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THERMAL MODELLING OF HIGH-FREQUENCY MAGNETIC COMPONENTS FOR POWER ELECTRONICS BY FINITE ELEMENT ANALYSIS

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*Thermal modelling of high-frequency magnetic
components for power electronics by Finite
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*"It's the questions we can't answer that teach us the most. They teach us how to think.
If you give a man an answer, all he gains is a little fact. But give him a question and
he'll look for his own answers."*

—Kvothe

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Abstract

Magnetic components design is vital to obtain low volume, high efficiency, inexpensive power electronics converters. Because of the interdependence between the electromagnetic and thermal fields, accurate models of both domains are crucial to properly optimize magnetic components. However, thermal modelling is often overlooked because classical models for magnetic components have limited application and accuracy.

The focus of this Ph.D. dissertation is developing generalized and accurate thermal models of magnetic components using Finite Element Analysis (FEA). FEA tools are commonly limited to validate final designs due to time consumption, but here a new methodology is proposed to overcome this limitation and the use of these tools to generate accurate thermal models for magnetic components is explained.

First, a two-step homogenization is developed to simplify 3D FEA thermal simulations while keeping proper accuracy. This simplification reduces the simulation time and ensure the convergence of the simulations, even for complex structures or litz bundles. Guidelines to apply this homogenization to inductors and transformers are described. A deviation lower than 5°C for most cases and an 8.1-fold reduction of the simulation time are achieved. The results are experimentally validated for DC and AC excitation.

Secondly, a model-extraction methodology from FEA thermal simulation results that predicts the maximum and surface temperatures of each region of the magnetic component for any operating conditions is proposed. This is specially useful to design these devices. The accuracy of this FEA-based thermal model is validated by 3D FEA simulations and experimental results, showing an absolute error lower than 5°C and relative error between -6.4% and 3.9% .

Finally, an optimization of magnetic components based on the simplified 3D simulations and the model extraction is performed. It is demonstrated that the new FEA-based optimization achieves the smallest volume possible and ensures a safe operation below thermal limits, unlike the traditional analytical-based optimizations. A 200 W classic Forward transformer is used as case study to validate the optimization, for which the analytical-based solution is 10°C over the thermal limits, while the FEA-based solution complies with this thermal constraint.

In conclusion, contrary to all previous approaches, combining the models developed in this thesis leads to maximum power density magnetic component designs while complying with the thermal specifications.

Resumen

El diseño de componentes magnéticos es vital para obtener convertidores de electrónica de potencia con volumen y coste reducidos a la vez que una elevada eficiencia. Debido a la interdependencia entre los campos electromagnético y térmico, se requieren modelos precisos para ambos dominios con el fin de obtener dispositivos correctamente optimizados. Sin embargo, la importancia del modelado térmico se subestima a menudo, recurriendo a algunos modelos para componentes magnéticos tradicionalmente utilizados, pero cuyo alcance y precisión son limitados.

El objetivo principal de esta tesis doctoral se centra en el desarrollo de modelos térmicos generalizados y precisos para componentes magnéticos por medio de Análisis de Elementos Finitos (AEF). El uso de herramientas de AEF se limita comúnmente a validar diseños finales debido al consumo de tiempo computacional que requieren. Se propone una nueva metodología para evitar esta limitación, así como el uso de estas herramientas para generar modelos térmicos precisos.

En primer lugar, se desarrolla una homogeneización de dos pasos para simplificar las simulaciones AEF 3D térmicas, manteniendo la precisión característica de estas herramientas. El objetivo principal de esta simplificación es reducir el tiempo de simulación y garantizar la convergencia de las simulaciones, incluso para geometrías complejas o devanados con hilo de litz. Se describen algunas pautas para aplicar esta homogeneización a bobinas y transformadores. Se ha conseguido una desviación menor de 5°C para la mayoría de casos y una reducción del tiempo de simulación de hasta 8.1 veces. Los resultados han sido validados experimentalmente para excitaciones de corriente continua y alterna.

En segundo lugar, se propone una metodología de extracción de modelos a partir de los resultados de simulaciones térmicas AEF, capaces de predecir la temperatura máxima y superficial en cada región del componente magnético para cualquier punto de operación. Esto es especialmente útil para diseñar estos dispositivos. La precisión de este modelo basado en AEF se ha validado por medio de simulaciones AEF 3D y resultados experimentales, obteniendo errores absolutos menores que 5°C y errores relativos entre -6.4% y 3.9% .

Finalmente, se propone una optimización de componentes magnéticos basada en el uso de las simulaciones 3D simplificadas y la extracción de los modelos de orden reducido. Se ha demostrado que esta optimización basada en AEF logra el menor volumen posible y garantiza una operación segura por debajo de los límites térmicos, a diferencia de las

optimizaciones tradicionales basadas en ecuaciones analíticas. Se usa un transformador Forward clásico de 200 W como caso de estudio para validar la optimización. La solución basada en métodos analíticos excede el límite térmico en 10°C , mientras que la solución basada en el uso de AEF cumple con esta restricción.

En conclusión, al contrario de otras metodologías de la literatura, la combinación de los modelos desarrollados en esta tesis conduce a diseños de magnéticos con la máxima densidad de potencia posible, asegurando que cumple con las especificaciones térmicas marcadas.

Nomenclature

a, b, c, d	Coefficients for the proposed multivariable fitting.
A_{cond}	Area of conductor.
A_e	Effective area.
A_p	Area product.
A_{ins}	Area of insulation.
AWG	American Wire Gauge.
h	Heat transfer coefficient or film coefficient.
k	Thermal conductivity.
k_a	Thermal conductivity of surrounding material (air, potting).
k_{bund}	Thermal conductivity of a litz bundle.
k_{cond}	Thermal conductivity of conductor material.
k_{eq}	Effective thermal conductivity.
$k_{eq_{ort}}$	Effective thermal conductivity in the orthogonal direction.
k_{eq_r}	Effective thermal conductivity in the radial direction.
$k_{eq_{\theta}}$	Effective thermal conductivity in the angular direction.
$k_{eq_{tan}}$	Effective thermal conductivity in the tangential direction.
$k_{eq_{TOT}}$	Effective thermal conductivity of several homogenized blocks.
k_{eq_z}	Effective thermal conductivity in the vertical direction.
k_{fe}	Thermal conductivity of ferrite.
k_i	Thermal conductivity of wire insulation.
k'_i	Homogenized thermal conductivity of different insulation.
$k_{i_{bund}}$	Thermal conductivity of bundle impregnation.
$k_{i_{imp}}$	Thermal conductivity of wire impregnation.

k_{ist}	Thermal conductivity of strand insulation.
k_{ins}	Thermal conductivity of insulation layers.
k_{ratio}	Thermal conductivity ratio (k_w/k_a).
k_w	Thermal conductivity of a homogenized wire.
l_{cond}	Width of conductors.
l_{ins}	Width of insulation layers.
n_{cond}	Number of conductor layers.
n_{ins}	Number of insulation layers.
N_{layers}	Total number of layers.
PF	Packing factor of a litz bundle.
P_{pri}	Losses in primary winding.
P_{sec}	Losses in secondary winding.
q''	Heat flux.
q''_x	Heat flux in the x direction.
R_{th}	Thermal resistance.
$[R_{th_{max}}]$	Coefficients matrix for maximum temperatures.
$[R_{th_s}]$	Coefficients matrix for surface temperatures.
r_1	Inner radius of a cylindrical foil winding.
r_n	Radius of the n foil layer.
r_{crit}	Critical radius.
τ	Volume fraction of conductor material.
τ^*	Introduced variable to replace τ .
$T_{c_{max}}$	Maximum core temperature.
$T_{c_{surf}}$	Average core surface temperature.
$T_{w_{max}}$	Maximum winding temperature.
$T_{w_{surf}}$	Average winding surface temperature.
ΔT	Temperature rise.
ΔT_{exp}	Measured temperature rise.
$\Delta T_{wireFEA}$	Temperature rise with 'wire level' homogenization.
ΔT_{sim}	Temperature rise obtained by simulation.
$\Delta T_{windFEA}$	Temperature rise with 'winding level' homogenization.
ϕ	Wire diameter.
v_{cond}	Volume of conductor material.
v_{ist}	Volume of strand insulation material.
$v_{i_{imp}}$	Volume of impregnation material.
$v_{i_{bund}}$	Volume of bundle insulation.
v_w	Volume of a wire.
w	Width of the cell under study.
W_a	Window area.

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Introduction

Motivation

Power electronics is an essential engineering field since every electrical system requires power processing units such as DC/DC converters, DC/AC inverters or AC/DC rectifiers. There is a tendency to increase the efficiency and power density of power systems while reducing their cost. This happens from consumer electronics to electric traction, passing through electric vehicles, charger stations and many other applications. The interest in increasing the efficiency and reducing the energy consumption of power systems is in part fuelled by the need to reduce global CO_2 emissions, 25% of which are generated in Europe [1]. Some initiatives on this line are the More Electrical Aircraft (MEA) philosophy and the *Clean Sky 2* European program [2] for the aerospace sector. On the other hand, the space sector is also interested in volume and mass reduction of power processing units because it reduces launch costs [3], [4].

The magnetic components of a power processing unit, such as inductors or transformers, account for most of the total volume, weight and cost, as shown in [5] and [6]. Thus, optimizing these components is vital. The optimization of magnetic components is challenging because of the interdependence between the electromagnetic and the thermal fields. As a consequence, it is crucial to use accurate models for both domains to get properly optimized magnetic components. Here is where thermal modelling proves its value. It is useful not only for obvious reasons, like avoiding the destruction of these components due to thermal stress or meeting the specifications imposed by the industry, but also to push magnetics to its limits by reaching the maximum power-density possible.

Thermal modelling of magnetic component for power electronics might represent a challenge for several reasons:

- **Heterogeneity.** A magnetic component comprises several parts made of different materials with dissimilar physical properties. Some of them are active, like the magnetic core or the winding, which in turn can be of different types (wire wound, litz bundles, foil, planar). At the same time, if the magnetic has different windings, they might have different electrical excitation. Others are passive, like the coil former, the connectors to the PCB or the clamps. Each part behaves very differently from an electromagnetic and a thermal point of view, resulting in a considerably high complexity to be analyzed.
-

- Number of conductors. Depending on the application, the winding can be made from just a few turns with a few parallel conductors, to hundreds of turns with many parallels. This fact makes the system very granular also increasing the complexity of the system to be modelled.
- Disparity of size between core and winding. This aspect forces a very demanding mesh when simulating magnetic components by means of Finite Element Analysis (FEA) simulations.
- Three-dimensional temperature distribution. Due to the geometry of magnetic components and their arrangement over a PCB layout and the possible use of additional cooling systems, the heat flux is rarely one-dimensional or even two-dimensional. Rather, heat flux usually has a three-dimensional (3D). Due to this fact and the previous statements, the resolution of the heat equation for 3D in such a heterogeneous component is too complex to be directly formulated analytically. Subsequently, the interaction between each region of the magnetics is difficult to be modelled or analyzed using analytical equations, thus making FEA tools a good candidate to analyze the thermal modelling of magnetic components.

The main focus of this thesis is to develop thermal models that overcome these difficulties while an accurate temperature estimation of the magnetic components. By means of these models, it is possible to achieve designs with higher power density than traditional designs. As an example, the developed models in this thesis will be used to optimize a transformer and compare its performance with other design obtained by a classical optimization.

Objectives

Some thermal models for magnetic components can be found in the literature, described in the next chapter. Nevertheless, their accuracy is limited or they are applicable to specific cases, in general.

Taking advantage of the accuracy of FEA tools and their capability to analyze complex systems, improved FEA-based thermal models are proposed in this thesis dissertation. The main objectives and contributions of this work are summarized next:

- Simplifying 3D Finite Element Analysis thermal simulations and ensuring their convergence. In this way, an accurate temperature prediction for any magnetic component can be achieved with much faster simulations, regardless of the

complexity of the geometry. A generic thermal homogenization for any type of winding based on FEA results is proposed.

- Generating an analytic thermal model from FEA simulations. To avoid the loss of accuracy due to mathematical simplifications, the potential of FEA simulations can be used to create accurate models for any operating conditions. However, FEA simulations provide a temperature estimation for each finite-element. This means that the matrix of results can be of the size of several million elements, which is not feasible to be handled to perform mathematical calculations. In this thesis, a method to obtain a simple matrix model from FEA results is presented, combining the accuracy and versatility of FEA tools with the simplicity of a linear thermal impedance matrix.
- Integrating the previous models within an optimization process. Taking advantage of the proposed fast and simplified 3D FEA simulations, combined with the method to extract a thermal model from them, it is possible to get better optimized magnetic components, compared to other optimizations, while ensuring a safe thermal operation below the established limits.

Methodology

In the last decades, Finite Element Analysis (FEA) tools have been used to get accurate electromagnetic or thermal estimations. They have been traditionally used at the last stage of the development of a product to check its performance, prior to start with physical prototyping. In this way, if any failure on the operation is observed by means of this FEA tools, it will be identified without incurring the cost of building a prototype that would show the same failure. Nevertheless, this is an iterative and time-consuming process.

Along this thesis, a different point of view is adopted. FEA tools are used not only as a virtual prototyping tool, but to build simplified and accurate thermal models of high-frequency magnetics for power electronics. The aim of these models is to reduce the design time while ensuring accurate temperature estimations. If these models are used within an optimization process, the volume of the magnetics can be reduced as much as possible, pushing the power density of these components to the limits.

Structure of the dissertation

After this introductory chapter, this dissertation is structured in the following sections:

The state of the art on thermal modelling for power electronics is reviewed in **Chapter 1**. Basic topics as the impact of the temperature dependence of the electrical performance, the physical properties, and the classification of the thermal models are explained in this chapter. The lack of FEA-based thermal models in the literature is identified.

In **Chapter 2**, a new winding thermal homogenization is introduced, to increase the speed of 3D FEA simulations while keeping good accuracy and ensuring their convergence. This proposed technique consists of replacing every conductor wire from a magnetic component by a single block. The properties of this block are calculated such that the equivalent system behaves similarly to the original one, from a macroscopic point of view. The main advantage of this transformation is the simplification of the thermal analysis without the penalty of losing accuracy. This is combined with the advantage that the convergence of the simulation of complex 3D geometries is guaranteed. Applications of this homogenization are discussed.

Chapter 3 introduces a methodology to create thermal models from FEA simulations. In this way, a simple coefficient matrix can be generated to accurately predict the temperature of a magnetic component for any operating conditions. This methodology combines the versatility of Finite Element Analysis tools to accurately reproduce any complex geometry for any boundary conditions. These FEA-based models can be implemented in optimization processes and circuit simulators, what is discussed in this chapter and Appendix B.

A FEA-based optimization process is proposed in **Chapter 4** to exemplify the usefulness of the developed models in optimizing magnetic components for power converters. This approach uses the fast and accurate 3D thermal simulations developed in Chapter 2 to analyze some designs. The model extraction from FEA results proposed in Chapter 3 is used to characterize those simulated designs. In order to avoid the extensive use of FEA simulations to characterize intermediate designs, a set of scaling laws is proposed in this chapter. The results of this FEA-based optimization are compared with a classic analytical approach. A 200 W Forward transformer is set as a case study. While the classical method leads to a design that exceeds the thermal limits, the proposed FEA-based optimization is able to achieve the smallest design that ensures a safe operation.

Finally, the main conclusions and contributions of this thesis are described in **Chapter 5**. To conclude, a brief overview of the possible future research lines is discussed.

State of the art on thermal modeling of magnetic components

The power electronics industry is constantly gaining importance worldwide, from low power consumption systems like space applications, wearable electronics or smartphones, to high power systems like electric traction applications or charging stations for electric vehicles. One of the main goals in those sectors is to reduce volume and to increase power density while getting a cost reduction. As a consequence, thermal issues are one of the main concerns of power electronics engineers. Magnetic components, such as inductors and transformers, are usually a key part in this kind of systems and represent a big percentage of their total volume. It is well known that the electromagnetic and thermal behaviors are interdependent. This yields to the need of developing proper models of both domains to get magnetic devices with higher power density and lower size while ensuring their operation under safe thermal limits. Electro-thermal coupled models are a useful tool to analyze and design magnetic components. Several models can be found in literature. An analytical approach for MF transformers was proposed in [7], where some 1D simplifications were performed to consider the convection and radiation boundary conditions. On the other hand, in [8] a dynamic model with thermal coupling is built from experimental results. A similar approach proposes an inverse model in which the correction of some analytical equations is made by means of prototype results [9]. The main disadvantage of these approaches is the need of hardware experiments. Virtual prototyping techniques are possible with Finite Element Analysis (FEA) tools. Some guidelines are proposed in [10] to develop a complete coupled model with FEA for semiconductors. The same strategy can be used for any other device, since it is proposed at mesh level. As a counterpart, techniques such as model reduction are required to handle the complete system [11]. To avoid this, a higher level thermal model based on 2D and 3D FEA is proposed in [12], which can be combined with [13] and [14] to get a complete electro-thermal model.

In this chapter, a general classification of the thermal models and a review of the most relevant models from the literature are described.

1.1 Temperature dependence of the performance of magnetic components

It is important to keep in mind the impact of temperature on the physical properties of magnetic components. The temperature dependence of the electromagnetic and the thermal properties of most common materials used for magnetic components is described next.

1.1.1 Temperature dependence of the electromagnetic properties

Most electromagnetic properties are dependent on the temperature and must be considered when designing a magnetic component.

This dependence with temperature can be clearly seen in the B-H hysteresis curves from figure 1.1 [15], for certain TDK soft ferrite. This effect is crucial, since the saturation flux density, B_{sat} , becomes lower as the temperature increases. As a worst case, the B_{sat} should be analyzed at the maximum operating temperature.

Furthermore, in the same figure, it can be seen the strong dependence of the permeability, μ , with temperature. Also, the *Curie temperature* can be identified (at 240°C in this case), at which the permeability starts falling drastically, thus losing the magnetic behaviour. To consider this dependence, some ferrite manufacturers suggest using a *temperature coefficient*, which is also dependent on the gap of the ferrite core [16].

Figure 1.1: Dependence of B-H curve (left) and permeability μ (right) with temperature of PC47 soft ferrite

Temperature also has a big influence on the core losses, as depicted in Figure 1.2 [15]. On the left, core losses over frequency curves are represented for different frequencies and temperature. On the left, for a particular pair of frequency and flux density, core losses are plotted over temperature. It can be seen that this soft ferrite material has an optimum temperature point, located at 100°C, at which losses are minimized. Ferrite manufacturers usually provide the coefficients of a quadratic approximation to model this dependence of

core losses with temperature [16]. However, this quadratic approximation has limited accuracy; a more accurate formulation is provided in Appendix C, as a minor contribution of this thesis.

Figure 1.2: Dependence of core losses, P_v , with temperature of PC47 soft ferrite: for different frequencies, f , and flux density, B , (left) and for a single B-f pair (right)

Regarding the electrical resistivity of conductors, it is well known that it increases as the temperature rises [17], thus increasing the electrical resistance too. This effect can be accounted with a simple temperature coefficient. In a similar manner, the high-frequency effects in conductors (skin, proximity and gap-proximity) are aggravated with the increase of temperature, as studied in [18] and [19]. As a consequence, higher winding losses are produced for higher temperatures.

1.1.2 Temperature dependence on the thermal properties

There is also a temperature dependence on the thermal conductivity [17]. It is mainly due to the motion of free electrons and lattice vibrations. For metallic materials, the strongest influence is due to the motion of free electrons. The molecular vibrations increase as temperature does, also increasing the free path of molecules. As a consequence, the flow of free electrons is reduced, thus reducing the thermal conductivity too. With a lower conductivity, the temperature rise is higher for the same value of internal heat losses generation.

In general, the thermal conductivity of the materials of interest in this study can be assumed constant for the usual range of operation (20°C to 100°C) [20], [21].

1.2 Thermal models classification

It is important to make a distinction between different thermal models to determine their range of application. The first criteria to classify them is the spatial direction on which

heat flux is propagated. Thermal models can also be either static or dynamic, depending on whether the time dimension is included in the model or not.

1.2.1 Heat flux spatial direction: 1D, 2D or 3D

Depending on how many spatial dimensions the heat flux is propagated, the thermal behaviour can be classified as one-dimensional, two-dimensional or three-dimensional, as depicted in Figure 1.3 [22]. In addition, those dimensions can be represented on cartesian, cylindrical or espherical coordinates. For example, the 3D temperature field in cartesian coordinates is a function of the three spatial dimensions: $T = f(x, y, z)$. Further details can be found in [20], [21].

Figure 1.3: Heat flux spatial dimension: 1D, 2D, 3D, from left to right

Keeping in mind this criterion is especially important to determine the applicability of one model to the actual heat flux. For example, in the magnetic component shown on the right side from Figure 1.3, it can be seen that the heat flux is different in any direction (x,y,z) and no symmetry is observed. As a consequence, a one-dimensional model like the one on the left would not be the most suitable approach to model the actual temperature distribution of that magnetic component. This is the most common situation for magnetic components and the main cause of the limitation of one-dimensional models.

1.2.2 Time domain: static or dynamic

Thermal models can also be classified as stationary (static) or transient (dynamic), depending whether the evolution over time is included in the model or not, respectively. For example, a dynamic 3D temperature distribution is a function of: $T = f(t, x, y, z)$. On the other hand, static models are used to express the temperature distribution at certain time instant, usually at steady-state conditions. They are commonly used to determine the temperature when a thermal equilibrium is reached. Further details can be found in [20], [21].

1.3 Thermal modelling methods

Several methods can be used to create a thermal model for magnetic components. All the models generated through these methods can be classified with the previously described criteria of the dimension and the time dependence. A general overview of the most representative thermal methods for magnetics found in the literature is described next.

1.3.1 Analytic equations

The solution of the heat equation for one-dimension can be easily obtained, as described in [20], [21]. If a higher dimension is required, the direct analytical solution of the heat transfer equation is only feasible for simple geometries, made by a single material and with simple boundary conditions non-dependent on the coordinates nor on the temperature. Some examples can be found in [23]. Nevertheless, the thermal distribution on magnetic components is generally three-dimensional. Furthermore, magnetics are made by different parts, with different materials and shapes. Hence, the direct solution of the heat equation for magnetics is normally unfeasible. As a consequence, several simplifications or Finite Element Analysis tools are required.

An attempt to simplify the heat equation for magnetics as a simple semi-empirical expression was made in [24] and reformulated in [25]. However, the accuracy is limited to the core family (EE cores) that were measured, under the same natural convection boundary conditions. In any other case, the temperature estimation by this expression would lead to results far from reality.

1.3.2 Thermal networks

Thermal networks are another approach to simplify the heat equation. They can be formulated as 1D, 2D and 3D. The approach from [26] proposes a one-dimensional thermal network, which is adjusted by means of experimental results. In [7] and [27], a three-dimensional network is proposed for medium-frequency solid-state transformers. A frequency-dependent thermal resistance to include high-frequency effects in the core was introduced in [28] as part of a three-dimensional network, leading to an accurate estimation of the core temperature. In every case, thermal networks have the limitation of being linear (while convection and radiation boundary conditions are non-linear) and only applicable to a particular case, since every magnetic component would be represented by a different network. Additionally, power losses are usually considered as concentrated in a single point, but not distributed along the solid bodies. This leads to a lower accuracy. On the other hand, the higher the granularity or number of nodes, the higher the accuracy.

Nevertheless, there is a point where it is much simpler to perform a FEA simulation than handling a very complex thermal network.

1.3.3 FEA-based models

Some examples of the direct solution of the heat transfer equation by means of Finite Element Analysis tools can be found [29]. They are a particular solution for the equation, not models by themselves. There is a lack of models which are extracted from the results of Finite Element Analysis simulations. Only some examples for the electro-magnetic behaviour [13] and a general idea for the thermal domain [12] can be found in the literature. This is because FEA tools are commonly used to obtain accurate results of a particular design, prior to build a prototype. However, they are not used to develop thermal models.

In this thesis, the main idea is taking advantage of the potential of FEA tools to create generalized, simple and accurate thermal models for magnetic components.

1.4 Thermo-electrical coupling

Every thermal model has in common the inputs and the outputs, regardless of their classification. Power losses and the boundary conditions are the input for thermal models. Specifically, the power losses are a result of the electro-magnetic operation. As a result, thermal models provide the temperature as output.

It is important to highlight that both the electromagnetic behaviour (thus the power losses too) are dependent on the temperature, as well as the thermal behaviour. So, there are two different options to couple the electromagnetic and the thermal behaviour, depending on whether this feedback with temperature is included in the model or not, as explained next.

1.4.1 One-way coupling

The resulting power losses obtained by the electromagnetic model or simulation are directly fed into the thermal model in a one-way coupling approach, as shown in Figure 1.4. Hence, the temperature dependence of the electromagnetic model is neglected.

Figure 1.4: One-way EM-thermal coupling.

If this approach is used, the ‘worst case’ temperature of the physical properties of the magnetics must be considered to get accurate results.

1.4.2 Two-way coupling

In the case of a two-way coupling approach, several iterations of both the EM and the thermal models or simulations are performed, as depicted in Figure 1.5. In this way, the temperature dependence of power losses is characterized.

Figure 1.5: Two-way EM-thermal coupling.

By means of this two-way coupling, an additional step can be carried out. If the dependence of the properties with temperature is implemented, they can be included in the loop in such a way that they are recalculated in each iteration, prior to update the power losses or the temperature.

Some examples of the electromagnetic-thermal coupling for magnetic components can be found in [8], [30].

1.5 Chapter summary

In this chapter, the thermal dependence of the material properties of magnetic components is reviewed to highlight the importance of thermal modelling of magnetic components. A general classification of the thermal models and the concept of thermo-electrical coupling are described.

An overview of the most representative models for magnetic components found in the literature is described. A more specific review of the literature will be explained in each chapter. However, the general overview of the literature from this chapter reveals a lack of thermal models based on Finite Element Analysis, as summarized in the next table:

Table 1.1: General overview of the thermal models for magnetic components from the literature.

	Solution of the heat equation	Analytical simplifications	Thermal networks	FEA-based models
Static	✓	✓	✓	X
Dynamic	✓	✓	✓	X

The main aim of this thesis dissertation is taking advantage of the accuracy and versatility of Finite Element Analysis tools to create generalized, simple and accurate thermal models for high-frequency magnetics for power electronics.

An improved method for 3D FEA thermal simulations

Homogenization techniques have been extensively used in every engineering field during history due to their consequent simplification of the original problem. Independently from the area of study, those techniques are based on the same idea: replacing an heterogeneous system, made by a mixture of substances or materials, by a single equivalent object with uniform physical properties. The computation of these equivalent properties is done in such a way that the new system behaves in a similar manner than the original heterogeneous system, from a macroscopic point of view. As a result, the equivalent system can be handled more easily. Some systems like alloys made by several elements with different diameters and weights, composite materials commonly used in aeronautics, chemical solutions with different solute concentrations or the litz bundles from a winding in a transformer or an electrical machine, are clear examples of their natural complexity due to their heterogeneity, not only regarding the different dimensions of the substances that composed them, but their dissimilar physical properties too. Analyzing them might be

Figure 2.1: Example of the proposed two-step homogenization. 3-D thermal distribution of a PQ core inductor: detailed FEA simulation (left); homogenized winding (right).

unfeasible due to excessively demanding computational cost, but the use of homogenization techniques can avoid this limitation.

In this chapter, a homogenization method to analyze the thermal behavior of the winding of any magnetic component for power electronics is proposed, with the main aim of speeding-up 3D FEA simulations and ensuring their convergence. The general idea is replacing every wire from the winding by an equivalent single block: its maximum temperature by applying the power losses is the same as the replaced winding, as shown in Figure 2.1. This process consists of a two-step homogenization based on a multi-variable curve fitting of FEA simulation results. The general features of the proposed method are next:

- Applicable both to inductors and transformers, with any winding type (solid wire, litz bundle round or squared, foil) and core geometry. It is also valid for both DC and AC current excitation.
- Its application to 3D FEA thermal simulations leads to considerably reducing the required computational time while ensuring accurate results.
- Its application to thermal networks leads to more accurate results than using traditional approaches.

The structure of this chapter is as follows. First, the scope of this method is explored. Then, the construction of the proposed method for every common winding types (round wires, litz bundles, foil) is detailed, also including a comparison with traditional approaches. A guide to use this homogenization for inductors and transformers is explained throughout the next section. The validation of the proposed method and its application to 3D FEA thermal simulations is carried out, by means of DC and AC tests. The key point regarding the achievable computational time reduction by using this method is also analyzed in the same section. Finally, how to apply it to simplify thermal networks and the consequent accuracy increase are explained.

2.1 Winding thermal homogenization

Given the interdependence between the electromagnetic and thermal domains, the use of a thermal model to calculate the temperature rise of a magnetic component due to its generated internal power losses is mandatory. As reviewed in Chapter 1, facing the actual heat dissipation of a magnetic component by solving the three-dimensional heat equation with internal heat loss is unfeasible. Some empirical approaches or simple equations based

on one-dimensional simplifications, even proven being useful as first approximations, their use is quite limited to certain geometries or operating conditions.

The reality is that there are infinite combinations of core shapes, winding strategies and operating conditions. Even with the same core shape, magnetics might have quite different windings between each other. The winding type, as well as its size, strongly depends on the application. Furthermore, heat transfer is rarely one-dimensional; it might be approximated at microscopic level, but not for the size of these devices. Both convection and radiation may be stronger at one side of the component than other due to its location within an enclosure. There might be other electronic devices with internal losses arranged close to the magnetic component, in the PCB layout. All these conditions are quite common in industrial power electronic converters and lead to three-dimensional heat flux.

3D Finite Element Analysis simulations or thermal networks are two commonly appealing resources to deal with this complex thermal modelling. Nevertheless, they still have some downsides. Regarding the first, they are considered as the most accurate tool to solve physics problems. However, the difference in mesh size of the core compared to the winding wires generates a very demanding mesh in terms of computational requirements. Other details like the size of the strands from a litz bundle might even cause convergence problems and make unfeasible the solution of the simulation. Regarding the thermal networks, the same heterogeneity of the magnetic component elements increases their granularity, so much that they also become unfeasible.

So, using some simplifications looks like the next logical step. Homogenization techniques are a useful approach to handle this problem. They have been used in the past for many other engineering fields [31], [32] to obtain the equivalent properties of heterogeneous systems, hence simplifying them. All these approaches share a similar strategy: expressing the equivalent properties as a function of the individual properties and the volume concentration of each material. There are some differences between them, like considering shape factors, the interaction between particles or a relationship with the surrounding boundary conditions.

As stated in the review from Pietrak *et al.* in [33], some of the previous approaches can be used to get the effective (equivalent) thermal conductivity which is the magnitude of interest of the present work. And more particularly, they are perfectly suitable to homogenize electrical litz bundles for electrical machines [34]–[38] and inductors or transformers for power converters [39], [40]. The effective thermal conductivity, k_w , calculated by these approaches will govern the heat equation, $\vec{q}'' = -k_w \vec{\nabla} T$, corresponding to the homogenized wires.

Nevertheless, even using them, a magnetic component might still be too complex to be analyzed due to the high number of wires. In this work, one step further is carried out. After simplifying each wire, every wire from the winding will be replaced by a single block, whose conductivity is equal to k_{eq} . This equivalent property will make it behave as the actual winding, in the sense that it will have the same temperature rise for the same power losses than the original magnetic component. This is shown in Figure 2.1. The mathematical derivation of this model will be explained and compared with other methods in the next sections.

Certain specific modifications make it valid for any magnetic component, made by any winding type, and for both DC and AC current excitation. This versatility is a strong difference between the models found in the literature and the proposed model. Further advantages of this method are the reduction of computational time and the decreased risk of convergence problems of 3D simulations, as well as the increase of accuracy of thermal networks.

2.2 Proposed two-step thermal homogenization

A magnetic component is a very heterogeneous system. For instance, transformers are made by different coupled windings, which might be made of different types, as the one from [41], whose primary winding is made by litz bundles and secondary winding by foil. Most conventional transformers have wound windings with different wire diameters in primary than in secondary. But transformers are not the only example. Even inductors, with a single winding, are still heterogeneous in the sense that there might be hundreds of wires, litz bundles or foil layers. All them have the corresponding insulation to isolate each conductor. This heterogeneity becomes even stronger for litz bundles, where each bundle is made by an external insulation that groups from tens to hundreds strands, embedded in an impregnation matrix. Additionally, the winding might also be potted to reduce its thermal resistance. And each of the mentioned materials (conductors, insulators, impregnation, potting, etc) have different geometrical characteristics and physical properties.

In order to face this heterogeneity and make possible the modelling of these systems, homogenization techniques are very useful. In this chapter, a two-step thermal homogenization is proposed. The modelling flow diagram is shown in Figure 2.2. The purpose of making two steps is getting a final model where the whole winding is replaced by a single block with homogeneous thermal conductivity, k_{eq} . The two steps are:

- **Wire level homogenization:** First, a wire and its surrounding air (or potting) are taken as a representative sample of the winding. The conductor (usually

Figure 2.2: Design flow to obtain the thermal equivalent winding. Example of a pot core with round solid and litz wire. From left to right: cross-sectional area of a pot core inductor; homogenization at wire level and at the subsequent winding level; inductor with the obtained equivalent layer.

copper or aluminium), with thermal conductivity k_{cond} , and its insulation (enamel, polyamide, polyvinyl formal, polyurethan among others [42]), with conductivity k_i , are homogenized as a single equivalent wire with conductivity k_w . In the case of a litz bundle, the insulation of each strand (k_{ist}), the impregnation material if any (k_{imp}), and the insulation of the bundle (k_{bund}) are included in this first homogenization.

- **Winding level homogenization:** Continuing from the homogenized wire (k_w) is surrounded by air or potting with conductivity k_a . The width of the representative sample is denoted as w , and the diameter of that wire as ϕ . Considering the conductivity ratio (k_w/k_a) and the size ratio (ϕ/w), it is possible to obtain the **equivalent thermal conductivity of the winding**, k_{eq} , as a function of the previous ratios. This relationship will be obtained by a multi-variable curve fitting of FEA results, what guarantees a considerably high accuracy.

Once obtained the homogeneous thermal conductivity of the winding, the equivalent system is only composed of a ferrite core, the homogenized winding and the surrounding air or potting material. The resulting geometry is simplified with respect to the original system and the number of different materials with different sizes is reduced, so that it can be handled more easily. After this homogenization, local details like the individual temperature of the wire n are lost. But the most important feature of the equivalent system is that it will serve to estimate the maximum and surface temperatures of the magnetic component, which will correspond to the temperatures of the actual component.

2.2.1 Wire level homogenization

The first step of the homogenization consists in obtaining the effective thermal conductivity of a wire, as a function of the size and conductivity of each material that composes it (conductor and any insulation). The requirements to apply this step to any conductor type (solid wires, round and squared litz bundles, and foil) is explained next.

Solid wire and insulation thickness analysis

Solid wires might seem simple to be modeled. However, the required mesh of a FEA simulation can be excessively restrictive due to the low thickness of its insulation, compared to the conductor diameter. A particular simplification for this case is possible: neglecting the thermal conductivity of the insulation, k_i , as depicted in Figure 2.3. In other words, the conductivity of the homogenized wire is approximated to the conductivity of the conductor:

$$k_w \approx k_{cond} \quad (2.1)$$

Figure 2.3: Wire level homogenization of solid wire.

This is not a homogenization in its own sense, but it is proven practical for the purpose of the proposed methodology. The accuracy of the subsequent temperature estimation if using this approach is not compromised due to the next reasons:

- *The thermal gradient at the cross-section of a wire is negligible*, due to the reduced dimensions of the wires, the high thermal conductivity of the conductor and its internal heat generation. This statement is valid for commercial wires smaller than AWG 10 (2.59mm) [17], rated for low and medium power applications (< 100 kW).
- *The radius of commercial wires is lower than the 'critical radius'*. The critical radius is defined as the value of radius from which the addition of insulation would increase the thermal resistance of a wire. Its value is calculated as:

$$r_{crit} = \frac{k_i}{h} \quad (2.2)$$

2.2. Proposed two-step thermal homogenization

Where k_i is the conductivity of the insulator and h is the film coefficient. By adding insulation to a round wire, the conduction resistance increases, but the convection resistance decreases as the outer surface area is incremented. Since the total resistance is the sum of both, it can be minimized for an insulation with thickness lower than $r_{crit} - r_i$, as shown in Figure 2.4. Above that value, the total resistance starts rising again.

Figure 2.4: Thermal resistance of an insulated wire over the difference of the total radius minus the inner radius, $r - r_i$. The total resistance, R'_{tot} , is the sum of the conduction resistance, R'_{cond} , plus the convection resistance, R'_{conv} . The critical radius, r_{crit} , corresponds to the value of that difference at which R'_{tot} is minimum.

The most restrictive case regarding the critical radius is the lowest value. By adding insulation over that limit, the thermal resistance of the wire is increased. Then, the lower the insulation conductivity k_i and the higher the film coefficient h , the lower the critical radius. Typical values of k_i and h range $[0.1, 0.4] \frac{W}{m \cdot K}$ and $[1, 100] \frac{W}{m^2 \cdot K}$, respectively. So, the worst combination is $k_i = 0.1 \frac{W}{m \cdot K}$ and $h = 100 \frac{W}{m^2 \cdot K}$, making $r_{crit} = \frac{0.1}{100} = 0.001m = 1mm$. This value is higher than AWG18 ($diam = 0.83mm$). That means that the total resistance of the wire can be decreased as long as the sum of the inner conductor radius plus the insulation thickness is lower than 1 mm. Hence, assuming that the thermal conductivity of the whole wire is equal to the conductivity of the wire without insulation is conservative, leading to a thermal resistance higher than the actual value. But still, the difference in the estimation of the temperature gradient within a conductor is insignificant (in the range of some decimals).

Let's use a commercial AWG20 wire (0.425 mm outer radius, 0.02mm insulation thickness) as an example to illustrate the effect of neglecting the insulation. Figure 2.5 shows the thermal gradient in the cross-section of the AWG20 wire when injecting Joule losses of 1W, having a film coefficient of $h = 5 \frac{W}{m^2 \cdot K}$. Two situations are compared:

a) *Wire with actual insulation:* In the first situation, the temperature in the copper area is uniform and equal to 94.922°C . Due to the insulation, a slight temperature gradient can be seen from the inner radius (94.922°C) and the outer radius (94.896°C). The temperature decreases as the radius increases because the heat exchange surface increases too and there is no heat generation with the insulation.

b) *Wire assuming homogeneous conductivity equal to k_{cond} for the whole cross-section:* In the second situation, the temperature across the whole wire is uniform and equal to 94.896°C . The reason for this fact is the high conductivity of the conductor ($400 \frac{\text{W}}{\text{m}\cdot\text{K}}$) and the small radius of the wire (0.425 mm). As a consequence, the thermal gradient from the center to the outer radius is imperceptible.

Figure 2.5: Thermal gradient in a copper conductor carrying current: with and without insulation.

Litz bundle

Litz bundles are a recommended option to reduce high-frequency effects [43], due to the low diameter of their strands. As a counterpart, the effective thermal conductivity of a bundle, k_w , is considerably lower than a round solid wire of the same diameter. It is hereby important to define the difference between the two different types of litz bundles, shown in Figure 2.6:

- *Non-impregnated bundles:* there is no impregnation material, so the strands are just attached together by the external bundle insulator [36]. The thermal resistance of this solution is lower compared to the next one, since the power losses in each strand

are dissipated just by conduction at the contact points and by radiation between each other.

- *Impregnated bundles*: the strands of the bundle are embedded in a potting matrix, to reduce the thermal resistance of the bundle by ensuring a heat transfer path by conduction through this additive material [37].

Figure 2.6: Wire level homogenization of litz wire: a) impregnated bundle, b) non-impregnated bundle; homogenized wire on the right.

In order to characterize the effective thermal conductivity of litz bundles, several approaches can be found in the literature and are reviewed in the following.

Hashin-Shtrikman's equation [44] was proposed in 1963 to study the elastic properties of two-phase composite materials. This equation puts the effective properties of a composite as a weighed volume relationship of its components' properties. This approach can be applied to get the effective thermal conductivity of a bundle as:

$$k_w = k_i \cdot \frac{(1 + \tau) \cdot k_{cond} + (1 - \tau) \cdot k_i}{(1 - \tau) \cdot k_{cond} + (1 + \tau) \cdot k_i} \quad (2.3)$$

Where k_{cond} and k_i are the conductor and insulation conductivity, respectively. τ is defined as the volume fraction of conductor material over the total volume:

$$\tau = \frac{vol_{cond}}{vol_{tot}} \quad (2.4)$$

Siesing *et al.* reveal in their study based in experimental measurements [36] that the previous Hashin-Shtrikman's equation is more suitable for impregnated bundles, whose impregnation, strand insulation and bundle insulation properties are so close that can be considered as equal (k_i).

Nevertheless, if the packing factor, PF , is low ($PF < 50\%$) or the bundle is non-impregnated, this assumption is no longer valid and might lead to unacceptable error in the estimation of k_w . For those cases, the impregnation and insulation materials cannot

be assumed as equal. A modification to consider these details was proposed by Simpson *et al.* in [37], where k'_i is used instead of k_i in Equation (2.3). It is defined as:

$$k'_i = \frac{k_{i_{st}} \cdot vol_{i_{st}} + k_{i_{imp}} \cdot vol_{i_{imp}} + k_{i_{bund}} \cdot vol_{i_{bund}}}{vol_{i_{st}} + vol_{i_{imp}} + vol_{i_{bund}}} \quad (2.5)$$

Subscripts i_{st} , i_{imp} and i_{bund} are referred to the strand insulation, impregnation and bundle insulator, respectively. So k'_i takes into account their conductivity and volume fractions over the total volume. These fractions can be calculated with the value of PF , usually provided by the manufacturer [45]. The analysis from [37] shows an error in the estimation of the effective conductivity of litz bundles lower than 11 % for low PF litz bundles ($PF < 50$ %) and lower than 37 % for high PF bundles ($PF > 50$ %).

Comparison with traditional approaches

Other approaches can be found in recent literature for high packing factor bundles. The thermo-electrostatic field analogy proposed by Jaritz *et al.* [39] shows a deviation with experimental measurements in the band of [1.3, 12.5] %. It shows good accuracy, but it might be complex. The methodologies from Wrobel *et al.* [38] and Kyaw *et al.* [40] are based in the use of thermal networks. Both approaches require calibration with measurements in order to avoid the tolerances from the data provided by manufacturers or the lack of that information. The second one makes a differentiation between square-packed and hexagonal-packed bundles in the approach. Their maximum error compared with experimental data is 34 % and 18 %, respectively.

At this point, it can be stated that the most suitable approach for this first step homogenization to get the effective k_w of low PF litz bundles is the modified Hashin-Shtrikman's equation from [37]. It uses a simple average of the different insulator properties, k'_i from Equation (2.5), which is used instead of k_i in Equation (2.3). For high PF , other approaches like [40] and [38] can be used at the cost of the additional complexity of calibration with measurements. Nevertheless, the modified Hashin-Shtrikman's equation still provides an acceptable deviation, close or within the uncertainty band obtained by the previous methodologies, what is combined with the additional advantage of its simplicity. The accuracy analysis by using this approach will be shown in Section 2.4.

The same conclusions can be stated for both round and squared section litz bundles. The difference between them affects the second step of the proposed homogenization, explained in the next Section 2.2.2.

Foil

Since foil windings are already made by layers of copper and insulation, there is no need to apply this first homogenization step. Further considerations will be regarded for this type of winding in the second step, at winding level.

2.2.2 Winding level homogenization

Following the schema depicted in Figure 2.2, the next step is performed at winding level. The properties of the homogeneous wires (k_w) are already established at this point but, still, they are surrounded by air or potting material (k_a). A further step can be carried out.

A geometric transformation is performed: replacing every homogeneous wire and the surrounding material into a single equivalent winding block, characterized by its effective thermal conductivity, k_{eq} . This property is uniform and constant across the whole equivalent winding. And its value is such that the maximum temperature rise, provoked by certain internal power losses, is equal to the winding of the original magnetic component given the same amount of losses.

The geometric transformation involved in this second step has certain decrease of the external heat exchange surface with the ambient as a consequence. Nevertheless, since the wires of a properly designed magnetic component are very tight between each other and due to their relatively low diameter, it can be assumed that this change of the external surface has a minimum effect on the global heat transfer.

The corresponding calculations and procedure to perform the winding level homogenization for different winding types is described next.

Round wires: solid wire and litz bundle

Figure 2.7: Winding level homogenization for round wires.

The procedure explained here is applicable for any round wire. In other words, after obtaining a wire with homogeneous k_w from the previous step, the situation is the same

regardless the original winding was made by solid wires or litz bundles. That wire and its surroundings (k_a) are homogenized in a block with k_{eq} , as shown in Figure 2.7.

Some analytical approaches are compiled by Chauvicourt *et al.* in [34], but the accuracy of these approaches is limited to low volume fraction of conductors within a surrounding matrix. This situation is very common in composite materials like alloys of carbon fiber, for which they were developed. However, this is not the case of the wires from the winding of a magnetic component. They are usually very tight and packed, leading to high filling ratios. Hence, these approaches are not recommended for this purpose. A more precise model to get k_{eq} for equivalent windings is developed and is **one of the main contributions of this work**.

From a physical point of view, the effective conductivity k_{eq} can be expressed as a function of the size and conductivity ratio of each material that composes the system to be simplified, what is in agreement with every analytical approach from [34]. Other variables, like shape factors, can be disregarded, as every wire is round in this case. First, a representative sample of the winding must be established. Since at this point the winding is made by evenly arranged homogeneous wires, a squared cross-sectional –from now on denoted as ‘cell’– of any of them can be considered. This cell of width w contains the homogeneous wire of diameter ϕ , whose homogeneous conductivity is k_w , and the surrounding air or potting material, with conductivity k_a . Then, the equivalent k_{eq} will be expressed as a function of those variables: (ϕ/w) as the size ratio; and $k_{ratio} = k_w/k_a$ as the conductivity ratio (where k_w is obtained according to Section 2.2.1). The main goal of defining these ratios is re-casting the problem in a *nondimensional* form. In this way, the resulting k_{eq} can be computed for any value of the implied variables. Furthermore, the number of involved quantities is reduced to the minimum, thus reducing the complexity of the system [23].

Now that the variables that influence k_{eq} are defined, the next step is obtaining the exact value of k_{eq} for each combination of those variables. To do so, a two-dimensional FEA simulation will be run for each combination. FEA simulations are used due to their accuracy, to avoid the limitation of the analytical approximations. The simulation setup, depicted in Figure 2.8, reproduces the sample to be analyzed. By following a similar process to obtain the effective thermal conductivity of a sample described in [20] and [21], the boundary conditions are set as follows:

- Since the sample cell is evenly ‘repeated’ along the window of a magnetic component, top and bottom surfaces are defined as adiabatic to account for this symmetry [20]. The problem could be defined in the perpendicular direction, but the result would be the same since a centered wire in a squared cell has a double symmetry.

- Certain heat flux, q_x'' [W/m²], is injected on the left surface of the cell, so that a temperature gradient appears between both left and right sides.
- The average temperature of the left and right surfaces is denoted as T_1 [K] and T_2 [K], respectively. The temperature rise between them is equal to $\Delta T = T_1 - T_2$.
- Temperature T_2 is imposed as a constant temperature.

Figure 2.8: 2D FEA thermal simulation setup to get k_{eq} of round wires at winding level.

Then, according to [20], the effective thermal conductivity of the cell is expressed in terms of the temperature rise between its surfaces, consequence of the applied heat flux, across its width:

$$k_{eq} = q_x'' \cdot \frac{w}{\Delta T} = q_x'' \cdot \frac{w}{T_1 - T_2} \quad (2.6)$$

As previously stated, k_{eq} is a function of the size ratio ϕ/w and the conductivity ratio $k_{ratio} = k_w/k_a$. Figures 2.9 and 2.10 illustrate this dependence. k_{eq} is normalized with k_a to generalize the conclusions.

In Figure 2.9, it can be seen that k_{eq} is equal to k_a for $\phi/w = 0$, since it is the same than having an infinitesimally small wire diameter compared to the width of the cell. As the contrary case, if $\phi/w = 1$, the wire diameter is equal to the width of the cell; then k_{eq}/k_a tends to certain value, close to k_w , since the wire occupies the most part of the cell.

In Figure 2.10, ϕ/w is fixed and the impact of the conductivity ratio k_w/k_a on k_{eq} is analyzed. For $k_w/k_a = 1$, k_{eq}/k_a is also equal to 1, since both the wire and the rest of the cell would have the same conductivity. On the contrary, if k_w/k_a tends to infinite, k_{eq}/k_a will tend to certain value close to k_w .

Then, in order to get the value of k_{eq} for a wide range of combination of ϕ/w and k_w/k_a , a 2D simulation is performed for each one. The diameter over the width of the cell, ϕ/w , is swept 31 steps from 0.5 and 0.8, and 100 more from 0.8 to 0.998. A lower

 Figure 2.9: Size dependence of k_{eq} for given k_w and k_a .

 Figure 2.10: Conductivity dependence of k_{eq} for given ϕ and w .

number of steps is chosen for the first range since the variations are minimum, compared to the last. Values below of $\phi/w < 0.5$ are skipped, since they are not a common case for windings in a magnetic component; the wires are not usually as separated from each other as that. Additionally, the analytical approaches from [34] are proven accurate enough for that low range of $\phi/w < 0.5$, more commonly used in structural composites. On the other hand, $k_{ratio} = k_w/k_a$ is swept with 600 logarithmic steps from 10^0 to 10^5 . For each combination of k_{ratio} and ϕ/w , the effective thermal conductivity is obtained by simulation and labeled as k_{eqsim} . These results are represented in Fig. 2.11. The effective conductivity is normalized with k_a in order to generalize the resulting conclusions.

Figure 2.11: Normalized k_{eq} with k_a for a sweep of k_{ratio} and ϕ/w , obtained by 2D FEA simulations.

Now, a multi-variable fitting of the exact values of k_{eq}/k_a is desired to create a mathematical expression to obtain those values without the need of using any simulator. Since Hashin-Shtrikman's method shows a relatively proper accuracy for low filling values of ϕ/w , a modification is proposed: instead of using $\tau = vol_x/vol_{tot}$, a new variable named τ^* is used. Accordingly, k_{eq}/k_a is expressed as:

$$\frac{k_{eq}}{k_a} = \frac{(1 + \tau^*) \cdot k_{ratio} + (1 - \tau^*)}{(1 - \tau^*) \cdot k_{ratio} + (1 + \tau^*)} \quad (2.7)$$

Where the value of τ^* is such that it makes the Equation (2.7) equal to the exact $k_{eq_{sim}}/k_a$ obtained by FEA simulations. To do so, τ^* is expressed as a sum of three exponential functions of (ϕ/w) :

$$\tau^* = \frac{\pi}{4} \cdot \left(\frac{\phi}{w}\right)^2 + a \cdot \left(\frac{\phi}{w}\right)^b + c \cdot \left(\frac{\phi}{w}\right)^d \quad (2.8)$$

Where the first exponential corresponds to the volume relationship ($\pi/4$) of a cylinder centered on a squared section of normalized width ($w = 1$) and the other terms are used to adjust τ^* to fit $k_{eq_{sim}}/k_a$. The coefficients a , b , c and d in turn are expressed in terms of k_{ratio} according to Equations (2.9):

$$a = \frac{0.15323 \cdot k_{ratio} - 0.21445}{k_{ratio} + 6.1801} \quad (2.9a)$$

$$b = \frac{14.297 \cdot k_{ratio} + 78.569}{k_{ratio} + 8.3734} \quad (2.9b)$$

$$c = \begin{cases} 0 & k_{ratio} < 6.7 \\ \frac{0.04902 \cdot k_{ratio} - 0.24267}{k_{ratio} + 19.568} & 6.7 \leq k_{ratio} \end{cases} \quad (2.9c)$$

$$d = \begin{cases} 0 & k_{ratio} < 6.7 \\ \frac{270.55 \cdot k_{ratio} - 472.41}{k_{ratio} + 3.6959} & 6.7 \leq k_{ratio} \end{cases} \quad (2.9d)$$

It should be noted that coefficients c and d are 0 for k_{ratio} lower than 6.7.

This procedure leads to a properly fit with $R - square > 0.9999$ of the exact effective thermal conductivity of a wire surrounded by air or potting. The accuracy of this approach will be analyzed next.

To conclude, the effective thermal conductivity of homogeneous wires surrounded by air or any potting material is calculated by means of Equations (2.7), (2.8) and (2.9).

Despite the complexity of the process to get this multi-variable fitting, the result is a set of equations simple to be handled by the designer of any magnetic component.

Comparison with traditional approaches

Instead of using the previously developed Equations (2.7), (2.8) and (2.9), another approaches could be used by applying some of the traditional methods compiled in [34]. Those approaches were analyzed in Section 2.2.1 to compute the effective conductivity of litz bundles, but has not been analyzed in the literature as a second-step homogenization. So, in the following, those approaches are compared with the proposed equations. The applicable approaches from [34] are: Hashin-Shtrikman's, Maxwell's, Rayleigh's and Milton's equations.

In order to carry out this comparison, every approach is compared with the exact values $k_{eq_{sim}}$ for a winding made by homogeneous wires. Those were obtained by means of Finite Element Analysis simulations, depicted in Figure 2.11. For this purpose, the relative error between the estimated conductivity ($k_{eq_{estim}}$) and the exact value from FEA simulations ($k_{eq_{sim}}$) is defined as:

$$Error_{rel}[\%] = (k_{eq_{estim}}/k_{eq_{sim}} - 1) \cdot 100 \quad (2.10)$$

The results are shown in Figure 2.12. It can be seen that Hashin-Shtrikman and Rayleigh approaches are suitable for size ratios in the range of $\phi/w < 0.8$. Otherwise, they would incur in significant underestimation of k_{eq} as high as 83% and 88%. On the other hand, the consequent error of using Maxwell's approach leads to an underestimation between 15% and 95% for the whole range $0.5 < \phi/w < 1$; while Milton's approach varies

Figure 2.12: Comparison of different approaches to get the homogenized k_{eq} .

between +18% and −75%, leading either to over or under-estimation respectively. This high deviation is undesirable, since it would lead to an ulterior error in the temperature estimation.

Nevertheless, the performance of the proposed approach in this work leads to a small deviation in the band of $[-2.4, 4.6]\%$. This can be seen in Figure 2.12, where the error is depicted with the same scale than for the previous approaches and with a more detailed scale. It is specially important getting a relatively low error in the estimation of k_{eq} at high size ratios ($\phi/w \approx 1$), since it is the most common situation for magnetic components. For this reason, it has been justified that the proposed approach is the most suitable to perform the winding-level homogenization of round wires (solid wire or litz bundles), compared to other traditional approaches.

Square litz bundle

Squared-section litz bundles are very useful to get high window filling ratios, since they lead to an almost perfect packing with negligible space between bundles. As a consequence, a second homogenization is not required for this winding type.

The procedure to obtain the effective thermal conductivity of a winding made by square litz bundles is exactly the same than the first step homogenization for round litz wires (Section 2.2.1). Then, Equation (2.3) combined with Equations (2.4) and (2.5) will be used to compute $k_{eq} = k_w$ in this case, as depicted in Figure 2.13.

Figure 2.13: Homogenization of square litz bundle winding.

Foil

Foil windings show very good performance at high-frequencies due to their low thickness, as explained in [43]. Additionally, their thermal impedance is lower than litz bundles or solid wires, since the heat exchange surfaces are much larger in comparison. Those features make this winding type very suitable for some high-frequency applications.

Foil barrel-wound magnetics achieve high window filling ratios [41], as square litz bundles do. Then, the procedure to get the effective thermal conductivity of foil winding is relatively easy. It is important to highlight that foil winding is characterized by its anisotropic thermal conductivity, since each copper layer is insulated. In this case, one-dimensional heat transfer equations are used to get k_{eq} for each spatial direction. Each one will be solved in a similar manner to a composite wall [20] (see Appendix A). Due to geometric differences, a distinction between rectangular center leg (XY) and round center leg (RZ) magnetics is made.

- XY rectangular winding:

Foil windings are usually arranged so that conductor layers and their corresponding insulation are evenly alternated, as shown in Figure 2.14, where the layers 1 and 3 (red)

Figure 2.14: Homogenization of foil winding for rectangular center leg cores.

represent conductor layers, 2 and 4 (yellow) are insulation layers. For rectangular center leg cores, a simple thermal network using cartesian coordinates can be analyzed to compute the effective thermal conductivity for each spatial direction:

- $k_{eq_{ort}}$: Orthogonal direction across the perpendicular direction to the winding.
- k_{eq_z} : Top-down direction, along the vertical direction of the winding.
- $k_{eq_{tan}}$: Tangential direction, along the winding direction.

By solving the corresponding one-dimensional thermal network for Figure 2.14 (further details of the mathematical elaboration in Appendix A), the effective conductivity of foil winding in the orthogonal direction can be expressed as:

$$\frac{1}{k_{eq_{ort}}} = \frac{1}{k_{cond}} \cdot \frac{l_{cond}}{l_{cond} + l_{ins}} + \frac{1}{k_{ins}} \cdot \frac{l_{ins}}{l_{cond} + l_{ins}} \quad (2.11)$$

Which is defined as a parallel association of thermal resistances. It is function of the total width of conductor and insulation layers, l_{cond} and l_{ins} . In the example from Figure 2.14, these quantities are equal to $l_{cond} = l_1 + l_3$ and $l_{ins} = l_2 + l_4$. Their corresponding conductivity are k_{cond} and k_{ins} , respectively.

Regarding the effective conductivity along the top-down direction of the winding, it can be defined as a series association of thermal resistances (see Appendix A):

$$k_{eq_z} = k_{cond} \cdot \frac{A_{cond}}{A_{cond} + A_{ins}} + k_{ins} \cdot \frac{A_{ins}}{A_{cond} + A_{ins}} \quad (2.12)$$

In this case, it is a function of the corresponding heat exchange areas in the vertical direction –labeled as A_1 to A_4 in Figure 2.14, where the total area of conductor layers is $A_{cond} = A_1 + A_3$ and total of insulation is $A_{ins} = A_2 + A_4$ –, and the conductor and insulation thermal conductivity, k_{cond} and k_{ins} .

As a last point, the influence of the tangential conductivity on the heat flow along this direction can be ignored due to the axial symmetry of the foil winding around the center core, and due to the internal heat loss. In other words, the thermal gradient along the same foil layer, in the tangential direction is negligible, so that k_{tan} can be disregarded.

- RZ circular winding:

Figure 2.15: Homogenization of foil winding for cylindrical center leg cores.

Following the same process than the previous rectangular center leg cores, circular windings analyzed in a similar manner than a cylindrical composite wall (more details regarding the mathematical elaboration can be found in Appendix A). An example of this case is shown in Figure 2.15. In this case, the three considered directions are:

- k_{eq_r} : Radial direction, perpendicular to the winding.
- k_{eq_z} : Top-down direction, along the vertical direction of the winding.
- k_{eq_θ} : Tangential direction, along the winding direction.

Its radial effective conductivity is expressed as:

$$k_{eq_r} = \frac{\ln\left(\frac{r_n}{r_1}\right) \cdot k_{cond} \cdot k_{ins}}{k_{ins} \sum_{n_{cond}} \ln\left(\frac{r_{n+1}}{r_n}\right) + k_{cond} \sum_{n_{ins}} \ln\left(\frac{r_{n+1}}{r_n}\right)} \quad (2.13)$$

Due to increment of heat exchange area between the inner and outer layers in the radial direction, a summation of different terms is required for both the conductor and insulation layers ($n_{cond} = 1, 3$ and $n_{ins} = 2, 4$ for the example in Figure 2.15).

Regarding the effective conductivity along the vertical direction is described in the same way than rectangular foil windings, by means of the Equation (2.12). The same reasoning as in the previous case is applied here for the effective conductivity around the axial symmetry, so that $k_{eq_{tan}}$ can be ignored.

2.2.3 Methodology summary

Finally, the two-level homogenization process is summarized in Table 2.1 for the different winding types, to ease the readability. In any case, additional design steps from Section 2.3 will be followed if interleaving or insulation are used.

Table 2.1: Equations summary for different winding types

WINDING TYPE	PROCEDURE
Round solid wire	1) $k_w = k_{cond}$ 2) Eq. (2.7) - (2.9)
Round litz bundle	1) Eq. (2.3) and (2.5) 2) Eq. (2.7) - (2.9)
Square litz bundle	1) Eq. (2.3) and (2.5).
Square foil	1) Eq. (2.11) and (2.12)
Round foil	1) Eq. (2.12) and (2.13)

2.3 Homogenization for inductors and transformers

In the previous section, the generic procedure to homogenize a magnetic component made by a single winding has been explained. That case corresponds to an inductor. However, it might happen that different wires (in size or type) are used in the different windings of a transformer. Additionally, it is usual to interleave the windings of a transformer and use insulation tape between them. What is more, both inductors and transformer usually carry high-frequency current, thus leading to high-frequency effects such as skin, proximity

or gap proximity effects. As a consequence, further considerations have to be carried out when using the homogenization technique for these cases.

2.3.1 Differences between DC and AC

There are some differences between DC and AC excitation in magnetic components, which must be considered. They are summarized next:

- The thermal dynamics are much lower than the electrical dynamics due to AC excitation [20], [21]. A switching converter might operate in the range of hundreds of kHz (period of microseconds), while a thermal transient due to the start-up or a load-step can take several minutes for the temperature to reach a steady value. As a consequence, using the average power losses in a magnetic component instead of the instant value is a valid assumption. This is true for the common size of magnetic components for low and medium power applications; it would not be valid for a magnetic for very low power (microWatts to miliWatts), with a size of a few microns [21].
- Core losses due to DC bias [46] are negligible compared to AC core losses. For AC excitation, additional effects shall be considered (hysteresis, eddy currents and relaxation effect). Subsequently, core losses are distributed non-uniformly in this case. An Electromagnetic-Thermal coupled simulation can be used to accurately model local effects, as discussed in the next subsection. However, considering uniform losses in the core does not means a significant loss of accuracy. This statement will be validated in Section 2.4 with experimental results.
- Winding losses also differ from DC to AC excitation. In high frequency AC, the current density on the conductors tend to be concentrated in the surface [47]. This situation can be aggravated with the presence of another conductor in the vicinity carrying current in the same direction (proximity effect, [48]) or the closeness of the magnetic core gap [49], [50]. As a consequence, the generated power losses are concentrated in the same external annular section than the current density. While for DC excitation, the current density is uniform across the whole section of the conductor, and thus the losses too. However, it will be seen in the next subsection that considering uniform winding losses is a valid assumption for most common magnetics in power electronics.

2.3.2 Losses distribution

Some examples to illustrate the validity of the previous statements are shown next. First, the assumption of uniform winding losses is analyzed. Then, the option of mapping the power losses distribution by means of an EM simulation is explored.

Uniform winding losses assumption

Due to the high thermal conductivity of the conductor, the relatively low diameter of the conductors (lower than 2 mm for low and medium power applications) and the behaviour of heat transfer in radial coordinates [20], [21], it is valid to assume a uniform power losses distribution across the whole section of the conductor.

Let's consider a commercial solid wire of diameter AWG20 (0.81 mm conductor plus 0.02 mm insulation thickness) as an example to illustrate the previous statement. On one hand, a DC current sufficient to generate 2 W is injected through the conductor; on the other hand, a 500 kHz AC current generating the same losses are injected to another wire with the same characteristics. The only difference is that in the second case, losses are concentrated in the 'skin depth' area [47]. The results of a 2D thermal simulation of both cases is shown in Figure 2.16, where the 'skin depth' area is marked. The difference in the temperature distribution is imperceptible (the same scale in both figures).

Figure 2.16: Temperature distribution in the cross section of an AWG20 wire with DC excitation (left) and AC excitation (right).

To see it in more detail, the temperature over the radial axis is plotted in Figure 2.17. It can be seen that the difference is negligible and only differs on the third decimals. However, the next observations can be noticed:

- Since the thermal resistance from the center to the insulation layer decreases as the radius increases [20], [21], then the temperature rise is slightly lower (and flatter) when the losses are concentrated in the ‘skin depth’ area than in the DC case.
- The same temperature is reached in the external conductor surface. Then it starts decreasing with a higher slope due to the insulation.
- Even if the difference between the temperature distribution in DC compared to AC is inconspicuous, the assumption of uniform power losses in AC do not penalizes the accuracy (and can be considered slightly more conservative).

Figure 2.17: Temperature over radial axis in an AWG20 wire with DC excitation (left) and AC excitation (right).

Direct mapping of FEA losses

In order to avoid the possible differences between the assumption of having uniform losses in the core and/or the winding and the actual power losses distribution, it is possible to get the actual power losses distribution map by means of a EM FEA simulation. This power loss distribution is the input for the FEA thermal simulation.

There are two possible ways to couple an EM with a thermal simulation:

- **One-way coupling:** the resulting power loss map is taken as the input of the thermal simulation (Figure 2.18). If the ‘worst case’ temperature of the physical properties of the magnetics are considered, this option is sufficient to get accurate results.
- **Two-way coupling:** several iterations of both the EM and the thermal simulations are performed. This is useful if the temperature dependence of the physical

Figure 2.18: One-way EM-thermal coupling.

Figure 2.19: Two-way EM-thermal coupling.

properties of the magnetic components are characterized in the simulation, such as the permeability or the thermal conductivity (Figure 2.19).

It is important to highlight that if the thermal homogenization is performed, also an EM homogenization would be required to properly characterize the losses distribution due to high-frequency effects. Some approaches for this purpose can be found in [51]–[55].

Nevertheless, it will be seen in Section 2.4 that there is no need of using an EM FEA simulation if power losses are properly calculated and the physical properties of the magnetics are characterized for the ‘worst case’ temperature.

2.3.3 Interleaving techniques and insulation layers

Regarding transformers, it is common using interleaving techniques [56] or insulation tape between layers, due to double insulation requirements like for space applications [57]. In these cases, the next considerations must be taken into account.

Interleaving levels

Depending on the interleaving level, depicted in Figure 2.20 a different procedure must be followed:

- (a) Winding arrangement in blocks (no interleaving). Primary and secondary windings are differentiated, possibly made by wires of different diameter and with different power losses. Then, the corresponding effective thermal conductivity for each winding, k_{eq1} and k_{eq2} , is calculated according to Section 2.2. Regarding the losses assignation, the total primary losses P_{pri} and total secondary losses P_{sec} would be

Figure 2.20: Interleaving levels in a transformer: a) no interleaving; b) layer-by-layer interleaving; c) fully interleaved.

assigned to the blocks k_{eq1} and k_{eq2} , respectively –unless the previously described EM-thermal coupling is performed.

- (b) Layer-by-layer interleaving. The procedure to compute k_{eq1} and k_{eq2} is the same than the previous case but the equivalent blocks are arranged accordingly to the primary and secondary arrangement. In this case, if losses are assumed uniform and the EM-thermal coupling by simulation is not used, each equivalent layer have assigned the proportional part of the losses to the number of layers (N_{layers_1} and N_{layers_2}). For example, if primary winding dissipates 1 W and there are 5 layers, the losses will be equally split among them (1 W/5 = 0.2 W per primary layer).
- (c) Fully interleaved winding. Since primary and secondary wires are alternated within the same layer, the only solution is to homogenize the whole winding into a single block. First, the effective thermal conductivity of each cell is calculated (k_{eq1} and k_{eq2}). Then, (2.12) is used to obtain the effective conductivity of each layer. Finally, depending on the shape of the center leg of the magnetic component, the corresponding equations from Section 2.2.2 will be used to get the total effective conductivity, k_{eqTOT} . In this case, the total winding losses are assigned to the whole homogenized block –again, unless EM-thermal simulation coupling is used.

Specially in the case of high interleaved windings, it is recommended the EM-thermal simulation coupling if possible, since the proximity effect is more patent and the losses distribution is less uniform. For the other interleaving situations, uniform losses can be assumed, as described.

Insulation between layers

Using insulation tape between windings is a common practice, usually mandatory for safety reasons and industry standards. If this is the case, the procedure to perform the homogenization of the winding is depicted in Figure 2.21. The first step (from 1 to 2) is to homogenize each conductor, according to the *wire level homogenization* from Section 2.2.1, thus obtaining their k_w . Each layer separated between tape layers is homogenized afterwards, by following the *winding level homogenization* from Section 2.2.2, assigning to them their corresponding k_{eq} (step 3). Optionally, every insulation tape and winding layer can be homogenized in a single block with conductivity k_{eqTOT} , by treating them as foil layers, using the corresponding equations from Subsection 2.2.2 (step 4).

Until this point, the procedure to obtain a simplified thermal model for any inductor or transformer with round, square or foil winding, also with interleaving or insulation layers, is explained. The proposed homogenization method is applied to 3D FEA simulations to reduce the computation time and is validated by means of experimental results in the next Section.

Figure 2.21: Homogenization procedure for magnetics with insulation between layers.

2.4 Experimental validation and application to 3D FEA

One of the main applications of the proposed thermal homogenization is the use on 3D FEA simulations, to speed them up without a significant loss of accuracy. In this section, the scope of this homogenization is analyzed. Also experimental prototype tests are provided in order to check and validate it.

2.4.1 Experimental setup

Two different scenarios are common for magnetics in power electronics: DC and AC excitation currents. The suggested experimental setup is represented in Figure 2.22. The configuration (a) is valid for DC current. The device under test (DUT) is connected to a constant voltage source, so that a DC current is injected to the magnetic component. In (b), a sinusoidal signal is provided by a waveform generator (GW Instek, model AFG-2005), then amplified by an RF amplifier (AR, model 150A 100B). It is connected afterwards to the magnetic component, so an AC current is supplied. In the case of inductors, (b.1), a capacitor in series is required. The impedance of this resonant series LC tanks is minimum at its resonant frequency, with the aim of injecting a high peak current with an acceptable voltage level at that particular frequency. For transformers, (b.2), a resistive load R_{load} (1Ω in this case) is connected in the secondary side, so that the AC current is flowing through both primary and secondary windings.

Figure 2.22: Experimental setup for DC and AC excitation.

As a result the flowing current, either DC or AC, certain power losses are caused in the tested magnetics. Therefore, their temperature starts rising until a steady thermal equilibrium is reached. By means of this setup, experimental thermal measurements are obtained. The aim of this is comparing them with 3D FEA simulations (labeled 'wire level FEA') and other 3D FEA simulation in which the proposed method is applied (labeled as 'winding level FEA'). The differences between both simulations are described next:

- 'Wire level FEA'. They are fully detailed 3D FEA simulations. Only the insulation of the wires and the strands within a litz bundle are homogenized with the wire level homogenization (Section 2.2), since including those details leads to an excessively demanding mesh [52].
- 'Winding level FEA'. In this case, the suggested two-step homogenization is used. Every wire from the previous step and the material between them are homogenized, according to Sections 2.2 and 2.3. This two-step homogenization leads to a considerably reduced simulation time compared with the 'Wire level FEA' approach, which will be analyzed at the end of this section.

In the physical prototypes, the average surface temperatures of the core and the winding of each magnetic component are measured with a FLUKE-Ti400 thermal camera, following the good practices described in [58]–[60]. The thermography is post-processed by the FLUKE Smart View software. The comparison between these measurements and 3D simulations is performed. The 3D simulations are performed with Icepak (ANSYS Electronics Desktop 2019.1).

Relative error metrics

To compare the thermal camera results and the 3D simulations, the surface temperature of the camera are considered as the reference. Then, the relative error between the measured temperature rise ($\Delta T_{exp} = T_{exp} - T_{amb}$) and the temperature rise obtained by the corresponding 3D simulation ($\Delta T_{sim} = T_{sim} - T_{amb}$) is calculated according to (2.14).

$$Error_{T_{surf}}[\%] = \left(\frac{\Delta T_{sim}}{\Delta T_{exp}} - 1 \right) \cdot 100 \quad (2.14)$$

Since the thermal camera is not able to reach the hot-spots, maximum temperatures can only be analyzed by simulation. The 'Wire level FEA' 3D simulations (just wires and litz bundles are homogenized) is considered as the reference for the maximum temperatures. Then, they are compared with the 'Winding level FEA' 3D simulations (two-stage homogenization is performed). The relative error of these magnitudes is calculated by (2.15):

$$Error_{T_{max}}[\%] = \left(\frac{\Delta T_{windFEA}}{\Delta T_{wireFEA}} - 1 \right) \cdot 100 \quad (2.15)$$

Prototypes description

Several prototypes are tested, in order to cover the whole scope of common applications in magnetics for power electronics. There are different inductors and transformers with various core shapes and winding types; they are described in Table 2.2. The next thermal conductivity has been considered for the different objects or materials in all cases: $k_{fe} = 4.5$ [W/m·K] for the ferrite cores [61], $k_a = k_{imp} = 0.03$ [W/m·K] (conductivity of air at 353 [K] from [62]) for the still air surrounding the wires and the 'impregnation material' of the litz bundles, $k_{bund} = 0.155$ [W/m·K] and $k_{ist} = 0.245$ [W/m·K] for the bundle insulation and strand insulation materials [39]. The accuracy for the DC and AC tests, as well as the simulation time reduction, are analyzed along the next paragraphs.

Table 2.2: Description of the tested prototypes

Prototype	A (inductor)	B (inductor)	C (inductor)	D (inductor)	E (transformer)
Core	P36/22	P36/22	PQ32/30	PQ32/30	RM8/I
Winding type	Solid	Litz 35 x 0.2mm	Litz 35 x 0.2mm	Foil (0.27mm·2) (cop.+iso.+cop.+iso.)	Solid
ϕ [mm]	0.81 (AWG20)	1.45 (AWG15)	1.45 (AWG15)	—	pri=0.51 (AWG24) sec=0.81 (AWG20)
insulation thickness[mm]	0.02	0.055	0.055	65 μ m (·2)	0.02
w [mm]	0.87	1.6	1.6	4	pri=0.64 sec=1
Number of turns	92	28	42	22	pri=28 sec=15
k_w [W/m·K]	390	0.205309	0.205309	390	390
k_{eq} [W/m·K]	0.179535	0.079864	0.079864	1.4094	pri=0.0639 sec=0.0721

2.4.2 DC tests

First, DC current is injected to the inductors 'A', 'B', 'C', 'D' in order to have certain winding conduction losses, according to Figure 2.22. Core losses only correspond to the DC bias effect and can be considered negligible, compared to the winding losses. Table 2.3 summarizes the power losses and ambient temperature for each test. The temperature distribution obtained by the thermal camera and the 'Winding level FEA' 3D simulations are shown in Figure 2.23.

Figure 2.23: Temperature distribution of the DC tests: 'Winding level FEA' simulations (top) and thermal camera pictures (bottom).

Table 2.3: DC test conditions

Prototype	A	B	C	D
$T_{amb}[\text{°C}]$	26	24	26	20
$P[\text{W}]$	3.793	3.244	3.160	4

Average surface and maximum temperatures of each device are analyzed next to validate the proposed model for DC excitation. A set of bar graphs are shown for each prototype. They correspond to the experimental measurements, 'Wire level FEA' and 'Winding level FEA' obtained temperatures. Notice that the bars for experimental maximum temperatures are not shown since that data is not available by means of the camera. The error margin inherent to the measurements, which is related to the discrepancy of the measured surface emissivity and the background radiation, is included for every measurement.

Prototype A

As the first proof of concept of the homogenization at winding level, a pot core inductor with round solid wire is tested (prototype 'A'). Since the winding is surrounded by the core almost completely, the winding is not directly exposed to ambient. In this way, the possible source of error due to the modelling of the boundary conditions in simulation is mitigated.

The results are shown in Figure 2.24. It can be seen that the simulation with the equivalent layer ('Winding level FEA') deviates only a 5 % for the winding surface temperature estimation ($T_{w_{surf}}$) and -2.2 % for the core surface temperature ($T_{c_{surf}}$).

Furthermore, the evaluation of the maximum temperatures in the winding ($T_{w_{max}}$) and the core ($T_{c_{max}}$) respectively have -7.5% and -4.4% deviation, compared to the detailed 'Wire level FEA' simulation.

Figure 2.24: DC test results of prototype 'A'.

Figure 2.25: DC test results of prototype 'B'.

Prototype B

To go one step further, prototype 'B' is built with litz wire. So, the performance of the homogenization at both wire and winding levels together is verified. The data of the litz wires provided by the manufacturers have certain tolerance, so there is some added uncertainty related to their physical properties.

However, it can be seen in Figure 2.25 that the results of the 'Winding level FEA' simulation are close to the 'Wire level FEA' estimations. The evaluation of $T_{w_{surf}}$ has almost no deviation compared to the experimental measurement

(1 % and -3.7 % respectively) and a slightly higher for $T_{c_{surf}}$ (9.2 % and 10.1 %). This mismatch is due to a deviation of the boundary conditions in simulation with respect to the measurements.

The difference in the tendency of the results of prototypes ‘A’ and ‘B’ is explained next. Prototypes ‘A’ and ‘B’ use a P36/22 core, but ‘A’ is made by solid wires and ‘B’ by litz bundles. After the ‘winding homogenization’, the wires are replaced by a block. This change in the geometry leads to a different external surface exposed to the ambient (lower external surface after the homogenization). Since the solid wires from ‘A’ are more compactly packed than ‘B’, the decrement of external surface is lower in ‘A’ than in ‘B’. Consequently, the impact of convective heat transfer is stronger in ‘A’ than in ‘B’ after the winding homogenization. This is the reason why the maximum winding temperature after the homogenization is higher after the winding homogenization (two-level homogenization) in ‘A’ and lower in ‘B’, compared to their corresponding wire homogenization (first level homogenization).

Despite the difference between the measurements and the simulations, the most important fact is the correlation between simulations. Furthermore, that difference is almost within the error band inherent to measurements. There is a low difference between the detailed ‘Wire level FEA’ and the completely homogenized ‘Winding level FEA’, it can be seen that the geometry modification does not have a significant impact on the accuracy. However, the second one is much faster, as analyzed in Section 2.4.5.

Prototype C

In order to check the validity of the two-step homogenization to different core shapes, a PQ core inductor is built (prototype ‘C’). In this case, the winding is directly in contact with the ambient. Another uncertainty appears due to this additional condition, as previously explained. This is the reason of a higher deviation of $T_{w_{surf}}$ (Fig. 2.26). But still, the highlight of this comparison is the similarity of the results of both simulations, so the ‘Winding level FEA’ simulation offer the additional advantage of having a much simpler mesh.

Prototype D

Since the formulation of k_{eq} is also provided for foil magnetics, a PQ core inductor with foil windings is tested (prototype ‘D’). As shown in Fig. 2.27, the estimation of $T_{w_{surf}}$ and $T_{c_{surf}}$ has an acceptable deviation from the measurements. Comparing this case with the

previous ones, the evaluation of the maximum temperatures are slightly higher with the 'Winding level FEA' model.

Figure 2.26: DC test results of prototype 'C'.

Figure 2.27: DC test results of prototype 'D'.

2.4.3 AC tests

At this point, the proposed two-step homogenization is validated for different core shapes and winding types for DC excitation. In this subsection, the previous inductors ('A', 'B', 'C' and 'D') and a transformer ('E') are tested with AC excitation. In high-frequency excitation, certain local effects appear, like skin and proximity effects. Actually, in the transformer 'E', proximity effect becomes more critical. These high-frequency effects create certain local concentration of power losses in the winding. To recreate this power losses distribution, an EM simulation is performed, whose losses are directly mapped as the input of the thermal simulations [22]. As stated in Section 2.3.2, in order to couple an EM simulation with the 'Winding level FEA' thermal simulation some homogenization

2.4. Experimental validation and application to 3D FEA

techniques to calculate the EM equivalent properties are required. Some examples can be found in [51]–[55]. For the sake of simplicity, the approach from [51] is used in this case.

Table 2.4: AC test conditions

Prototype	A	B	C	D	E
$T_{amb}[^{\circ}\text{C}]$	24	22	23	22	20
$f[\text{kHz}]$	62	219	155	107	100
$I_{pk-pk}[\text{A}]$	3.2	3.49	4.74	3.68	$pri = 3.68$ $sec = 6.64$

The setup for AC excitation is depicted in Figure 2.22. The peak-to-peak of the sinusoidal current and frequency for each case (equal to the resonant frequency of the LC tank), as well as the ambient temperature, are summarized in Table 2.4. The temperature maps obtained in simulation and by measurements are shown in Fig. 2.28.

Figure 2.28: Temperature distribution of the AC tests: 'Winding level FEA' simulations (top) and thermal camera pictures (bottom).

As a consequence of the AC excitation, it is important to notice that two additional variables are added, compared to the DC tests, which are additional sources of deviation between measurements and simulations: core loss and the EM homogenization. Despite this fact, high accuracy is also guaranteed for AC currents.

Prototype A

In this case, the difference with the average surface temperatures is more noticeable than in the previous DC excitation. The main reason is the uncertainty associated to the EM material properties and the calculated effective EM properties for the winding. Despite the deviation between measurements and simulations, it is more important to analyze the error between the 'Wire level FEA' and the 'Winding level FEA', which have close results as depicted in Figure 2.29. This means that the same temperature estimation can be achieved with a much simpler simulation.

Prototype B

Due to the lower electrical resistance of the litz winding of prototype 'B', compared to 'A', the component achieves a lower temperature rise for a current with similar magnitude (Table 2.4). Simulation results are within the measurement error margin and close to each other (differences of 2.5 % and 1.2 % for maximum temperatures), as shown in Figure 2.30.

Prototype C

By means of this prototype, the validation is carried out for a litz based winding in a core shape highly exposed to the ambient with AC excitation. Due to hardware limitations, the peak-to-peak current is set as high as possible (Table 2.4). Even with this current, the reached temperature rise is lower than the previous cases due to the low electrical resistance of the winding. As a consequence, the relative error between measurements and errors is higher (Figure 2.31). However, in terms of absolute error, the difference is just around 5°C.

Prototype D

The homogenization of foil winding with AC current is validated by means of this prototype. The same conclusion can be extracted than in Prototype D, except that the deviation of the simulation results is within the measurement error margin (Figure 2.32).

Prototype E

Finally, the last concept to be proven is the homogenization of transformers. This prototype is added to the experimental test. It consists of an RM8/I core with solid copper wires of different diameters in the primary winding than in the secondary winding. The corresponding setup for this case is in shown Figure 2.22, with the excitation values from Table 2.4. The achieved results are shown in Figure 2.33. It can be seen that the deviation between simulations and measurements is lower than 6.3 %. Furthermore, the results of both 'Wire level FEA' and 'Winding level FEA' differ in less than 5.2 % between each other.

Figure 2.29: AC test results of prototype 'A'.

Figure 2.30: AC test results of prototype 'B'.

Figure 2.31: AC test results of prototype 'C'.

Figure 2.32: AC test results of prototype ‘D’.

Figure 2.33: AC test results of prototype ‘E’.

2.4.4 DC and AC results summary

To summarize the previous DC and AC test results, the next general conclusions can be extracted:

- Thermal measurements have an inherent error due to the discrepancy in the emissivity of the measured surfaces and the background radiation [58]–[60]. This error band is marked in the previous measurement results.
- The difference between the 3D thermal simulations using both the ‘Wire level FEA’ approach or the proposed ‘Winding level FEA’ and measurements is mainly affected by the tolerance of the available thermal and electromagnetic material properties data provided by the manufacturers.
- There are two additional variables in the AC tests: core loss and the electromagnetic homogenization required for the ‘Winding level FEA’. These variables are other

sources of error. Nevertheless, the proposed homogenization has been validated both for DC and AC excitation.

- Due to the geometric modification caused by the ‘Winding level FEA’ approach, the impact of the convective heat transfer is slightly varied compared with the detailed ‘Wire level FEA’. However, the difference between each other is lower than 5°C for most cases, except for the foil inductor, where it is lower than 9°C.

At this point, the two-step homogenization is validated for both DC and AC excitation for different core shapes and winding types. Furthermore, the most remarkable key point is the accuracy of the introduced two-stage thermal homogenization (‘Winding level FEA’), while leading to lower time-consuming simulations, compared to detailed simulations (‘Wire level FEA’). This achieved reduction of the computational time is analyzed next.

2.4.5 Computational time reduction

Once the accuracy of the method is quantified and validated, the next main feature is analysed: the time reduction of 3D FEA thermal simulations. Table 2.5 demonstrates that the homogenized ‘Winding level FEA’ 3D simulations lead to a time reduction of the computation time by a factor between 6 and 8, compared to the detailed ‘Wire level FEA’ 3D simulations.

These results are computed with the Icepak package from ANSYS Electronics Desktop 19.1, running Intel Core i7-6700 CPU and 32 GB RAM. The mesh level is set as default and 100 iterations, with the solve setting set to ‘Flow=0.0001’ and Energy to ‘1e-8’.

This statement can be illustrated by means of Figure 2.34 and 2.35, which represent the effect of the two-step homogenization on the meshing reduction. The first one is a cross-section of a solid wire with insulation, whose mesh count is 1936 elements. By considering the process from Section 2.2, the mesh elements are reduced to 542 elements. The same effect can be contemplated for the litz bundle, in which a higher mesh count of 3239 elements is reduced to 542. This is just an example of the the cross-section of

Table 2.5: Computation time for prototypes A, B, C, D and E with the Wire level FEA and the Winding level FEA homogenization.

Prototype	A	B	C	D	E
Wire level FEA	20’29”	19’25”	20’17”	10’55”	25’46”
Winding level FEA	2’55”	3’08”	2’48”	1’32”	3’11”
Reduction factor	7	6.2	7.2	7.1	8.1

one wire or bundle, but take into account that the meshing count of the whole component rises exponentially due to the abundant number of conductors in a winding.

It is important to highlight that the convergence of these simulations is ensured. Fully detailed simulations including the insulation of each wire or the strands were unfeasible due to an excessive mesh count, which led to convergence problems. So, at least, the first step ‘Wire level’ homogenization was required. To conclude, it can be stated that by means of the proposed two-step homogenization, that meshing count is drastically reduced, thus leading to the time reduction shown in Table 2.5 while achieving almost the same performance than a much more detailed 3D simulation.

2.5 Application to thermal networks

Thermal networks are a commonly used tool to model the thermal behavior of magnetic components, due to the good compromise between simplicity and accuracy. Nevertheless, stating a thermal network implies certain simplifications which might lead to inconsistency of the results. This loss of accuracy can be mitigated by means of the proposed k_{eq} from Section 2.2.

2.5.1 Implementation of the thermal network

Some examples of thermal networks for magnetics can be found in [63]–[66]. In these references, the magnetic component is divided in several regions. Each one is made of a current source, representing the power losses in that region, and four thermal resistances, to account for the conductive heat transfer in each direction of the region. An additional thermal resistance is used to model the heat exchange between the external surfaces and the ambient. The values of the internal thermal resistances depend on the geometry of

Figure 2.34: Meshing of a solid wire with (left) and without insulation (right).

Figure 2.35: Meshing of a litz bundle (left) and a homogenized bundle (right).

the region and its thermal properties; the external resistances depend on the convective and radiative conditions of the outer surfaces. The equations to obtain the values of the thermal resistances is described in [20], [21].

However, all the approaches from [63]–[66] use a coarse simplification regarding the properties of the winding: they consider the winding as a solid block, whose thermal conductivity is set equal to the copper conductivity. However, as demonstrated in Section 2.2, this is far from reality. The winding is not only made of conductor material, there is still air or potting material surrounding every wire, thus lowering the equivalent conductivity of the homogenized block one or two orders of magnitude lower than copper (from 400 W/m·K to 40 W/m·K or even less than 1 W/m·K, depending on the size and properties). To contrast this traditional assumption with the proposed k_{eq} , the accuracy is analyzed for a generic inductor next.

Figure 2.36: Cross-section of a 3D thermal FEA simulation of a P36/22 core inductor with 21 turns of AWG15 (left) and the corresponding 2D thermal network (right).

2.5.2 Validation

To demonstrate that the accuracy of thermal networks can be improved by means of k_{eq} (Section 2.2), a P36/22 core with 21 turns of AWG15 is analyzed as an example, Figure 2.36. The winding and core losses are 1 W and 0.2 W, respectively. The convective coefficient is assumed as 20 W/m²·K for the side face and 10 W/m²·K for the top surface. Heat flow to the center and to the bottom is neglected (adiabatic lines), thus simplifying the thermal network. The inductor is split in 6 regions.

A detailed 3D FEA thermal simulations is performed and considered as a reference. It is compared with a common 2D thermal network. The thermal network is calculated according to the assumption of considering the whole winding block with the conductivity of copper, according to [42]. Then, it is compared by a thermal network in which the proposed k_{eq} (Section 2.2) is used.

The results of the temperature rise in the winding and the core are summarized in Table 2.6. It can be seen that the assumption made by the approach from [65] leads to a large error in the estimation of the maximum temperatures. However, this difference is much lower by using the proposed k_{eq} .

It is important to notice that the accuracy of the thermal network can be increased even more by splitting the inductor in more regions (the higher the granularity of the network, the better the accuracy [21]).

Table 2.6: Thermal networks and 3D FEA results for a generic P36/22 inductor.

	3D FEA	[65]	Proposed
$\Delta T_{max_{wind}}$ [°C] (Error[%])	30.2	2.5 (-91.7 %)	20.6 (-31.8 %)
$\Delta T_{max_{core}}$ [°C] (Error [%])	14.1	1.8 (-87.2 %)	12.3 (-10.6 %)

2.6 Chapter summary and contributions

A two-step homogenization (k_{eq}) is introduced in this chapter to simplify the thermal analysis of magnetic components for power electronics, while keeping highly accurate results. The developed equations and specific guidelines to model round solid and litz wires, square litz wire and foil wound magnetics are described. Additional concerns regarding some special cases, like interleaving techniques and the use of insulation tape between layers, are discussed.

The proposed approach has been validated for DC and AC excitation, by means of diverse prototypes. Inductors and transformers with different core shapes and winding types have been tested, to cover the scope of most common applications of magnetics in power electronics.

The homogenization can be applied either to 3D FEA thermal simulations and to thermal networks. In the first case, the homogenization of the winding leads to a required simulation time between 6 and 8 times lower than a fully detailed 3D FEA simulation. What is more, this approach ensures the convergence of 3D simulations, by avoiding excessively demanding meshes due to complex geometries or the small detail of the strands within a litz bundle winding. In the case of thermal networks, the use of k_{eq} leads to an increased accuracy, compared to traditional thermal networks.

Regarding the impact on the simplification of 3D FEA simulations, which has been proven, the idea of using this fast and accurate simulations within optimization processes is a suitable option that can lead to better optimized magnetics. As a proof of concept of this idea, a FEA-based optimization process using the method from this chapter will be described and analyzed in Chapter 4.

Contributions related to this chapter

Journal publications:

1. [67] **G. Salinas**, A. Delgado, J. Muñoz-Antón, J. A. Oliver and R. Prieto, “Fast and Accurate Thermal Modeling of Magnetic Components by FEA-based homogenization,” in *IEEE Transactions on Power Electronics (TPEL)*.
doi: 10.1109/TPEL.2019.2921160
url: <https://ieeexplore.ieee.org/document/8731691>
2. [55] A. Delgado, **G. Salinas**, J. A. Oliver, J. A. Cobos and J. Rodríguez-Moreno, “Equivalent Conductor Layer for Fast 3D Finite Element Simulations of Inductive Power Transfer Coils,” in *IEEE Transactions on Power Electronics (TPEL)*.

doi: 10.1109/TPEL.2019.2921160

url: <https://ieeexplore.ieee.org/document/8882243>

Conference papers:

1. [68] **G. Salinas**, A. Delgado, J. A. Oliver and R. Prieto, “Fast FEA Thermal Simulation of Magnetic Components by Winding Equivalent Layers,” 2018 IEEE Energy Conversion Congress and Exposition (**ECCE**), Portland, OR, USA, 2018, pp. 7380-7385.

doi: 10.1109/ECCE.2018.8557714

url: <https://ieeexplore.ieee.org/document/8557714>

2. [54] A. Delgado, **G. Salinas**, J. A. Oliver, J. A. Cobos and J. Rodríguez, “Equivalent Parameters of Round and Litz Wire Conductors to Obtain an Equivalent Layer to Accelerate Finite Element Simulations of Wireless Power Transfer System,” 2018 IEEE Energy Conversion Congress and Exposition (**ECCE**), Portland, OR, USA, 2018, pp. 7375-7379.

doi: 10.1109/ECCE.2018.8558310

url: <https://ieeexplore.ieee.org/document/8558310>

3. [69] A. Delgado, **G. Salinas**, J. Rodríguez, J. A. Oliver and J. A. Cobos, “Finite Element Modelling of Litz Wire Conductors and Compound Magnetic Materials based on Magnetic Nano-particles by means of Equivalent Homogeneous Materials for Wireless Power Transfer System,” 2018 IEEE 19th Workshop on Control and Modeling for Power Electronics (**COMPEL**), Padua, Italy, 2018, pp. 1-5.

doi: 10.1109/COMPEL.2018.8460012

url: <https://ieeexplore.ieee.org/document/8460012>

Model extraction from FEA simulations

Using an accurate thermal model during the design of a magnetic component can be decisive to get the smallest possible design, while complying with the defined specifications by the industry. However, the most common approach is using analytic models and just using Finite Element Analysis tools as a final verification of a prototype [29], not during the design stage. This procedure often leads to some issues: the component does not meet thermal specifications, thus requiring a redesign of the component, or it meets the constraints but it is not as small as possible; it may also occur that both the specifications and the lowest size are reached, but it would be ‘by chance’. To avoid these situations, it is possible to develop a thermal model based on FEA simulation results, taking the advantage of their accuracy.

Figure 3.1: Maximum temperature rise over winding losses at constant core losses: 3D FEA results (red), $R_{th} = 53 \cdot V_e [cm^3]^{-0.54}$ [24] (blue) and proposed approach (green).

The need of an accurate, easy-to-handle and generic thermal model is patent. In general, the simplest approaches provide limited accuracy, and the complex thermal networks show good results at the expense of higher computational cost. In order to combine the good qualities of these approaches (simplicity and accuracy), a matrix thermal

model is proposed in this chapter, whose parameters are obtained by means of FEA tools. Figure 3.1 illustrates the temperature rise of a generic transformer for a sweep of the generated power losses. The temperature estimated by the proposed model is compared with the 3D FEA simulation results and the estimation by the commonly used equation from [24], which leads to a considerable deviation. The main feature of the proposed model is that it consists of a simple coefficients matrix, which is simple to use, and leads to a significant accuracy from T_{amb} until the limit temperature, T_{lim} , since it is obtained by the linearization and superposition of FEA results. Once the matrix is obtained, no more simulations are required, so its use is as simple and fast as a set of linear equations. Furthermore, this matrix is able to consider the heat transfer between each part of the magnetic component, what is too complex to be obtained by any other approach. Then, it allows to predict the temperature rise at any region of the component for any generated power losses, in contrast to the approach from [24], which considers the magnetic component as a single temperature node and is only applicable if there is natural convection.

In this chapter, a static thermal impedance matrix is developed by means of 3D FEA thermal simulations. As a result, an accurate estimation of the maximum and mean surface temperatures of each part of the magnetic component (core, primary, secondary, auxiliary, bobbin, etc) can be obtained for any operating point of the power converter.

First, the proposed static FEA-based thermal model is explained in general terms. Then, the specific remarks required when building the model for transformers and inductors are discussed. Finally, an experimental validation of the model is shown.

3.1 Need of a new model

The thermal behaviour of a magnetic component is one of the main limiting factors to design it. It can be as small as it is able to properly evacuate its power losses (core and winding losses), keeping its temperature below safe thermal limits or specifications. Actually, the tendency is going to higher power density components, what makes even more important to use an accurate thermal model which properly predicts the temperature rise of the component under operation.

Using the empirical equations from ferrite core manufacturers is one of the easiest approaches [24] or derived simple equations like the one proposed by Escibano in [25]. Nevertheless, their main drawback is that they consider any magnetic component as a single temperature node. This assumption itself, despite resulting in a very simple approach, is false in most cases since both core and winding usually reach different

temperature. Furthermore, any empirical approach is only valid in the same operating conditions as the experiments were performed (i.e. ambient temperature, forced or natural convection, etc). These approaches should be used just as a first order approximation.

There are more complex models, like the thermal models from [7] and [64], which are based on thermal networks. It shows a good correlation with a FEM thermal simulation, but the assigned boundary conditions were set as fixed film coefficients, so the fluid dynamics were not included in the simulation. Furthermore, another possible source of error is the assumption of the heat sources (power losses) as concentrated in certain nodal points of the resistance network. The reality is that losses are distributed across the whole solid bodies. As a result, this approach might have significant deviations.

A different approach is building a model based on experimental results, like [8]. Other similar approach is using an inverse model in which the correction of some analytical equations is made by means of prototype results [9]. Both have the inherent drawback of requiring a prototype and limiting the extrapolation of results to similar conditions than during the experiments.

On the other hand, virtual prototyping techniques are possible with Finite Element Analysis (FEA) tools. In [10] some guidelines are proposed to develop a complete coupled model with FEA for semiconductors. The same strategy can be used for any other device, since it is proposed at mesh level. As a counterpart, techniques such as model order reduction are required to handle the complete system [11]. To avoid this, a higher level thermal model based on 2D and 3D FEA is proposed in [12], which can be combined with [13] and [14] to get a complete electro-thermal model.

Nevertheless, the previously described approaches might result too complex to be implemented in a circuit simulator or to be used to design a magnetic component. In this chapter, a thermal coefficients matrix built by means of 3D FEA thermal simulations is proposed. Being a matrix makes the model easy to handle and to be implemented in any circuit simulation. Additionally, that matrix contains the required information to separately estimate the temperature rise at any region of the magnetic component (primary, secondary and auxiliary windings, core, bobbin and connectors) at any operating point of the power converter in which it is placed. This is useful to predict and avoid the saturation of the magnetic core due to a high temperature and to comply with very restrictive standards, like space applications [57], [70]–[72], airborne equipment [73], [74] or vehicle wireless chargers [75], in terms of maximum allowed temperature. The process to obtain this model is explained along the next sections.

3.2 FEA-based static thermal model

The main feature of the proposed FEA-based static thermal model is that it contains the required information to accurately estimate both the maximum and the average surface of any identified object within a system, under any operating point. This can subsequently be extended to inductors or transformers, being ‘systems’, where the magnetic core, each winding, and any surrounding accessories (like a plastic coil former or the pin connections to a PCB) are the objects composing those systems. As a result, it is possible to avoid overcoming safe thermal limits established by specifications or the corresponding industrial standard.

The key-point of using 3D Finite Element Analysis thermal simulations is to take advantage of their accuracy, as well as their capability to analyze complex systems. Any magnetic component can be considered as a complex system due to their geometry and heterogeneity of their parts. In these components, it is very difficult to compute the interaction between each part by means of analytical approaches. This limitation is overcome by FEA simulations. Furthermore, any boundary condition can be recreated by simulation, regardless its nature or complexity. For example, the component can be analyzed in conditions of natural or forced cooling. Advanced cooling systems like heat-pipes or heat-sinks combined with fans can also be included in the simulations.

The proposed model is described in the following; a flow diagram which summarizes the procedure to obtain the model is depicted in Figure 3.2.

Figure 3.2: Procedure to obtain the proposed FEA-based thermal model.

Design procedure summary

- 1) *Split the magnetic component in representative objects.* The model can be as granular as desired; it is able to predict the temperature in N objects. However, one 3D FEA simulation is required per object. So, it is recommended to split the magnetic

component in the least number of representative objects as possible. For instance, it would not be practical to analyze each wire separately, but it would be desirable to analyze the group of wires from each winding (i.e. primary; secondary; auxiliar) and the core.

- 2) *Establish the maximum allowed surface temperature and the ambient temperature.* By defining these quantities, the maximum allowed temperature rise per object is calculated, ΔT_{sim} . This value is usually set by specifications. In any other case, it must be established under safe thermal limits.
- 3) *Calculate the corresponding P_{test} for each object.* By means of simple formulae, the excitation (power loss) for the next thermal simulations, required to build the model, is calculated.
- 4) *Perform one 3D FEA simulation per object.* The whole magnetic component will be simulated, but losses will be assigned only to one object at a time, with the value obtained from the previous step, while the other objects will have assigned null losses. Power losses can be considered uniform and the thermal properties constant (an equal to the value at T_{sim}) without a relevant loss of accuracy, as demonstrated in the experimental validation from Chapter 2. The temperature rise of each object will be stored in every simulation. The information of two different quantities can be stored to create the model:
 - a) Maximum temperature rise per object. This might be the most critical quantity, since it should be used to check the compliance with specifications.
 - b) Mean surface temperature rise per object. Sometimes, the surface temperature is also a constraint by specifications. It is useful to have this information too.
 - c) Both. Including the information of both maximum and surface temperatures is very useful, but the resulting model might be too complex, depending on the final application.
- 5) *Create the coefficients matrix by using the stored information from the previous step.* This matrix represents the ‘thermal resistance’ of the magnetic component, though it does not strictly correspond to the physical definition of thermal resistance. This matrix has the same units as a thermal resistance, $[\frac{K}{W}]$, but it is not obtained as an expression dependent on the material properties and dimensions of the magnetic component. The terms of this matrix are calculated as the temperature of each object divided by the losses injected in each simulation. In this way, the self-heating per object, as well as the interaction (heat flux) between objects, are registered in

this matrix. Nevertheless, this step implies a linearization of the thermal behaviour of the component from T_{amb} to T_{slim} , to apply the principle of superposition [23]. Hence, it is more accurate for temperatures close to both limits of that range than in the middle of it, as previously shown in Figure 3.1.

- 6) *Get the temperature rise of every object that the magnetic component is composed of when applying certain power losses.* The use of the thermal coefficients matrix follows the same application rules than a common thermal resistance, with the difference that the equation will be presented in matrix form: $[\Delta T] = [R_{th}] \cdot [P]$.

Along the next paragraphs, the methodology is conceptually explained for a single object. Then, the extension of the analysis to N objects is presented.

3.2.1 Concept for a single object

In order to understand the concept of the proposed model, its application to a single object is explained in this section.

One-dimensional heat transfer without heat generation is simple and well known [20]. An analogy with electrical circuits can be used to model this thermal problem, where the thermal resistance between two nodes can be calculated as the temperature increment over the heat flow through them, and correspond to simple geometric relationships. However, this analogy is no longer valid when some heat generation is produced within the analyzed objects. In this case, the temperature distribution is non-linear and some simplifications are required to develop a thermal network for this case, like the one proposed by Fagundes in [63]. But this simplification might lead to undesirable deviations compared to reality.

Figure 3.3: Diagram of the model extraction for one object.

Additionally, heat transfer in magnetic components can rarely be assumed as one-dimensional. Usually, magnetic components are soldered on a PCB, whose conductivity is relatively low, and the film coefficient is dissimilar on the side faces than the top faces. As a consequence, heat transfer is different in each spatial dimension, having

a three-dimensional behaviour. Furthermore, the geometric expressions for the thermal resistances for three dimensions might become too complex, thus the solution of the heat transfer equation too. For this reason, Finite Element Analysis tools are chosen to develop the proposed model and overcome this problem.

Let's use the simple geometry from Figure 3.3 to illustrate the proposed model. Following the previous schema, the first step (splitting the geometry in representative objects) is skipped in this case, since a single object is analyzed.

Establishing maximum operating and ambient temperatures

Secondly, the maximum allowed operating conditions and the ambient temperature must be established. For example, the maximum surface temperature that the object should not exceed can be set as $T_{slim} = 120^\circ C$, and the ambient temperature equal to $T_{amb} = 20^\circ C$. Then, this constraint can be expressed as a maximum allowed surface temperature rise:

$$\Delta T_{slim} = T_{slim} - T_{amb} \quad (3.1)$$

Calculating P_{test}

The next step is defining the power loss that will be assigned in the next required 3D FEA thermal simulation of the analyzed object, P_{test} . This is a first guess by means of analytical expressions. It is calculated so that the object reaches the established T_{slim} , under the desired boundary conditions (i.e. defined ambient temperature, natural or forced convection, radiation, surrounding objects).

Assuming natural convection and radiation for this example, the heat that the object is capable to dissipate by means of these mechanisms while provoking a surface temperature rise equal to ΔT_{slim} is calculated. It is also assumed that the object is laying on a surface at fixed temperature, which is similar to having an electronic device on a PCB.

Regarding the natural convection, some empirical expressions are developed by Cengel in [21] and summarized Table 3.1. For the object of this example, the film coefficient for each external surface ('surf' index) must be calculated. Then, the heat that is dissipated via convection is expressed in terms of those film coefficients and the corresponding heat exchange areas (A) and ΔT_{slim} , according to (3.2).

$$\begin{aligned}
 Q_{conv} &= \sum_{surf} (h_{surf} \cdot A_{surf} \cdot \Delta T_{slim}) = \\
 &= (h_{sides} \cdot A_{sides} \cdot \Delta T_{slim}) + (h_{top} \cdot A_{top} \cdot \Delta T_{slim})
 \end{aligned} \tag{3.2}$$

In a similar manner, the heat dissipated via radiation is computed as:

$$Q_{rad} = \varepsilon \cdot \sum_{surf} A \cdot \sigma \cdot (T_{slim}^4 - T_{amb}^4) \tag{3.3}$$

Where ε is the relative emissivity of the object, σ is the Stefan-Boltzmann constant ($5.670373 \cdot 10^{-8} \frac{W}{m^2 \cdot K^4}$). The temperatures T_{slim} and T_{amb} must be expressed in Kelvin, due to the raising to the fourth.

Then, the value of the heat generation assigned to the object in the simulation is the sum of the heat that can be dissipated by convection and radiation:

$$P_{test} = Q_{conv} + Q_{rad} \tag{3.4}$$

This value of P_{test} is assigned to the object in the thermal simulation. It will cause a surface temperature rise close to ΔT_{slim} , but not equal. This is why a 3D FEA simulation is used. It will capture the actual non-linear behaviour of convection and radiation.

Table 3.1: Film coefficient for natural convection for typical geometries [20], [21].

Object	Film coefficient
Vertical plate or cylinder	$h = 1.42 \cdot \left(\frac{\Delta T_s}{L} \right)^{0.25} ; L = height$
Horizontal cylinder	$h = 1.32 \cdot \left(\frac{\Delta T_s}{D} \right)^{0.25} ; D = diameter$
Horizontal plate hot surface up	$h = 1.32 \cdot \left(\frac{\Delta T_s}{L} \right)^{0.25} ; L = \frac{4 \cdot Area}{perimeter}$
Horizontal plate hot surface down	$h = 0.59 \cdot \left(\frac{\Delta T_s}{L} \right)^{0.25} ; L = \frac{4 \cdot Area}{perimeter}$

Simulation setup

In this case, only one 3D FEA thermal simulation is required, since just one object is analyzed. In any other case, the number of required simulations will be equal to the number of chosen representative objects, N . It is very important to highlight that the simulation

setup must capture the boundary conditions that the object would experiment in the real setup. For example, if the object is going to be exposed to any cooling system, like heat-pipes or fans, those must be included in the simulation setup. Surrounding objects, like PCBs, bobbins or a metallic structure, must also be included in order to recreate the real situation of the analyzed component. It is strongly recommended to simulate the fluid dynamics, so that the actual surrounding fluid flow (usually air) is developed and the use of simplified film coefficients is avoided [20]; higher accuracy will be achieved in this way.

Regarding the example from Figure 3.3, natural convection and radiation will be activated in the simulation. The PCB on which the object is lying must also be drawn. Finally, the previously calculated P_{test} losses are assigned to the object. The thermal conductivity of every participating element must also be configured. The setup will be static, since the steady-state regime is analyzed and not the dynamic behaviour. At this point, the simulation can be performed.

Extracting results to build the thermal coefficients ‘matrix’

Depending on the information required by the user, two quantities can be extracted from the simulation: the maximum temperature of the object or the mean surface temperature. Considering both of them would lead to a model able to predict both the maximum and the surface temperatures for any power losses.

Just to illustrate the potential of the model, both temperatures are considered for this example. But they will be separately treated. In other words, two different matrices will be created.

Since only one object is treated, each ‘matrix’ will contain just one element. On the one hand, the matrix containing the behaviour of the maximum temperature of the object is equal to:

$$R_{th_{max}} = \frac{T_{max_{object}} - T_{amb}}{P_{test}} = \frac{\Delta T_{max_{object}}}{P_{test}} \quad (3.5)$$

On the other hand, the ‘matrix’ containing the information of the mean surface temperature is expressed as:

$$R_{th_s} = \frac{T_{s_{object}} - T_{amb}}{P_{test}} = \frac{\Delta T_{s_{object}}}{P_{test}} \quad (3.6)$$

The previous matrices will contain the same number of elements than the chosen number of representative objects, N .

Using the thermal coefficients ‘matrix’

Once the thermal coefficients matrices are obtained, they can be used in a similar manner than any other thermal resistance. For instance, being P_{loss} (in Watts) the internal heat generation (or power losses) in the object, its maximum temperature and mean surface temperature are respectively equal to:

$$T_{max_{object}} = R_{th_{max}} \cdot P_{loss} + T_{amb} \quad (3.7)$$

$$T_{s_{object}} = R_{th_s} \cdot P_{loss} + T_{amb} \quad (3.8)$$

It is important to highlight that by means of the thermal coefficients matrices, the maximum temperature and the mean surface temperatures of the object can be calculated for any value of P_{loss} .

3.2.2 Modelling N objects

The process to expand the model to N objects is very similar to the model for a single object. There are some additional details, like the importance of subdividing the magnetic component in representative objects, and some slight differences, like the dimensions of the resulting matrix due to the higher number of objects.

Guidelines to identify representative objects

The proposed model can be extended to an infinite number of objects. This would lead to a model capable to predict the temperature in those infinite objects. However, subdividing the analyzed device in the minimum representative parts (objects) is key to avoid an excessive complexity of the resulting matrix ($N \times N$) while keeping the proper relevant information. As a general rule, an ‘object’ should comprehend pieces of the analyzed device which share a physical similarity (i.e. made by the same materials, similar size or similar boundary conditions).

For instance, considering the imaginary system from Figure 3.4 and assuming the parts with the same color are made of the same material, a proper subdivision of objects would be next:

- *Object #1*: It is constituted by a lot of several smaller parts, which are made of the same material and are very close to each other.

Figure 3.4: FEA-based thermal model for N objects: the black arrows represent heat flux through the external surface of each object to the ambient; the red arrows the heat flux between objects, by any heat transfer mechanism (conduction, convection and radiation).

- *Object #2:* This object is made of a different material and, also, has different boundary conditions (object #3 is in contact with it).
- *Objects #3 and #4:* These objects are made of the same material. But they do not share the same conditions; #3 is in contact with #2 and the ambient, but #4 is only surrounded by the ambient.

It is important to highlight that even if two elements are not touching each other (there is no conductive heat transfer between them), they can interact by means of convection and/or radiation.

Establishing the maximum operating temperature of each object and the ambient temperature

It is possible that two objects, made of different materials, have different maximum operating temperatures. This is not a problem, the only difference is that those objects will have a different T_{slim} . Considering this statement and the same ambient temperature for all objects, the maximum allowed surface temperature rise must be calculated for each one according to Equation (3.1) ($\Delta T_{slim,1}, \Delta T_{slim,2}, \dots, \Delta T_{slim,N}$). As a consequence, this will affect the value of the assigned power losses to those objects in the simulation.

Calculating P_{test} for each object

Since each object has different geometry, boundary conditions (convection and radiation) and different ΔT_{slim} in some cases, the value of P_{test} for each object ($P_{test_1}, P_{test_2}, \dots, P_{test_N}$) must be calculated according to Equations (3.2), (3.3) and (3.4).

Simulation setup

Regarding the simulation setup, the general comments explained in Section 3.2.1 are applicable for N objects. The only difference here is that the number of required simulations is equal to the number of objects, N . The difference between simulations is that losses are assigned to one object and the rest have no heat generation. For the example in Figure 3.4, four simulations with the next power losses are required:

- *Simulation #1*: object 1 with power losses equal to P_{test_1} ; others, zero.
- *Simulation #2*: object 2 with power losses equal to P_{test_2} ; others, zero.
- *Simulation #3*: object 3 with power losses equal to P_{test_3} ; others, zero.
- *Simulation #4*: object 4 with power losses equal to P_{test_4} ; others, zero.

Building the thermal coefficients matrix

From the previous simulations, the maximum temperature and the mean surface temperature in each object are stored. Two matrices can be built. Their dimension is $N \times N$.

Regarding the first one, the elements of each column stores the maximum temperature rise of the object per simulation (the first subscript of each temperature rise is the number of the object; the second is the number of the simulation) divided over the assigned P_{test} in that simulation ($P_{test_1}, P_{test_2}, \dots, P_{test_N}$). Each colour of the matrix denotes that the numbers in that column are extracted from the same simulation. So this matrix is arranged according to (3.9).

Regarding the second matrix, each column stores the mean surface temperature rise per object at each simulation (the first subscript of each temperature rise is the number of the object; the second is the number of the simulation), divided over the assigned P_{test} . In his way, the matrix is arranged according to (3.10).

The matrix $[R_{th_{max}}]$ contains more critical information, since it will be used to estimate the hotspots in every object. And hotspots are usually the most restrictive quantity. A

$$[R_{th_{max}}] = \begin{pmatrix} \frac{\Delta T_{1max,1}}{P_{test_1}} & \frac{\Delta T_{1max,2}}{P_{test_2}} & \dots & \frac{\Delta T_{1max,N}}{P_{test_N}} \\ \frac{\Delta T_{2max,1}}{P_{test_1}} & \frac{\Delta T_{2max,2}}{P_{test_2}} & \dots & \frac{\Delta T_{2max,N}}{P_{test_N}} \\ \vdots & \vdots & \ddots & \vdots \\ \frac{\Delta T_{Nmax,1}}{P_{test_1}} & \frac{\Delta T_{Nmax,2}}{P_{test_2}} & \dots & \frac{\Delta T_{Nmax,N}}{P_{test_N}} \end{pmatrix} \quad (3.9)$$

$$[R_{th_s}] = \begin{pmatrix} \frac{\Delta T_{1s,1}}{P_{test_1}} & \frac{\Delta T_{1s,2}}{P_{test_2}} & \dots & \frac{\Delta T_{1s,N}}{P_{test_N}} \\ \frac{\Delta T_{2s,1}}{P_{test_1}} & \frac{\Delta T_{2s,2}}{P_{test_2}} & \dots & \frac{\Delta T_{2s,N}}{P_{test_N}} \\ \vdots & \vdots & \ddots & \vdots \\ \frac{\Delta T_{Ns,1}}{P_{test_1}} & \frac{\Delta T_{Ns,2}}{P_{test_2}} & \dots & \frac{\Delta T_{Ns,N}}{P_{test_N}} \end{pmatrix} \quad (3.10)$$

clear example is the saturation of ferrite materials. If the temperature at any point of a ferrite core reaches the *Curie temperature*, it would saturate. Using this matrix is very useful to predict and avoid this situation.

Nonetheless, $[R_{th_s}]$ will be used to estimate the mean surface temperature per object. This quantities might also be constraint by specifications in certain industrial applications, like Wireless Power Transfer coils, or in consumer electronics, like wearables or smartphones.

For both matrices, the diagonal elements represent the self-heating effect of each object due to their internal heat generation. While the other elements in the matrices represent the heat exchange between each other. For example, in case that an object is isolated or too far from others, the corresponding term would be zero. If two objects are not in contact (no conduction heat transfer), it does not mean that the corresponding term in the matrix is zero because it might still interact with another object by means of convection and radiation. And here lays the most important feature of the presented matrices. They model the heat transfer between any object through any heat transfer mechanism, no matter how complex the geometry or the boundary conditions are.

Using the thermal coefficients matrix

Once the matrices are obtained, they are ready to predict the temperature of each object for any combination of heat generation (power losses).

Their use is simple and is governed by the rules of any thermal resistance, with the difference that they have a matrix representation. So, in order to obtain the vector containing the maximum temperature rise in each object $(\Delta T_{max1}, \Delta T_{max2}, \dots, \Delta T_{maxN})$ for any value of losses in each object $(P_{loss1}, P_{loss2}, \dots, P_{lossN})$, Equation (3.11) is used.

$$\begin{pmatrix} \Delta T_{1max} \\ \Delta T_{2max} \\ \vdots \\ \Delta T_{Nmax} \end{pmatrix} = \begin{pmatrix} \frac{\Delta T_{1max,1}}{P_{test1}} & \frac{\Delta T_{1max,2}}{P_{test2}} & \dots & \frac{\Delta T_{1max,N}}{P_{testN}} \\ \frac{\Delta T_{2max,1}}{P_{test1}} & \frac{\Delta T_{2max,2}}{P_{test2}} & \dots & \frac{\Delta T_{2max,N}}{P_{testN}} \\ \vdots & \vdots & \ddots & \vdots \\ \frac{\Delta T_{Nmax,1}}{P_{test1}} & \frac{\Delta T_{Nmax,2}}{P_{test2}} & \dots & \frac{\Delta T_{Nmax,N}}{P_{testN}} \end{pmatrix} \cdot \begin{pmatrix} P_1 \\ P_2 \\ \vdots \\ P_N \end{pmatrix} \quad (3.11)$$

While to calculate the mean surface temperature rise of each object for a particular combination of power losses per object the Equation (3.12) is used instead.

$$\begin{pmatrix} \Delta T_{1s} \\ \Delta T_{2s} \\ \vdots \\ \Delta T_{Ns} \end{pmatrix} = \begin{pmatrix} \frac{\Delta T_{1s,1}}{P_{test1}} & \frac{\Delta T_{1s,2}}{P_{test2}} & \dots & \frac{\Delta T_{1s,N}}{P_{testN}} \\ \frac{\Delta T_{2s,1}}{P_{test1}} & \frac{\Delta T_{2s,2}}{P_{test2}} & \dots & \frac{\Delta T_{2s,N}}{P_{testN}} \\ \vdots & \vdots & \ddots & \vdots \\ \frac{\Delta T_{Ns,1}}{P_{test1}} & \frac{\Delta T_{Ns,2}}{P_{test2}} & \dots & \frac{\Delta T_{Ns,N}}{P_{testN}} \end{pmatrix} \cdot \begin{pmatrix} P_1 \\ P_2 \\ \vdots \\ P_N \end{pmatrix} \quad (3.12)$$

The inherent features of the previous matrices makes possible the calculation of the temperature rise for any number of objects with any value of internal heat generation. As a consequence, it is a very suitable model for inductors and transformers, which are made of several parts or objects, with different power losses in each of them (any operating point). This will be explained in Section 3.3.

3.2.3 Impact and limitations

The main advantages and scope of the developed model can be summarized next:

- Its use can be extended to N objects. Any component can be subdivided in N objects, whose behaviour will be represented in the resultant matrix.
- Self-heating effects of each object, as well as heat transfer between objects by conduction, convection and radiation, are considered.

- Its use can be extended for any operating conditions, in terms of heat generation. The matrix can be used to compute the maximum and/or the mean surface temperatures, given any value of the internal heat generation (power losses) per object. If applied to an electrical component, this thermal model can be used to calculate the thermal rise of the component for any electrical operating condition.
- Valid for the boundary conditions that set in the required simulations, no matter their complexity. For example, if an IC package is being analyzed, which is mounted on a PCB and assembled within an enclosure with heat-pipes and radiation, the model will be valid for those conditions as long as they are reproduced in the required simulations to get the model.
- The accuracy is ensured from T_{amb} to T_{slim} , due to the linearization between those temperatures. As a consequence, it allows to properly identify the actual thermal limits of the analyzed device, ensuring that the established temperature limits are not exceeded. Additionally, if used within an optimization process, it would help to avoid over-sized design solutions.

Despite the wide scope and applicability of the proposed approach, it has some limitations:

- The developed model is only valid for the boundary conditions (convection and radiation) established in the simulations. For example, if natural convection is set in the required simulations, the resulting matrix cannot be used to extrapolate the temperature rise of the analyzed device in an environment with forced convection; and the other way around. So, a different matrix must be created for different boundary conditions.
- Since it is a static model and the dynamic behaviour is not included, the implementation for circuit simulators would require techniques such as moving window integration, or a combination with a dynamic model (see Appendix B). However, the obtained results for steady-state are reliable, as shown in Section 3.4.
- The obtained model has its maximum accuracy for temperatures close to T_{slim} and to T_{amb} . The accuracy is slightly lower in the middle of that range. This is due to the linearization that is assumed from T_{amb} to T_{slim} in order to apply superposition. This step is required to avoid the non-linearity of the radiation heat transfer and the turbulent air flow [20], [21].

The previous aspects should be kept in mind to ensure a proper usability of this model. By setting the desired boundary conditions of the analyzed objects, the proposed model offers crucial thermal information which can be used to ensure a safe operation of any magnetic component working at any operating point given by the complete electronic system.

3.3 Model extraction for transformers and inductors

Once the thermal coefficients matrix model has been defined for N selected objects, certain additional particularities must be considered when applying it to transformers and inductors. Some general guidelines to split the magnetic component into the most representative *objects* are clarified next, as well as the distinction between objects with internal power losses and passive objects.

Splitting the magnetic component into representative objects

Despite the versatility of the matrix model, identifying the fewest representative objects as possible of a magnetic component is crucial to obtain a matrix easy to handle while representing the most relevant information. As a general rule, both inductors and transformers should be split into the next objects:

- *Magnetic core*: the permeability and saturation magnetic flux of ferromagnetic materials, like soft ferrites, is dependent on the temperature. As a consequence, evaluating its temperature is very important. The magnetic core can be considered as a single object or it can be split into two parts if the boundary conditions are very differentiated between the upper and the lower halves.
- *Main windings*: every winding with circulating current must be considered, since they have internal losses due to conduction (Joule effect) and high frequency effects (skin and proximity effects). They should be treated as separated objects according to their differences (i.e. different power losses and diameter). For example, if a transformer has a single main primary winding and two main secondary windings, they should be considered as three different objects. In the case of inductors, only one winding is considered as an object, even if several parallel wires are arranged.
- *Auxiliary windings*: some transformers usually have an auxiliary winding (or multiple) in order to supply control or protections circuitry, usually driving low current levels. Even if they have negligible power losses, their temperature might be

critical for design purposes or safety. So, each auxiliary winding should be considered as an independent object. Specific details must be considered for them when building the matrix model, as discussed later in this section.

- *Passive parts:* some parts of the magnetic component might be critical in certain designs, like the coil former in which the windings are wound or the pin connectors to a PCB. If the temperature of those elements is considered critical for the current design, they should be included in the analysis and treated as independent objects in the model.

Establishing maximum operating temperature of each object and ambient temperature

The ambient temperature is the same for the whole magnetic component. Nevertheless, it might occur that each object has a different maximum temperature established by specifications. This would lead to a different T_{slim} for each object. For example, soft ferrites are usually considered stable below $T < 100^{\circ}\text{C}$, while conductor wires might be able to stand higher temperatures limited by their insulation (usually $T < 120^{\circ}$, depending on the materials). However, some standards or specifications for space industry, airborne equipment, electrical vehicle chargers, wearables or consumer electronics might be even more restrictive and must be considered for this step, depending on the environment of the project.

Calculating P_{test} for the magnetic core, each winding and passive objects

Due to the nature of each part of the magnetics, different remarks are considered when calculating their P_{test} for the later simulations.

Regarding the magnetic core, its $P_{test_{core}}$ is calculated according to Equations (3.2), (3.3) and (3.4), using the corresponding film factors from Table 3.1 as described in Section 3.2.1. The surfaces in contact with a PCB will be considered as adiabatic, hence neglected for this step. If heatsinks and/or forced convection are used as cooling systems, the simplified film factor will be calculated for each surface according to [20], [21], [76].

For the wires of wound magnetics, since only the wires in the outer layer are in direct contact with ambient, all windings are treated as a group. An example is depicted in Figure 3.5. The group of windings is treated as a vertical cylinder (Table 3.1), with the additional remark that the heat exchange area must account for the external surface of each

wire, as highlighted in Figure 3.5. Assuming the window height to compute the external surface of the winding would lead to a wrong value: $A_{ext} = (2 \cdot r \cdot n_{wires}) \cdot \text{perim}_{wind}$. On the other hand, if properly considering the surface of each wire, the external surface must be:

$$A_{ext} = (\pi \cdot r \cdot n_{wires}) \cdot \text{perim}_{wind} \quad (3.13)$$

Where n_{wires} is the number of wires in the external layer of conductors, r is their radius and perim_{wind} is the perimeter of that layer around the core.

Figure 3.5: Cross-section of a generic magnetic component: consideration of the external surface for the calculation of P_{test} for the winding.

Then, $P_{test_{wind}}$ is calculated, with the normal procedure (Section 3.2.1). Finally, the losses assigned to each winding in their corresponding simulation are equal to $P_{test_{wind}}$, according to (3.3), regardless the number of main windings or wires per winding. This is because P_{test} is calculated only considering the external surfaces in contact to the ambient, and the winding only dissipates its losses to ambient through the external layer of conductors.

$$P_{test_{pri}} = P_{test_{sec}} = \dots = P_{test_{wind}} \quad (3.14)$$

In the case of foil windings, no special concern is required regarding the ‘extra’ surface of the windings (the heat exchange area with the ambient corresponds to a cylinder). But the same assignation of $P_{test_{wind}}$ must be considered, according to (3.3).

Auxiliary windings usually have negligible power losses compared to the main windings. Hence, no additional simulations are required for them. There is no need to compute a value of P_{test} for them as a consequence. Nevertheless, their temperature should be stored after the other required simulations so that this information will be included in the thermal coefficients matrix. The same reasoning is applied for any other passive object (like the

3.3. Model extraction for transformers and inductors

bobbin or the pin connectors to the PCB); they will not require additional simulations, but they will be considered when building the thermal matrix.

Required simulations: example of a generic transformer

A 3D FEA thermal simulation is required for each object with internal heat dissipation (power losses). In every simulation, the maximum and the average surface temperature rises of the objects defined in the first step must be stored. Taking the generic transformer from Figure 3.6. The chosen representative objects are: magnetic core, primary winding, secondary winding, auxiliary winding and the coil former. Only the first three objects have power losses (they are active objects), so just three simulations are required. Each simulation has losses in one active object, with the corresponding value calculated in the previous step.

Figure 3.6: Required simulations and obtained results for a simple PQ transformer (quarter symmetry representation). Natural convection and adiabatic bottom surface of the core conditions are considered as boundary conditions.

Building the thermal matrix: example of a generic transformer

Continuing with the example from Figure 3.6, the number of objects is five (magnetic core, primary winding, secondary winding, auxiliary winding and coil former). Then, the matrix containing the required information to calculate the maximum temperature of each object correspond to the next expression:

$$[R_{th_{max}}] = \begin{pmatrix} \frac{\Delta T_{core-max,1}}{P_{test_{core}}} & \frac{\Delta T_{core-max,2}}{P_{test_{pri}}} & \frac{\Delta T_{core-max,3}}{P_{test_{sec}}} & 0 & 0 \\ \frac{\Delta T_{pri-max,1}}{P_{test_{core}}} & \frac{\Delta T_{pri-max,2}}{P_{test_{pri}}} & \frac{\Delta T_{2max,N}}{P_{test_N}} & 0 & 0 \\ \frac{\Delta T_{sec-max,1}}{P_{test_{core}}} & \frac{\Delta T_{sec-max,2}}{P_{test_{pri}}} & \frac{\Delta T_{sec-max,2}}{P_{test_{sec}}} & 0 & 0 \\ \frac{\Delta T_{aux-max,1}}{P_{test_{core}}} & \frac{\Delta T_{aux-max,2}}{P_{test_{pri}}} & \frac{\Delta T_{aux-max,3}}{P_{test_{sec}}} & 0 & 0 \\ \frac{\Delta T_{bobbin-max,1}}{P_{test_{core}}} & \frac{\Delta T_{bobbin-max,2}}{P_{test_{pri}}} & \frac{\Delta T_{bobbin-max,3}}{P_{test_{sec}}} & 0 & 0 \end{pmatrix} \quad (3.15)$$

Each column contains the maximum temperature rise of each object obtained in a simulation, divided by the corresponding P_{test} of that simulation. This happens for simulations 1 to 3. It can be seen that columns 4 and 5 are filled with zeroes, since they are associated to the auxiliary winding and the coil former, which have null losses and they do not require an additional simulation for them. Nevertheless, columns 4 and 5 must be included with the aim of calculating the maximum temperature of these objects in the last step of the procedure, which will be provoked by the power losses from nearby objects.

Additionally, the corresponding matrix to calculate afterwards the average surface temperature of each object in this example corresponds to (3.16).

$$[R_{th_s}] = \begin{pmatrix} \frac{\Delta T_{core-s,1}}{P_{test_{core}}} & \frac{\Delta T_{core-s,2}}{P_{test_{pri}}} & \frac{\Delta T_{core-s,3}}{P_{test_{sec}}} & 0 & 0 \\ \frac{\Delta T_{pri-s,1}}{P_{test_{core}}} & \frac{\Delta T_{pri-s,2}}{P_{test_{pri}}} & \frac{\Delta T_{2max,N}}{P_{test_N}} & 0 & 0 \\ \frac{\Delta T_{sec-s,1}}{P_{test_{core}}} & \frac{\Delta T_{sec-s,2}}{P_{test_{pri}}} & \frac{\Delta T_{sec-s,2}}{P_{test_{sec}}} & 0 & 0 \\ \frac{\Delta T_{aux-s,1}}{P_{test_{core}}} & \frac{\Delta T_{aux-s,2}}{P_{test_{pri}}} & \frac{\Delta T_{aux-s,3}}{P_{test_{sec}}} & 0 & 0 \end{pmatrix} \quad (3.16)$$

The same remarks are highlighted for this matrix, regarding columns 4 and 5.

Calculating the maximum and the average surface temperatures of each part of the magnetic component

Finally, the maximum and the average surface temperatures of each part of the magnetic component can be obtained by multiplying the previous matrices for any value of generated power losses, following the same procedure than in Section 3.2.2. Then, the maximum temperature rise in each object for certain combination of power losses per object is calculated by (3.17).

$$\begin{pmatrix} \Delta T_{core_{max}} \\ \Delta T_{pri_{max}} \\ \Delta T_{sec_{max}} \\ \Delta T_{aux_{max}} \\ \Delta T_{bobbin_{max}} \end{pmatrix} = [R_{th_{max}}] \cdot \begin{pmatrix} P_{core} \\ P_{pri} \\ P_{sec} \\ P_{aux} \\ P_{bobbin} \end{pmatrix} \quad (3.17)$$

In a similar manner, the average surface temperature rise of each object for the same combination of power losses is calculated according to (3.18).

$$\begin{pmatrix} \Delta T_{core_s} \\ \Delta T_{pri_s} \\ \Delta T_{sec_s} \\ \Delta T_{aux_s} \\ \Delta T_{bobbin_s} \end{pmatrix} = [R_{th_s}] \cdot \begin{pmatrix} P_{core} \\ P_{pri} \\ P_{sec} \\ P_{aux} \\ P_{bobbin} \end{pmatrix} \quad (3.18)$$

Having the capability of analyzing the temperature of each identified part of the magnetic component (either an inductor or a transformer) for any power losses is very valuable for power electronics designers. By means of the model described in this chapter, it is possible to check the thermal limits specifications of their magnetics under any electrical operating point.

3.4 Experimental validation

In this section, an experimental setup is presented in order to check and validate the proposed FEA-based thermal model. First, a description of the devices under test and the electrical setup are described. Then, some aspects to recreate the boundary conditions of the experiments in the 3D FEA simulations are discussed. Finally, the estimated temperature rise of the tested magnetics by means of these models are compared with thermal measurements.

3.4.1 Devices under test and experimental setup

Table 3.2 contains the description of the tested prototypes. A generic inductor and a generic transformer with three windings (primary, secondary and auxiliary) with different core shapes are chosen as representative samples in order to cover the scope of the proposed model. They are shown in Figure 3.7.

Table 3.2: Description of the tested prototypes

Prototype	A (inductor)	B (transformer)
Core	P36/22	E25/13/7
ϕ [mm]	1 (AWG19)	0.81 (AWG21)
gap [mm]	0.44	None
Number of turns	30	9:20:10
T_{amb} [°C]	26	26

Figure 3.7: Tested prototypes and resonant capacitors: a) P36/22 inductor, b) E25/13/7 transformer.

Regarding the thermal measurements, an OEM-PLUS Series fiber optics thermometer was used to register the hot-spot temperatures and the dynamic change of the temperature rise of the devices. Additionally, a thermal picture was obtained when the steady state temperature was reached by means of a FLUKE-Ti400 thermal camera and processed by means of the FLUKE Smart View software. In the inductor, two thermocouples were used: one was placed between the base of the core and the PCB in which it is mounted; the second was in the middle of the winding. For the transformer, four thermocouples were used instead: between the core and the PCB, between the bobbin and the primary winding, in the middle of the secondary winding, and in the interface between secondary

and auxiliary windings. A cross-section of both magnetics showing the location of the thermocouples is represented in Figure 3.8.

Figure 3.8: Cross-section of the tested prototypes and location of the thermocouples.

In order to build the matrix model and compare it with the measurements, two objects are identified in the inductor (core and winding), and four objects in the transformer (core, primary, secondary and auxiliary windings). Since the thermocouples are approximately located at the hotspot of each object and the thermal camera shows their surface temperatures, both maximum and mean surface temperatures will be stored in the later simulations to get the corresponding matrices.

AC excitation is injected to the prototypes in order to generate core and winding power losses, hence provoking certain temperature rise. The electrical setup required to do so is depicted in Figure 3.9. A sinusoidal voltage waveform at certain frequency is generated by a waveform generator, amplified by means of an RF amplifier and supplied to the rest of the circuit. The value of the capacitor C_{res} is such that the resonant frequency with the Device Under Test (DUT) corresponds to the input signal frequency or is close to it. In this way, the impedance of this resonant tank is low enough to get a relatively high input current with a relatively low input voltage, thus avoiding the saturation of the RF amplifier (due to its limited voltage gain). Since the voltage is imposed by the RF amplifier, a variable resistor (R_{damp} , R_{load}) is placed in order to adjust the demanded current by the circuit. In this way, the voltage and current can be selected as desired. The values of every component and operating conditions are summarized in Table 3.3.

The experiments were taken in natural convection and radiation conditions. As shown in Figures 3.7 and 3.8, the prototypes are mounted on a PCB which is laying on a wooden table. These conditions were replicated in the simulations. The fluid dynamics were

Figure 3.9: Electrical setup for the tested inductor (a) and transformer (b).

Table 3.3: Components and operating conditions of the prototypes

	A	B
C_{res}	$3.3 \text{ nF} \times 2$	$4.7 \text{ nF} \times 3$
L, L_m, L_{lk}	$500 \text{ } \mu\text{H}, -, -$	$-, 169 \text{ } \mu\text{H}, 830 \text{ nH}$
R	$0.5 \text{ } \Omega$	$1.8 \text{ } \Omega$
V_{pk-pk}	976.6 V	104.7 V
I_{pk-pk}	2.26 A	5.19 A
$freq$	175 kHz	340 kHz
T_{amb}	$26 \text{ } ^\circ\text{C}$	$26 \text{ } ^\circ\text{C}$

simulated as well to accurately define the behaviour of the surrounding air flow and the subsequent heat exchange by convection.

3.4.2 Obtained matrix models

At this point, the procedure to obtain the matrices $[R_{th_{max}}]$ and $[R_{th_s}]$ for both prototypes, described in Section 3.3, is followed. The ambient temperature of the simulations was set equal to the experiments, $T_{amb} = 26 \text{ } ^\circ\text{C}$. The limit surface temperature is established as $T_{slim} = 100 \text{ } ^\circ\text{C}$, resulting in a temperature rise of $\Delta T = 74 \text{ } ^\circ\text{C}$, what is used for the calculation of P_{test} for each object. The calculated values of P_{test} for both prototypes are summarized in Table 3.4, considering natural convection and radiation.

For the inductor ‘A’, the resulting matrices are (3.19) and (3.20), being the core and the winding the objects 1 and 2, respectively.

Table 3.4: Values of P_{test} for the prototypes

P_{test} [W]		A	B
Core		4.493	1.721
Winding	Main	2.171	–
	Pri	–	1.622
	Sec	–	1.622

$$[R_{th_{max}}]_A = \begin{pmatrix} 15.27 & 21.36 \\ 14.53 & 26.27 \end{pmatrix} \begin{bmatrix} K \\ \frac{K}{W} \end{bmatrix} \quad (3.19)$$

$$[R_{th_s}]_A = \begin{pmatrix} 14.02 & 16.01 \\ 14.45 & 26.65 \end{pmatrix} \begin{bmatrix} K \\ \frac{K}{W} \end{bmatrix} \quad (3.20)$$

While for the transformer ‘B’ the matrices are are (3.21) and (3.22), being core, primary, secondary and auxiliary windings the objects 1, 2, 3 and 4, respectively.

$$[R_{th_{max}}]_B = \begin{pmatrix} 31.5 & 40.1 & 32.4 & 0 \\ 28.6 & 54.6 & 36.8 & 0 \\ 27.9 & 40.0 & 48.9 & 0 \\ 26.0 & 38.3 & 30.7 & 0 \end{pmatrix} \begin{bmatrix} K \\ \frac{K}{W} \end{bmatrix} \quad (3.21)$$

$$[R_{th_s}]_B = \begin{pmatrix} 27.9 & 26.7 & 25.1 & 0 \\ 28.2 & 52.1 & 32.8 & 0 \\ 25.9 & 32.1 & 41.4 & 0 \\ 25.2 & 31.6 & 29.0 & 0 \end{pmatrix} \begin{bmatrix} K \\ \frac{K}{W} \end{bmatrix} \quad (3.22)$$

3.4.3 Comparison with measurements

In this section, the performance of the obtained matrices for both prototypes is analyzed. First, by considering the operating conditions from Table 3.3, the power losses are calculated. Since the excitation is sinusoidal, Steinmetz equation is used to estimate the core loss [77], [78], and the AC resistance to calculate the winding losses is obtained by means of an ANSYS PEmag [79], [80]. The corresponding losses for the inductor and the transformer are summarized in Table 3.5. These calculated losses are the input for the thermal model.

Table 3.5: Calculated power losses for the inductor and transformer prototypes.

Inductor		Transformer	
P_{core} [W]	1.095	P_{core} [W]	0.1
P_{wind} [W]	0.937	P_{pri} [W]	0.3
		P_{sec} [W]	0.8
		P_{aux} [W]	0

Steady-state temperature

In this case, after the magnetics have reached a thermal equilibrium, the temperature is kept constant at certain value. This steady-state temperature is the temperature of interest that can be obtained with the $[R_{th}]$ matrix. A comparison between the results of the model, measurements and 3D FEA simulations is carried out next.

In the prototypes, the maximum temperatures are obtained by means of thermocouples, shown in Figure 3.10. In the inductor, the temperature on the base of the core ($\Delta T_{c_{base}}$) and in the middle of the winding ($\Delta T_{w_{mid}}$) are measured. It is important to notice that the maximum temperature of the core is not located in the base, but on the top surface. It has been inspected by means of the thermal camera that there is a difference of 6°C between base and top. Hence, the maximum temperature of the core is considered as $\Delta T_{max}|_{core} = \Delta T_{c_{base}} + 6$. Regarding the winding temperature, the thermocouple was placed in the geometrical middle part of the winding. However, the maximum temperature might have close but different location, thus adding certain error band in the measurement.

The same conclusions are considered for the transformer. Nevertheless, the location of the thermocouples are in: 1) base of the core (being in this case $\Delta T_{max}|_{core} = \Delta T_{c_{base}} + 5.6$), 2) between the bobbin and the primary winding, 3) in the middle of the secondary winding, 4) between secondary and auxiliary windings.

Figure 3.10: Thermocouple temperature rise of the prototypes: a) inductor, b) transformer.

Figure 3.11: Thermal camera measurements of the prototypes: a) inductor, b) transformer.

The surface temperatures are measured by a FLUKE-Ti400 thermal camera, following the good practices described in [58]–[60]. The thermography is then post-processed to obtain the average surface temperatures by the FLUKE Smart View. The thermal pictures are shown in Figure 3.11. The average surface temperature of the core is labeled as $\Delta T_s|_c$. Only the external layer of conductors can be measured, labeled as $\Delta T_s|_w$. In the inductor, it corresponds to the average winding surface temperature. In the transformer, it corresponds to the surface temperature of the secondary winding; the surface temperature of the auxiliary winding is labeled as $\Delta T_s|_{w_{aux}}$.

The power losses are the input for any thermal model or simulation. So, the proposed matrix and the corresponding simulation have exactly the same value of power losses, shown in Table 3.5. Since these input losses are estimated, they are subjected to certain deviation that propagates to the obtained temperature. This is the reason why 3D FEA simulations are also added in the comparison: the results from $[R_{th}]$ and the simulation have exactly the same input, so that they can directly be compared. The comparison with measurements is also done as demonstration and validation. It is important to notice that the difference between the estimation made by $[R_{th}]$ and the experimental measurements is not only due to the error of $[R_{th}]$ but due to the inherent error of the measurements (due to a mismatch between the assumed emissivity of the measured surfaces compared to the actual one, and due to the background radiation) plus the uncertainty associated to the power losses estimation.

Finally, the comparison of the temperature measurements and the temperature estimation is shown in Figures 3.12 and 3.13. The maximum temperature rise at each region of the inductor and the transformer is collected in the top graphs; the average surface temperature rise is collected in the bottom graphs. Each quantity has three bars: the purple one represents the experimental measurements; the orange one is the temperature obtained by 3D FEA simulations; and the green one is the temperature

Figure 3.12: Maximum (left) and average surface (right) temperature rises of the analyzed inductor in the core and the winding. Comparison of: a) Experimental measurements (purple), b) 3D FEA simulations (orange), c) Proposed $[R_{th}]$ matrices. The error between $[R_{th}]$ and the 3D simulations is highlighted with the corresponding percentage, since that approach is originated from those simulations. Experimental results are shown as proof and validation.

Figure 3.13: Maximum (top) and average surface (bottom) temperature rises of the analyzed transformer the core and the primary, secondary and auxiliary windings. Comparison of: a) Experimental measurements (purple), b) 3D FEA simulations (orange), c) Proposed $[R_{th}]$ matrices. The error between $[R_{th}]$ and the 3D simulations is highlighted with the corresponding percentage, since that approach is originated from those simulations. Experimental results are shown as proof and validation.

estimation by using the proposed $[R_{th}]$ matrices. The error margin in the measurements is denoted in the corresponding bars. The experimental surface temperature of the primary winding of the transformer is not available since the thermal camera only captures the external layer of conductors. The percentage shown over the estimated temperatures with $[R_{th}]$ corresponds to the deviation between this model and the 3D FEA simulations. This represents the measurement of the intrinsic error of the $[R_{th}]$ model. It is calculated as:

$$Error[\%] = \left(\frac{\Delta T_{[R_{th}]}}{\Delta T_{3D}} - 1 \right) \cdot 100 \quad (3.23)$$

It can be seen that the absolute error between the proposed model and the experimental results is lower than 5°C , and the relative error between the proposed model and the 3D FEA simulations ranges between -6.4% and 3.9% .

Results conclusions

The main observations that can be stated by the previous results are summarized next:

- Regarding the direct comparison between the proposed $[R_{th}]$ matrix and the 3D FEA simulations, the results show a good accuracy, with an error lower than 5% in most cases.
- The deviation compared with measurements are mainly due to the uncertainty associated to the power losses estimation, which are the input for both the $[R_{th}]$ model and the 3D FEA.
- As a detail to notice, the estimation on the surface temperatures also differ from the measurements because every surface of each object is used in the simulations to compute the average surface temperature. However, only the external surfaces are measured with the thermal camera.
- Despite the difference with measurements, the accuracy of $[R_{th}]$ is proven by its comparison with 3D simulations. Furthermore, the results are within the error band of measurements or close to them.
- Natural convection and radiation were the thermal conditions for the prototypes. Both boundary conditions are non-linear. And, despite this fact the results of the proposed $[R_{th}]$ are proven accurate. Then, it can be stated that this model can also be used for any other boundary condition, like heat-pipes or forced convection.

- An inductor and a transformer with different core shapes, winding arrangement and operating conditions were tested. In this way, the scope of $[R_{th}]$ is proven for a wide variety of the most common configurations of high-frequency magnetics for power electronics. So, the proposed model can be used to characterize any magnetic component.

3.5 Chapter summary and contributions

A methodology to extract thermal models from 3D Finite Element Analysis thermal simulations is developed in this chapter, as one of the main contributions of this thesis. It can be generalized to any component, but the corresponding particularities to model inductors and transformers are explained. The main characteristics of this FEA-based thermal models is the accuracy and the usability of the models for any combination of power losses in the magnetic component. The accuracy of these models is validated by experimental results and 3D FEA simulations.

By using these FEA-based thermal models in optimization processes, over-sized designs would be avoided, pushing the power density to the limits while ensuring a safe operation within the established thermal specifications, as proven in the next chapter.

Additionally, as a minor contribution, Appendix B describes three dynamic thermal models which can be used to complement the FEA-based thermal model from this chapter and to implement it in circuit simulators.

Optimization of magnetic components using FEA-based thermal models

This chapter illustrates the impact of thermal modelling on the optimization of magnetic components, which is a crucial application in power electronics. The proposed models from Chapter 2 and 3 are used to develop a Finite Element Analysis based optimization. The first one is used to speed-up the required simulations, while the second one is used to characterize the thermal behavior of each core family.

This proposed FEA-based optimization is applied to a 200 W classic Forward converter and compared to a traditional analytic-based optimization, based on a commonly used thermal equation. It will be seen that the analytic approach leads to a non-compliant design that exceeds the established thermal limits. On the other hand, the proposed method provides the smallest design as possible while a safe operation within the thermal specifications is ensured.

Figure 4.1: Generic optimization set and Pareto front.

4.1 Multi-objective optimization of magnetic components

Efficiency-power density ($\eta - \rho$) Pareto-front is a recurrent tool to optimize power electronics converters [81]. In this chapter, power losses-volume Pareto will be analyzed, which is equivalent to the efficiency-power density Pareto. Losses, volume and temperature are the three main variables that affect the results obtained by the Pareto optimization. Accordingly, thermal modelling can be decisive to get a proper optimization. Using extensive approximations would lead to a Pareto in which the solutions might deviate from reality. Some interesting optimizations can be found in [7], [82], [83]. However, simplified thermal networks are used, with the inherent drawback of assuming certain approximations to model the interaction between the core and the winding, the simplification of the film coefficients for convection and radiation, and the assumption of concentrating the power losses at certain points (instead of assuming more realistic distributed power losses). On the other hand, the resulting designs are more representative and closer to reality if an accurate thermal model is used instead.

This chapter consists of a practical application of the developed models from the previous chapters, highlighting the impact of the main contributions of this thesis. To do so, a Pareto optimization is described for magnetic components, at device level. A comparison between the traditional equation-based Pareto and the proposed FEA-based Pareto is shown in this chapter. The proposed optimization uses the winding homogenization from Chapter 2 and the FEA model extraction from Chapter 3. An additional key-point of this procedure is performing just a few homogenized 3D FEA thermal simulations for each ferrite core family while using some scaling laws to interpolate other design solutions. This contribution is explained in detail along this chapter.

The optimization software for magnetics ANSYS PExprt [84] is used in this chapter for both optimizations. By means of this software, it is possible to get analytical solutions. Then, the proposed models will be used to complete a FEA-based optimization, which will be compared with the analytical-based results.

Multiple variables affect the design of magnetic components, starting from the voltage, current and frequency given by the power converter in which they are operating, as well as the number of turns, the core size and its magnetic properties. As a consequence of these parameters, certain power losses (P_{loss}) are produced in the magnetic component which, in turn, provoke certain temperature rise (ΔT). Both the power losses and the volume (Vol) are to be minimized in the optimization process.

These variables usually follow an opposite tendency, leading to a complex set of infinite solutions. However, when applying physical limits or industrial standards and specifications, the number of possible solutions is reduced. Some examples of these

limitations are the maximum operating temperature, the saturation of the core magnetic materials and the cost, which will be discussed later (Section 4.2).

After applying the previous limits, the Pareto front represents the boundary of the optimum solutions [85], as shown in Figure 4.1.

Pareto front technique is often used at system level to optimize a whole power converter. However, a Pareto front is applied at component level to optimize a **Forward transformer**, since the focus of this work is the modelling of magnetic components.

The accuracy on the calculation of intermediate variables, as power losses or temperature, strongly affects the quality of the Pareto-front. The higher the accuracy, the closer the performance of the optimum designs to the reality. In an opposite case, a poor accuracy of the used models could lead to designs whose performance is worse than expected or even out of limits. This would be discovered when testing a prototype, but not during the optimization process. This situation is undesirable, since a redesign could be required, as well as more prototyping, which increases the cost of the process. To avoid this situation, a FEA-based optimization is proposed and explained later in this chapter.

The main goal of this study is to show the impact of the thermal modelling on the quality of the Pareto front optimized solutions. So, two different Pareto fronts are compared at the end of the chapter. Both use the same input of power losses, but their difference lays on the used thermal model. The first Pareto front is based on a traditionally used simple equation, while the second Pareto front is computed by using the thermal models from Chapter 2 and 3, as well as certain proposed scaling laws, discussed later.

4.2 Electromagnetic, thermal and other specification limits

In order to narrow the numerous designs provided by the optimization, some constraints and trade-offs must be applied. They are related to the power converter specifications, the thermal limits, available core families and materials, volume and cost, described next.

4.2.1 Power converter specifications

The power converter specifications are the first considerations when designing a magnetic component, since they set the electrical operating conditions. These variables (voltage, current and frequency) are fixed in the present study for the sake of simplicity, assuming that all designs operate under the same electrical conditions imposed by the power converter. Considering the switching frequency as a variable to sweep in the optimization would imply considering other devices in the process too, like the semiconductor switches. However, the focus is optimizing the magnetic component at device level.

Table 4.1 summarizes these variables for the optimization of the classic Forward transformer treated in this chapter.

Table 4.1: Classic Forward transformer specifications for the optimization.

V_{in}	V_{out}	P_{out}	Δi_{out}	f_{sw}	$n_p : n_s : n_{aux}$
30 V	45 V	200 W	40 %	100 kHz	1 : 3 : 1

These electrical variables affect the selection of the magnetic core, the number of turns and wire diameter. Furthermore, they also influence the magnetic flux density and the power losses of the transformer.

Magnetic flux density

Regarding the magnetic flux density, B , every magnetic core material has a saturation level, B_{sat} , which is also dependent on the temperature. The maximum magnetic flux density under operation, B_{max} , is usually limited to certain percentage of B_{sat} , as a safe margin to avoid the saturation of the material. It is important to highlight that B_{sat} also depends on the temperature. In this case, the maximum operating flux is constrained to $B_{max} < 0.6 \cdot B_{sat}$, and B_{sat} is taken at 100°C as a worst case. This limitation can be found in some industrial standards, like [70], [74].

Power losses

As a consequence of the power losses, certain temperature rise occurs in the component. The cooling conditions (convection and radiation) and the geometry of the transformer affect the magnitude of that temperature increment. Following the same reasoning, the higher the volume of the magnetic component, the better the heat dissipation capability since the heat exchange surface with the ambient increases as well. Finally, a higher volume also induces to higher cost. These aspects are discussed next.

4.2.2 Temperature limits

One of the main limitation of magnetic components is the operating temperature. The first limitation that must be considered is the physical temperature rating of the materials. Most industrial standards describe more constraint limits, to ensure a safe margin, like the European standards for space applications [57], [70]–[72], the American military standards [73], [74] or the SAE standards for electrical vehicles and wireless chargers [75]. The established thermal limitations used in this work are discussed next.

Temperature rating of common materials

The temperature rating of each part of the magnetics and the corresponding materials are summarized next:

- Magnetic core: soft ferrites are generally suitable for high-frequency applications. The *Curie temperature* of these materials is around 200°C, from which they start losing their magnetic properties and their permeability starts decreasing sharply, tending to zero. Some materials have permeability less dependent on temperature than others. But, as a general rule, soft ferrites generally have a lower dependence for temperatures lower than 120°C [16]. In this study, a limit of 100°C is considered for soft ferrites, according to some industrial standards.
- Copper: copper conductors are able to withstand high temperatures [17], but the operating temperature is usually more limited by the insulation of the conductors. In [86], it is stated that HC copper suffers softening effect for temperatures above 150°C and that a successful operation for 20 to 25 years is ensured, supported by experimental data. Also, copper is able to withstand short-circuit conditions up to 250°C for a few seconds. In this study, the limit of the winding temperature is set by the wire insulation.
- Wire insulation: insulation materials are usually more critical. The most common material are enamel, polyamide, polyvinyl formal and polyurethane among others [42]. Electrical wires are usually classified in temperature classes: 105°C, 130°C, 155°C, 180°C and 200°C. For safety reasons, the maximum winding temperature is set as 105°C in this study.
- Soldering material: due to the thermal cycling, consequence of the pulsating current on the component, some mechanical failures like cracks can appear in the soldering materials [87]. The higher the temperature variation, the lower the lifetime until failure occurs. This is usually restricted by specifications but, as a general rule and in this study, it will be limited to 100°C.
- Impregnation material: some applications require embedding the winding in a *potting* or *impregnation* material. Some examples are CV-2500 [88], with maximum operating temperature of 300°C, or solithane [72], with 120°C. In this work, the winding is not potted.
- Coil former: the winding of a magnetic component is usually wound on a plastic coilformer. One of the most common materials is the high-grade liquid-crystal

polymer, which is able to withstand soldering temperatures as high as 350°C during manufacturing, but maximum operating temperatures of 180°C [89]. Since the previous materials are more restrictive, the thermal limits for the optimization are restricted by them, not by the coil former.

- Pin connectors: the metallic pin connectors are usually made of CuZn39Pb3 [90]. This material is able to work without mechanical failure below 250°C, so it does not mark an actual limit on the optimization.

It is important to highlight that the current density is usually limited by industry –as the inherited $5A/mm^2$ for the space sector–, but the actual physical limit to ensure a safe operation is the temperature. This is the reason why this study is focused on the thermal limitation. Some discussions regarding this aspects can be found in [91] for planar magnetics.

There is a temperature dependence of the electromagnetic and the thermal properties of any material, as discussed in Section 1.1. For simplicity, these properties are considered as constant around the operating temperature.

Influence of ambient temperature

The ambient temperature is crucial in every design. It limits the maximum allowable temperature rise of every component. Accordingly, the ambient temperature also impacts the electromagnetic properties of the materials. Hence, the higher the ambient temperature, the lower the flux density saturation. In this chapter, the magnetic flux saturation level is taken at 100°C as the worst case.

Furthermore, the boundary conditions determined by the cooling systems must be determined, since they strongly influence the temperature rise of the components. In this chapter, natural convection is assumed in order to verify the scope and applicability of the FEA model extraction from Chapter 3 under non-linear thermal boundary conditions. In this way, the applicability to other cooling methods like forced convection, which is practically linear, is verified too.

Further details of the environmental conditions and the thermal management of planar magnetics for space applications are analyzed in [92]. Deeper analysis and design guidelines for high-frequency high-power transformers are addressed in [93], as a minor contribution of this thesis.

4.2.3 Core families, materials and wires

In order to focus the scope of the work, just two different families of cores are analyzed: EE cores and pot cores [16]. These families are chosen due to their geometrical differences. The winding is more exposed to the ambient in EE cores than pot cores as they are ‘open’ core geometries, while pot cores have most parts of the winding enclosed between the two core halves.

The most suitable materials for the chosen operating frequency (Table 4.1) according to [16] have been included in the optimization: 3C90, 3C92, 3C94. The same recommendations are followed to select the wire section, considering high-frequency effects: AWG15, AWG20, AWG25.

4.2.4 Volume

The volume is often limited by specifications, due to mechanical limitations. The area and the height are usually specified in order to fit the electrical equipment in certain available space. However, no industrial specifications regarding volume are set here since the target of this study consists in getting the smallest solution as possible for the established requirements and comparing the results with traditional methods.

4.2.5 Cost

The variable of cost has been neglected for this study since it is tightly related to the volume of the magnetic component in a proportional way. The bigger the volume, the higher the cost, but the minimization of cost is not the target of the present study and is not analyzed.

4.3 Volume calculation of the magnetics

The effective volume is usually provided in the core datasheets, but it does not include the winding by definition. As a practical magnitude to evaluate the size of a magnetic component as a part of a power converter, the *box volume* concept was introduced by [6], and it is illustrated in Figure 4.2.

The expression of the box volume for transformers is presented in (4.1).

$$V_{box} = K_{VT} \cdot A_p^{3/4} \quad (4.1)$$

Where A_p is the area product, which is equal to the product of the window area and the core cross section area [94], according to (4.2).

Figure 4.2: Concept of transformer box volume.

$$A_p = W_a \cdot A_e \quad (4.2)$$

And K_{VT} is a constant dependent on the core size. There is a value of K_{VT} for each core family: $K_{VT} = 19$ for pot cores, calculated in [6], and $K_{VT} = 10$ for EE cores. In [6], the value for EE cores is 24, assuming that the winding occupies higher volume than in the present study. So, $K_{VT} = 10$ has been calculated as more appropriate for this optimization..

This concept was used in [6] to study the volume and weight of magnetic components as a function of frequency, and in [81] to optimize LC output filters of AC power sources. Notice that a very simplified expression for the thermal resistance of the components first proposed in [24], was used in those studies. It is also used in the analytic-based optimization of this chapter in order to compare it with the proposed FEA-based approach.

4.3.1 Cooling system volume and thermal management elements

In this work, magnetic components are analyzed in conditions of natural convection. So, no additional cooling system is included in the optimization. Nevertheless, some research regarding the impact of forced convection in the converter power density can be found in [95]. Further details regarding advanced thermal management techniques were analyzed in [92] for planar transformers for space applications. Nevertheless, the same conclusions can be stated for conventional magnetics and other industrial environments.

4.4 Power losses estimation

In magnetic components, the winding and the magnetic core suffer power losses during operation, mainly due to high-frequency effects. Both are described next.

Since the main goal of this chapter is analyzing the impact of thermal modelling on the Pareto-front results, the same power losses calculation is used in the classic and the FEA-based optimizations. Analytical approaches are used to calculate the power losses for the sake of simplicity.

Regarding the intermediate variable of power losses, P_{loss} , high frequency effects in the core and the winding are considered. For this step, the analytical computation of results from ANSYS PExprt is used [84].

For simplicity, only a one-way coupling between the EM and the thermal models is considered (see Section 1.4 for more details). In other words, the physical properties of the materials are considered constant at the worst-case temperature, hence the power losses too. If considering an EM-thermal feedback (two-way coupling) and recalculating the physical properties for every iteration in the optimization, both Pareto fronts would move in the same manner. The results using this feedback would be slightly more accurate, but this is computationally expensive and it is not the goal of this study.

The underlying equations for the core losses and the winding losses are described next.

4.4.1 Core losses

Ferromagnetic cores suffer certain power losses due to their inherent properties and the interaction with a current flowing through the winding. The causes of these core losses are briefly described next:

- **Hysteresis:** due to the magnetization and the demagnetization of the ferromagnetic core, a hysteresis loop is created in each cycle of operation. The energy required for this magnetization cycle is lost as heat. The research on this phenomena dates back to 1892 and the research of Charles Proteus Steinmetz [77]. Further studies can be found in later bibliography, like [96], as well as a measurement setup [97] to characterize this effect.
- **Eddy currents:** Ferromagnetic materials are also electrical conductors, so they have certain electrical resistivity. Due to the magnetic field created by the flowing current of the winding, certain currents are induced in the domains of the magnetic core, named *Eddy Currents*. This circulating currents generate certain power losses due to the domains wall motion, which are dissipated as heat. Some overview on this effect was given in [98]. This effect is more patent in metallic materials than in ferrites.
- **DC bias:** if the magnetic component operates with a DC premagnetization level -in other words, if the magnetic inductance current has a DC offset-, some losses are

added to the hysteresis. This happens in topologies like the Flyback converter, in which the transformer only works in the positive quadrant of the B-H loop, but not in other topologies like the Push-Pull. Further insight of this effect can be found in [99].

- **Relaxation effect:** it was proven in [100] and [101] that there are some losses during the phase of constant flux, when the voltage across the magnetic component is zero. This effect has more influence in topologies with high dead-times ($t_{dead} > 1\% \cdot T$), like Half-Bridge or Full-Bridge converters.

Several equations can be found in the literature to model the overall core losses:

- **SE (Steinmetz Equation):** this equation accurately estimates the core losses for sinusoidal excitation. It depends on three coefficients, usually provided by the manufacturers in the material datasheet.
- **MSE (Modified Steinmetz Equation):** this was a pioneer attempt to model core losses for non-sinusoidal waveforms and DC-bias, but internal inconsistencies were found [99], [102].
- **GSE (General Steinmetz Equation):** in this formulation, also losses for non-sinusoidal waveforms are modelled, but not DC-bias. However, the results do not correlate consistently with measurements [103].
- **iGSE (improved General Steinmetz Equation):** this equation is consistent and matches with measurements [78]; also for non-sinusoidal excitation. But losses due to relaxation and DC bias are neglected.
- **i2GSE (improved-improved General Steinmetz Equation):** this is a modification of the iGSE to include the relaxation and DC bias losses [100]. However, it requires experimental data to get the corresponding parameters of the equation.

For this study, iGSE is chosen since no experimental data is required to use the equation and it is proven accurate. The expression is next:

$$P_v = \frac{1}{T} \int_0^T \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt \quad (4.3)$$

$$k_i = \frac{k_{fe}}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos\theta|^\alpha 2^{\beta-\alpha} d\theta} \quad (4.4)$$

Being ΔB the peak-to-peak flux density. k_{fe} , α and β are the coefficients from Steinmetz Equation.

Core losses due to relaxation and DC are neglected in this study, since they have a low contribution on the total core losses in the case of a classic Forward converter.

As mentioned above, power losses are considered constant with temperature for simplicity. Some manufacturers provide some coefficients to add a quadratic dependence with temperature. However, this is a simplification only valid under certain conditions. Further information on this topic and a proposed set of equations to accurately model the temperature dependence of core losses using the information provided by manufacturers can be found in Appendix C, as a minor contribution of this thesis.

4.4.2 Winding losses

As a consequence of the high-frequency operation of switching converters, some effects appear in the conductors of the magnetic components, described next:

- **Skin:** as a result of opposed eddy currents induced by the changing magnetic field inherent to the pulsed operation, the current density tends to concentrate near to the external surface. Therefore, the effective cross-section of the conductors is reduced, hence increasing the effective resistance and thus the losses too [47].
- **Proximity:** when two conductors are adjacent and both carry alternating (or pulsating) current, they create certain magnetic flux which correspondingly induces eddy currents in each other. As a consequence, the current density tends to concentrate on the opposing outer surfaces of both conductors. Correspondingly, the effective cross-section of both conductors is reduced, hence increasing the effective resistance and the losses too [48].
- **Gap proximity:** a similar effect occurs when the air-gap of a magnetic core is located close to a conductor. The magnetic field induces certain eddy current in the conductor, increasing the effective resistance [49], [50]. However, this effect only appears if the magnetic core has an air-gap, like inductors or Flyback transformers.

The previous effects are aggravated with higher frequency operation.

The most common approach to compute the total winding losses is calculating the Fourier transform of the current flowing through each winding of the magnetic component. In this way, the current can be represented in the frequency-domain, with one harmonic per frequency. Then, the resistance of the winding, R_{ac} , can be conveniently calculated

as a function of frequency. It is possible to calculate the winding losses by means of the information of the harmonics of the current and the R_{ac} [80].

$$P_{wind} = R_{DC} \cdot I_{DC}^2 + \sum_{harmonic} \left(\sum_{i=wire} \left(\sum_{j=wire} R_{ac_{ij}} \cdot I_{rms_i} \cdot I_{rms_j} \right) \right) \quad (4.5)$$

The terms $R_{ac_{ij}}$ are the AC resistance of the impedance matrix from [80]. The diagonal terms ($i = j$) represent the skin effect due to the current flowing in that conductor. The other terms represent the proximity effect, caused by the interaction between conductors.

Different approaches can be used to obtain the value of the R_{ac} for each frequency. There are some geometrical simplifications, like [48] and [104], whose accuracy was proven accurate for high filling factor of the winding and foil conductors. There are also some semi-empirical approaches, like [105], which are equations based in EM FEA results, more accurate than the previous models. However, the most accurate way is obtaining the R_{ac} values directly from FEA simulations, with software like ANSYS PEmag [79].

Furthermore, there is some influence of the temperature on the winding losses, as studied in [18], [19], due to the temperature dependent resistivity of the conductors.

In this study, the implemented analytical approaches implemented in ANSYS PExprt [84] are used for the winding losses estimation. The temperature dependence is neglected for simplicity, as previously explained.

4.5 Classic optimization using analytical approaches

In this section, the analytical-based optimization is discussed. The software optimization tool ANSYS PExprt [84] is used for this purpose, since it has implemented several analytical equations.

In this software, first the type of magnetic component and the power converter topology are chosen. In this case, a transformer for a classic DC/DC Forward is selected. Then, the specifications described in Section 4.2 are the input to start the optimization. Figures 4.3 and 4.4 are some screenshots of the PExprt settings, as an example.

Then, different methods to compute the losses can be set. To use the described methods in Section 4.4, the ‘Harmonics and AC Resistance (Dowell)’ and ‘Steinmetz (modified)’ options are chosen to compute the winding and core loss, respectively (Figure 4.5). It is important to keep in mind that the two optimization processes treated in this chapter, share the same values of power losses. The only difference between them is the used thermal model to calculate the temperature rise.

4.5. Classic optimization using analytical approaches

Figure 4.3: Electrical specifications settings in PExprt.

Figure 4.4: Operating conditions settings in PExprt.

The **thermal model in this analytical-based** optimization that provides the estimation of the temperature for the chosen specifications is based on the commonly used simple equation from [24]:

$$R_{th} \left[\frac{K}{W} \right] = 53 \cdot V_e [cm^3]^{-0.54} \quad (4.6)$$

No Finite Element Simulations are used in this process.

At this point, the iterative analytical-based optimization is ready to be initiated. Then, all possible designs that meet the specifications are collected, as in Figure 4.6.

The Pareto front obtained by this analytic optimization is compared to the FEA-based optimization in Section 4.7.

Figure 4.5: Calculation methods settings in PExprt.

Core [Name]	Wire [...]	Materi...	Volum...	Heigh...	Footp...	Turns	T...	Magn...	Parall...	Powe...	Wind...	I
P22/13	AWG25	Ferrox...	1997....	13.40	380.13	12	291.72	613.6...	1	7.7815	42.86	
E25/9/6	AWG20	Ferrox...	1820....	18.90	160.02	19	260.75	680.2...	1	7.6458	46.25	!
E25/9/6	AWG20	Ferrox...	1820....	18.90	160.02	19	259.90	680.2...	1	7.6198	46.25	!
E25/10/6	AWG20	Ferrox...	1935....	19.30	161.29	19	256.07	681.2...	1	7.6604	48.23	!
E25/9/6	AWG25	Ferrox...	1820....	18.90	160.02	19	255.68	680.2...	2	7.2498	59.08	!
E25/10/6	AWG20	Ferrox...	1935....	19.30	161.29	19	255.27	681.2...	1	7.6354	48.23	!
E25/9/6	AWG25	Ferrox...	1820....	18.90	160.02	19	254.82	680.2...	2	7.2238	59.08	!
E25/10/6	AWG25	Ferrox...	1935....	19.30	161.29	19	251.47	681.2...	2	7.2648	61.61	!
E25/10/6	AWG25	Ferrox...	1935....	19.30	161.29	19	250.66	681.2...	2	7.2397	61.61	!

Figure 4.6: Results sample of PExprt.

4.6 Proposed optimization based on 3D FEA thermal simulations

In this section, a new optimization using the FEA-based methods from the previous chapters is proposed. It will be seen that this approach provides the smallest possible design while ensuring a safe operation within the established thermal limits. On the other hand, the classical analytic-based optimization leads to a design whose temperature exceeds the thermal limits by 30°C, as explained later in Section 4.7. This undesirable situation is avoided by using the proposed FEA-based optimization.

Taking as a starting point the results from the classic analytical-based optimization, the temperature estimation for each design is recalculated by means of the methods and models developed in this work. The step-by-step procedure is described next:

1. Determine all candidate designs

First, the power losses of every combination of cores and number of turns is computed. So, the collected results by the analytical-based approach are the starting point for this

FEA-based process, since both share the same calculated power losses. The difference lays on the temperature estimation, explained next.

2. Choose three intermediate designs per family and calculate k_{eq}

In order to characterize the thermal behaviour of a core family, just three intermediate designs are chosen to be simulated. As a general rule, designs with a high filling factor show better performance; otherwise, the magnetic component would be either over-sized or would have excessive losses. Then, the homogenized k_{eq} (Chapter 2) is calculated for those intermediate designs with the aim of creating an accurate but fast thermal simulation.

3. Obtain the matrix $[R_{th_{max}}]$ for the chosen designs on the previous step

Taking advantage of the homogenization performed in the previous step, the matrix $[R_{th_{max}}]$ (Chapter 3) is created for the selected designs. In this way, their temperature rise can be easily obtained for any value of core and winding losses. This is particularly convenient since it can be assumed that every design with the same core and a similar filling factor is characterized by the same $[R_{th_{max}}]$. This can only be assumed if the filling factor is similar. However, since the only designs with high filling factor were chosen in the previous step, this is not a limitation.

4. Scaling laws to characterize the thermal behaviour each core family

By means of the previously obtained $[R_{th_{max}}]$ matrix for the chosen designs, an interpolation of each term R_{ij} of the matrix is performed to characterize the thermal behaviour of the whole core family. This ‘scaling laws’ of the thermal matrix were proposed in collaboration with Y. Bouvier and applied in his PhD thesis to planar transformers [106].

This concept is illustrated in Figure 4.7, where the thermal matrix is obtained for the family of wound EE cores used in this study. Therefore, the interpolation of the R_{ij} terms for the EE core family is described according to the next expressions, as a function of the effective volume, V_e :

$$R_{th_{11}} \left[\frac{K}{W} \right] = 41.44 \cdot e^{-2.465 \cdot 10^5 \cdot V_e [m^3]} + 13.43; \quad (4.7)$$

$$R_{th_{12}} \left[\frac{K}{W} \right] = 38.29 \cdot e^{-3.23 \cdot 10^5 \cdot V_e [m^3]} + 17.36; \quad (4.8)$$

$$R_{th_{21}} \left[\frac{K}{W} \right] = 46.3 \cdot e^{-3.165 \cdot 10^5 \cdot V_e [m^3]} + 13.42; \quad (4.9)$$

$$R_{th22} \left[\frac{K}{W} \right] = 49.28 \cdot e^{-6.555 \cdot 10^4 \cdot V_e [m^3]} + 3.247; \quad (4.10)$$

The same approach is used to characterize the pot core family. Figure 4.8 shows the scaling laws of the R_{thij} terms for the pot core family that is included in the optimization. This behaviour of the R_{ij} terms for the pot core family as a function of the effective volume, V_e , can be expressed as:

$$R_{th11} \left[\frac{K}{W} \right] = 30.7 \cdot e^{-2.065 \cdot 10^5 \cdot V_e [m^3]} + 11.11; \quad (4.11)$$

$$R_{th12} \left[\frac{K}{W} \right] = 34.13 \cdot e^{-2.115 \cdot 10^5 \cdot V_e [m^3]} + 15.95; \quad (4.12)$$

$$R_{th21} \left[\frac{K}{W} \right] = 29.95 \cdot e^{-1.913 \cdot 10^5 \cdot V_e [m^3]} + 13.42; \quad (4.13)$$

$$R_{th22} \left[\frac{K}{W} \right] = 90.37 \cdot e^{-3.867 \cdot 10^4 \cdot V_e [m^3]} + 31.02; \quad (4.14)$$

It is important to mention that the fitting of the R_{ij} terms is only valid for the range of core sizes analyzed in this study. Also, the higher the number of simulated designs, the higher the accuracy of the temperature estimation; but that is a trade-off between the accuracy of the method and the required computational time. In this case, only three designs are characterized in order to validate the proposed method for just a reduced number of simulations.

Figure 4.7: Scaling laws of the thermal resistances R_{ij} for EE cores over effective volume (obtained by FEA in diamonds, obtained by curve fitting in dotted lines).

Figure 4.8: Scaling laws of the thermal resistances R_{ij} for pot cores over effective volume (obtained by FEA in diamonds, obtained by curve fitting in dotted lines).

5. Calculate the temperature rise of each design

Once every core family is characterized, the temperature rise of each design for the corresponding power losses is calculated.

The resulting Pareto obtained by this proposed FEA-based optimization is analyzed and compared to the classic analytical approach in the next section.

4.7 Validation

Along this section, the obtained solutions by the classic analytical-based optimization and the proposed FEA-based optimization are discussed and compared.

4.7.1 Comparison: equation-based vs. FEA-based

In both cases, the main goal of the optimization is obtaining the smallest achievable volume for the established specifications. The optimization results are depicted in Figure 4.9. The power losses of every analyzed design over their volume are represented. The maximum temperature is represented by the color scale, in $^{\circ}\text{C}$. Each dot represents a valid design, with different core, wire section, number of turns or number of parallel conductors. The black crosses are the thermally restricted discarded designs, according to each method; their temperature overcome the established limits.

It can be seen that the main difference is that a higher number of designs are discarded by the FEA-based method, compared to the analytic-based. This is because the analytic method over-estimates the temperature for every design, for this particular specifications. However, the FEA-based method provides a more realistic estimation of the temperatures, as shown later.

As a result of this fact, the optimum design according to the analytic method is discarded by the FEA-based, since it exceeds the thermal limits. **This dangerous situation is effectively avoided by the proposed FEA-based method**, whose optimum design has a slightly higher volume, but it is capable of dissipating the generated power losses, **keeping it within safe thermal limits**.

4.7.2 3D FEA simulation of the optimum designs

Table 4.2 summarizes the characteristics of the optimum designs obtained by both methods. It is important to highlight that the suitable solution with a P26/16 core was discarded by the analytical-based optimization, since its temperature estimations are not as accurate as the proposed FEA-based. This will be analyzed next.

Figure 4.9: Optimization results comparison: analytical-based versus proposed FEA-based. The Pareto front is defined with the pink line. The colour scale indicates the maximum temperature rise of each component. The discarded designs due to a temperature rise over specifications are identified by black crosses, while the smallest optimum design is highlighted with a yellow star. The box volume of these smallest designs is indicated. It can be seen that the smallest design according to the analytical optimization is discarded by the FEA-based optimization, whose temperature estimation is more reliable and accurate, for exceeding the maximum allowed temperature.

4.7. Validation

Table 4.2: Optimum designs obtained by the analytical and the proposed FEA-based optimizations.

	Core	Material	Turns	Magnetizing inductance	Total power losses	Window Filling	Temperature estimation	Box volume
Analytical based	P26/16	3C92	7:21:7	192 μ H	1.8 W	47.2%	75.7°C	10cm ³
FEA based	E30/15/7	3C92	11:33:11	183 μ H	1.9 W	55.6%	75.6°C	11.4cm ³

In order to validate the proposed FEA-based optimization, a 3D FEA simulation is performed in order to analyze and compare the solution of both the classic and the proposed optimizations, depicted in Figure 4.10.

Figure 4.10: 3D representation of the simulated optimum designs with the same scale, and the obtained temperatures by simulation: a) analytical optimization (left), b) FEA-based optimization (right).

The temperature estimation of both estimations is summarized and compared in Table 4.3 with 3D FEA thermal simulations. It can be clearly seen that the analytical-based optimization has a considerable deviation from the actual temperature of the component (-38.1% , 34.3°C difference), which leads to a design not compliant with the thermal limits ($110^\circ\text{C} > T_{slim}$). However, by using the proposed FEA-based optimization, the **temperature estimation has just a -11.7% deviation (7.4°C difference)** compared to the corresponding 3D simulation.

By using the classic analytic-based optimization, the compliance with specifications is not ensured, leading to an undesirable uncertainty. Furthermore, that analytic method does not necessarily lead to an under-estimation of the maximum temperature like in the example of this chapter. The opposite situation might occur, thus leading to over-sized designs. Both situations are undesirable and the boundary between them is unpredictable. However, they can be avoided by using the **proposed method, which leads to the smallest designs as possible while complying with the thermal limits.**

Table 4.3: Temperature estimation by the optimizations compared to 3D FEA thermal simulations.

$T_{amb} = 20^{\circ}\text{C}$	Maximum temperature estimated in the optimization (relative error)	Maximum temperature obtained by a 3D FEA simulation
Analytical based	75.7°C (−38.1%)	110°C
FEA based	75.6°C (−11.7%)	83°C

4.8 Chapter summary and contributions

In this chapter, the proposed models from Chapters 2 and 3 are used in an optimization of a high-frequency magnetic as demonstration of their usefulness in a typical application in the field of power electronics.

This **proposed Finite Element Analysis based optimization** uses the developed k_{eq} **homogenization from Chapter 2 to speed-up the required 3D simulations**. By means of the **model extraction from FEA results developed in Chapter 3**, a thermal model is generated for each simulated component, $[R_{th_{max}}]$. To avoid the extensive use of 3D simulations, just three intermediate designs from each core family are chosen to be simulated. Then, by means of certain scaling laws, the behaviour of the whole core family is obtained by interpolating the $[R_{th_{max}}]$ matrices of the simulated designs.

This optimization method is applied to a 200 W classic Forward converter and compared to a traditional analytic-based optimization, based on a commonly used thermal equation. It has been demonstrated that the analytic method leads to a design which exceeds the thermal limits (34.3°C over the limit) in this particular example. This undesirable situation is avoided by using the proposed FEA-based optimization, which leads to the smallest design as possible while ensuring the compliance with specifications. In this particular example, the **temperature estimation of the FEA-based method is only 7.4°C below the actual temperature** of the component, obtained by a 3D FEA simulations as a verification. **The proposed method allows pushing the power density to its limits without exceeding the established thermal limits.**

Contributions related to this chapter

Conference papers:

1. [92] **G. Salinas**, J. A. Oliver, J. Muñoz-Antón, A. Fernández and R. Prieto, “Thermal Management of Planar Magnetics for Spacecrafts: characterization and modeling,” 2019 European Space Power Conference (**ESPC**), Jean-Les-Pins, France,

2019.

doi: 10.1109/ESPC.2019.8932068

url: <https://ieeexplore.ieee.org/document/8932068>

2. [93] **G. Salinas**, Á. Giménez, J. A. Oliver and R. Prieto, “Design Guidelines for High-Power and High-Frequency Transformers,” 2019 IEEE Applied Power Electronics Conference and Exposition (**APEC**), Anaheim, CA, USA, 2019, pp. 1423-1429.

doi: 10.1109/APEC.2019.8722060

url: <https://ieeexplore.ieee.org/document/8722060>

Conclusions and future work

5.1 Conclusions and contributions

Thermal management is one of the main bottlenecks for reducing the size of power converters by increasing their power-density. Hence, accurate thermal models that properly predict the thermal gradient of the components on a power converter during operation are key to obtain better-optimized designs. Since magnetic components –inductors and transformers– are the bulkiest components in power converters and lack versatile and accurate models, this thesis is focused on the thermal modelling of these devices.

Finite Element Analysis tools are usually used at the last stage of the development of a product, prior to the physical prototyping phase, in order to check the performance of any component of a power converter. If the limits are exceeded, a redesign of the component is required, leading to another iteration of the design of the product. Furthermore, these tools are considered the most accurate. As a counterpart, they require some expertise by the user and are considerably time-consuming.

A different philosophy is adopted in this work: taking advantage of the potential of FEA tools to develop thermal models of magnetic components based on their results.

The **first main contribution** is a thermal homogenization, in which every conductor from a winding is replaced by a single block (Chapter 2). The effective properties of that block are obtained by means of FEA simulations. As a consequence of this homogenization, the required time of a 3D thermal simulation time can be reduced by a factor of 8, without penalizing the accuracy on the temperature estimation.

The **second main contribution** of this work is the development of a generic methodology to obtain a thermal coefficients matrix which can be used to predict the maximum temperature and average temperature at any region of a magnetic component (Chapter 3). This matrix model can be used to predict the temperature for any combination of power losses (in the core and each one of the windings), given by the electrical operation. An additional feature is that the matrix can be obtained for any boundary conditions, regardless of its complexity, since it is built by means of the results

of a set of FEA thermal simulations. This coefficients matrix is a static model; so, the coupling with a dynamic thermal model is proposed to make it suitable for the implementation in any circuit simulator, which is a **minor contribution** of this thesis. In this way, the electro-thermal behaviour of the inductor or transformer can be inspected while analysing the operation of the power converter.

Finally, the previous thermal models have been used to propose a FEA-based optimization of magnetic components, leading to the **last main contribution** of the thesis (Chapter 4). The thermal homogenization from Chapter 2 has been used to speed-up the required 3D simulations. The model extraction methodology from Chapter 3 is used to build a matrix to characterize the components from those simulations. Finally, being a **minor contribution**, a set of scaling laws is proposed to interpolate the thermal behaviour of intermediate design solutions within the same core family. This approach was compared to a classic analytic-based optimization, using a 200 W Forward transformer as a case study. It was found that the classic approach leads to a design that exceeds the thermal limits. However, the proposed FEA-based optimization avoids this situation and achieves the smallest design as possible while ensuring a safe operation within the established specifications.

As **minor contributions** of this thesis, Appendix B discusses three different dynamic thermal models in order to make possible the implementation of the FEA-based model from Chapter 3 in circuit simulators; and Appendix C describes an accurate formulation of the dependence of core power losses on the flux density and the temperature, by means of the data provided by manufacturers.

5.2 Future work

The next tasks are proposed as future work to complement the research of this thesis dissertation:

- Develop a k_{eq} homogenization for planar magnetics: A homogenization of the insulation layers (FR4 and prepreg) is proposed as part of this research in [92]. However, a further step can be carried out, like in Chapter 2, where the conductors are also included in the homogenization. Obtaining this simplification might be useful to get fast estimations of the maximum temperature of planar transformers by means of a simple simulation. The use of thermal networks to make a homogenization layer by layer is suggested, similarly to Appendix A.
- Compensate the geometry change due to the homogenization: As explained in the experimental results from Chapter 2, there is a source of error due to the geometric

change after performing the proposed homogenization. For example, if every wire of a winding is replaced by a block, it can be seen that the film coefficient has changed since the heat exchange surface to ambient has been altered. The impact on the results is minor since the temperature rise within the solids due to the internal heat generation has a higher effect than the dissipation to ambient. Nevertheless, a correcting factor can be applied to k_{eq} in order to compensate for this effect.

- Obtain the dynamic thermal capacitance by means of transient thermal FEA simulations: The static coefficient matrix model from Chapter 3 was coupled with a simple approximation, the *lumped systems* assumption [20], [21], to provide the dynamic capability. However, due to the heterogeneity of each part of the magnetic component, this assumption might have a certain mismatch on the transient temperature estimation, despite the steady temperature is accurately predicted. To avoid this, it is possible to extract a *thermal capacitance* model from transient simulations.
- Using a two-way EM-Thermal coupling for the optimization: in the optimization from Chapter 4 a one-way coupling was used, so that the power losses were calculated for a single temperature (the highest allowed temperature as a worst-case), but not recalculated by a feedback of the temperature estimation. The resulting power losses might be lower than calculated in this chapter, possibly leading to smaller optimum volumes. Also, the electromagnetic and thermal properties of the materials were assumed constant (using worst-case values) and not dependent on temperature. These properties can be also recalculated if two-way feedback is implemented.
- Include the variable of cost in the optimization: The cost was neglected in the optimization proposed in Chapter 4, for the sake of simplicity, since it is directly related to volume (the lower the volume, the lower the cost). Nevertheless, volume and cost are not linearly related. As a consequence, the final choice of a solution from the Pareto front might differ if the cost is not contemplated.
- Consider other cooling systems: in the optimization, only natural convection was considered, since the aim was to validate the linearized model extraction from Chapter 3 for non-linear boundary conditions. However, extending this study to forced convection or any other advanced cooling system like heat pipes can be interesting. The models developed in this thesis are suitable for this extended study since the boundary conditions for forced convection or heat pipes behave in a linear manner and the models have been validated for more complex non-linear conditions.

- Use Electro-Magnetic (EM) FEA coupled with the thermal simulations in the optimization: Since the aim of Chapter 4 was to illustrate the impact of the accuracy of thermal models in the Pareto front, the same analytically calculated power losses were considered for both the classic Pareto and the proposed FEA-based optimization. Nevertheless, a step further can be carried out by including an EM simulation coupled to the thermal simulations with the same setup. In this way, the power losses can be obtained by means of FEA, more accurately than with the used analytic simplifications.

5.3 Publication of results

The results obtained in this research are compiled in the next journal publications and conference papers.

Journal publications

1. [67] **G. Salinas**, A. Delgado, J. Muñoz-Antón, J. A. Oliver and R. Prieto, “Fast and Accurate Thermal Modeling of Magnetic Components by FEA-based homogenization,” in *IEEE Transactions on Power Electronics (TPEL)*.
doi: 10.1109/TPEL.2019.2921160
url: <https://ieeexplore.ieee.org/document/8731691>
2. [55] A. Delgado, **G. Salinas**, J. A. Oliver, J. A. Cobos and J. Rodríguez-Moreno, “Equivalent Conductor Layer for Fast 3D Finite Element Simulations of Inductive Power Transfer Coils,” in *IEEE Transactions on Power Electronics (TPEL)*.
doi: 10.1109/TPEL.2019.2921160
url: <https://ieeexplore.ieee.org/document/8882243>

Conference papers

1. [68] **G. Salinas**, A. Delgado, J. A. Oliver and R. Prieto, “Fast FEA Thermal Simulation of Magnetic Components by Winding Equivalent Layers,” 2018 IEEE Energy Conversion Congress and Exposition (**ECCE**), Portland, OR, USA, 2018, pp. 7380-7385.
doi: 10.1109/ECCE.2018.8557714
url: <https://ieeexplore.ieee.org/document/8557714>
2. [92] **G. Salinas**, J. A. Oliver, J. Muñoz-Antón, A. Fernández and R. Prieto, “Thermal Management of Planar Magnetics for Spacecrafts: characterization and modeling,” 2019 European Space Power Conference (**ESPC**), Jean-Les-Pins,

France, 2019.

doi: 10.1109/ESPC.2019.8932068

url: <https://ieeexplore.ieee.org/document/8932068>

3. [93] **G. Salinas**, Á. Giménez, J. A. Oliver and R. Prieto, “Design Guidelines for High-Power and High-Frequency Transformers,” 2019 IEEE Applied Power Electronics Conference and Exposition (**APEC**), Anaheim, CA, USA, 2019, pp. 1423-1429.

doi: 10.1109/APEC.2019.8722060

url: <https://ieeexplore.ieee.org/document/8722060>

4. [54] A. Delgado, **G. Salinas**, J. A. Oliver, J. A. Cobos and J. Rodríguez, “Equivalent Parameters of Round and Litz Wire Conductors to Obtain an Equivalent Layer to Accelerate Finite Element Simulations of Wireless Power Transfer System,” 2018 IEEE Energy Conversion Congress and Exposition (**ECCE**), Portland, OR, USA, 2018, pp. 7375-7379.

doi: 10.1109/ECCE.2018.8558310

url: <https://ieeexplore.ieee.org/document/8558310>

5. [69] A. Delgado, **G. Salinas**, J. Rodríguez, J. A. Oliver and J. A. Cobos, “Finite Element Modelling of Litz Wire Conductors and Compound Magnetic Materials based on Magnetic Nano-particles by means of Equivalent Homogeneous Materials for Wireless Power Transfer System,” 2018 IEEE 19th Workshop on Control and Modeling for Power Electronics (**COMPEL**), Padua, Italy, 2018, pp. 1-5.

doi: 10.1109/COMPEL.2018.8460012

url: <https://ieeexplore.ieee.org/document/8460012>

Derivation of the effective thermal conductivity for foil winding

In this appendix, the derivation of the equations for the effective conductivity for rectangular and circular foil winding, used in Section 2.2.2, is explained.

A.1 XY rectangular winding

A.1.1 Orthogonal effective thermal conductivity

Considering a plane wall made by several layers of conductor material and other layers of insulator (Figure 2.14), the total thermal resistance in the orthogonal direction (X or Y) can be expressed as the equivalent resistance of a series association of resistances [20], [21], as:

$$R_{th_{ort}} = \frac{l_{cond}}{k_{cond} \cdot A} + \frac{l_{ins}}{k_{ins} \cdot A} \quad (A.1)$$

Where A is the cross-section, l_{cond} and l_{ins} are the total width of conductor and insulation regions, respectively. For Figure 2.14: considering the red layers as conductors and yellow ones as insulation, $l_{cond} = l_1 + l_3$ and $l_{ins} = l_2 + l_4$. k_{cond} and k_{ins} are their corresponding thermal conductivity.

Then, according to the definition of a thermal resistance, $R_{th_{ort}}$ can be expressed in terms of its effective conductivity:

$$R_{th_{ort}} = \frac{l_{tot}}{k_{eq_{ort}} \cdot A} \quad (A.2)$$

Rearranging the previous equation, the orthogonal conductivity $k_{eq_{ort}}$ can be computed as a function of the thermal conductivity and length of each material region:

$$k_{eq_{ort}} = \frac{(l_{cond} + l_{ins}) \cdot k_{cond} \cdot k_{ins}}{l_{cond} \cdot k_{ins} + l_{ins} \cdot k_{cond}} \quad (A.3)$$

A.1.2 Effective thermal conductivity in the vertical direction (Z)

To calculate the effective conductivity in the vertical direction, Z, the parallel association of the thermal resistance at each region is performed:

$$R_{th_z} = \left(\frac{1}{R_{th_{cond-Z}}} + \frac{1}{R_{th_{iso-Z}}} \right)^{-1} = \left(\frac{k_{cond} \cdot A_{cond}}{l_z} + \frac{k_{ins} \cdot A_{ins}}{l_z} \right)^{-1} \quad (A.4)$$

Then, defining R_{th_z} as a function of the effective conductivity:

$$R_{th_z} = \frac{l_z}{k_{eq_z} \cdot A_{tot_z}} \quad (\text{A.5})$$

Following the same concept than in the previous case, and assuming that all regions have the same vertical length (l_z), the effective conductivity in the vertical direction can be described as:

$$k_{eq_z} = \frac{k_{cond} \cdot A_{cond} + k_{ins} \cdot A_{ins}}{A_{cond} + A_{ins}} \quad (\text{A.6})$$

A.2 RZ circular winding

To define the effective thermal conductivity of a circular winding, the procedure is similar to the previous section, but using cylindrical coordinates (Figure 2.15).

A.2.1 Effective radial thermal conductivity

In the radial direction, the thermal resistance of each region n is defined as [20], [21]:

$$R_{th_{n-r}} = \frac{\ln(r_{n-out}/r_{n-in})}{2\pi \cdot l_z \cdot k_n} \quad (\text{A.7})$$

Being r_{n-out} and r_{n-in} are the outer and inner radius of the n layer, respectively; l_z is the height of the winding; and k_n is the thermal conductivity of the n layer.

At this point, analyzing the circular foil winding in the radial direction, the total radial thermal resistance is equal to the series association of each layer's thermal resistance:

$$R_{th_r} = \sum_n R_{th_{n-r}} = \frac{\ln(r_{out}/r_{in})}{2\pi \cdot l_z \cdot k_{eq_r}} \quad (\text{A.8})$$

Where r_{out} and r_{in} are the outer and inner radius of the whole winding ; k_{eq_r} is the radial effective thermal conductivity.

Therefore, by rearranging the previous equation, k_{eq_r} can be described in terms of the radius and conductivity of each layer:

$$k_{eq_r} = \frac{\ln(r_n/r_1) \cdot k_{cond} \cdot k_{ins}}{k_{ins} \sum_{n_{cond}} \ln(\frac{r_{n+1}}{r_n}) + k_{cond} \sum_{n_{ins}} \ln(\frac{r_{n+1}}{r_n})} \quad (\text{A.9})$$

A.2.2 Effective thermal conductivity in the vertical direction

Both rectangular and circular foil winding cases share the same expression for the effective conductivity in the vertical direction, (A.6).

Implementation of $[R_{th}]$ with dynamic thermal models for circuit simulators

Virtual prototyping is used to check the performance of electronic devices prior to the development of a physical product. This is the reason why checking the thermal behavior under real electrical operating conditions is so relevant. So, having a thermal model of the magnetics coupled with the electrical simulation of the whole circuit can be a useful tool for design engineers (see Chapter 1 for more details regarding the electro-thermal coupling). A Flyback converter schematic including an electro-thermal model of the transformer in ANSYS TwinBuilder circuit simulator is shown in Figure B.1

Figure B.1: ANSYS TwinBuilder schematic of a Flyback converter with a electro-thermal VHDL-AMS of the transformer.

Since the proposed FEA-based thermal model from Chapter 3 is static, the time dependence needs to be added in order to make it suitable for the implementation in circuit simulators. Taking advantage of its matrix form, the next solutions are propose to implement the dynamic thermal behaviour:

- Moving average.
- Diagonal capacitance matrix.
- Complete capacitance matrix.

The implementation of the static model in VHDL-AMS is discussed, prior to analyze the dynamic models. The choice of using VHDL-AMS is convenient to take advantage of the electromagnetic models already implemented in ANSYS PExprt and ANSYS PEmag, which can be imported in the circuit simulator ANSYS TwinBuilder [79].

B.1 Implementation of the static model

The matrix equation (3.9) can be written as split equation form, one for each object. Considering a system with two objects, the corresponding implementation in VHDL-AMS is:

```
// Output ports
PORT real out: TempMax1 = TempMax1
PORT real out: TempMax2 = TempMax2

// Component temperature per objects
EQU {TempMax1:= P1*R11 + P2*R12 + Tamb} DST: SIM(Type:=SSGM, Sequ:=PRE);
EQU {TempMax2:= P2*R22 + P1*R21 + Tamb} DST: SIM(Type:=SSGM, Sequ:=PRE);
```

In this implementation, the values of P_1 and P_2 correspond to the electromagnetic losses of objects 1 and 2. If a magnetic component is modelled in ANSYS TwinBuilder, these losses can be calculated by using the ANSYS PExprt or ANSYS PEmag models [79]. Otherwise, an electromagnetic model must be implemented. Some methods are described in [13] and [14].

B.2 Adding the dynamic behaviour

The previous implementation of the thermal model is static, so the calculated temperatures use the instantaneous values of P_1 and P_2 at any simulation time step. By adding a moving average integration of the estimated temperature or including a thermal capacitance model, the dynamic evolution of the temperature will be also recreated.

B.2.1 Moving average technique

By performing a *Simple Moving Average* [107] of the estimated temperatures, an average of the instantaneous values of temperature during certain period is calculated. In each time step, a new instantaneous value is added and the oldest data drop out. After certain simulation time, the averaged temperatures will stabilize at some point.

Using the same example than in the implementation of the static model –a system with two objects–, the VHDL-AMS coding is:

```
// --- Input port ---
PORT real in: Period = Tend; // "Period" is equal to the end simulation time,
unless other value is externally overwritten.

// --- Instantaneous temperatures ---
EQU {TempMaxNoAvg1:= P1*R11 + P2*R12 + Tamb} DST: SIM(Type:=SSGM, Sequ:=PRE);
EQU {TempMaxNoAvg2:= P2*R22 + P1*R21 + Tamb} DST: SIM(Type:=SSGM, Sequ:=PRE);

// --- Moving average temperatures ---
UMODEL MOV_AV TempMax1 (INPUT:=TempMaxNoAvg1, TSTART:=0, TSTOP:=Tend, PERIOD:=Period)
SRC: DB(Lib:="Simplorer_Elements\Basic_Elements\Basic_Elements");
UMODEL MOV_AV TempMax2 (INPUT:=TempMaxNoAvg2, TSTART:=0, TSTOP:=Tend, PERIOD:=Period)
SRC: DB(Lib:="Simplorer_Elements\Basic_Elements\Basic_Elements");
```

As a drawback of this method, the temperature estimation will incur in non-realistic oscillations during the start-up of the power converter or during a transient. On the other hand, the low computational cost is an advantage. Figure B.2 shows an example of a Flyback transformer by using this approach.

Figure B.2: Simulation results of the moving average thermal model of a Flyback transformer in ANSYS TwinBuilder: output voltage (red), drain-source voltage (orange), average core loss (green), average winding loss (blue) and maximum temperature (red).

B.2.2 Diagonal capacitance matrix

Considering that the interior temperature of the objects is uniform at all times during the heat transfer process, they can be treated as *lumped systems* without much sacrifice from accuracy [21]. This assumption is generally applicable when the *Biot number*, Bi , is lower than 0.1. Bi is the relationship between the convective heat exchange of the object to ambient and its inner conduction resistance, according to (B.1).

$$Bi = \frac{L_c/k}{1/h} \quad (B.1)$$

Being $L_c = V/A_{ext}$ the characteristic length (volume over external area), k the thermal conductivity of the object and h the film coefficient. This assumption is generally valid for any magnetic component.

Just to better visualize the presented concept, an electrical analogy can be considered. Even if the static proposed FEA-based matrices $[R_{th_{max}}]$ and $[R_{th_s}]$ are not thermal resistances ‘physically’, the circuit representation of a system with two objects from Figure B.3 serves as an example on how to mathematically formulate the dynamic model. A similar procedure was described in [108].

Figure B.3: Circuit representation of the dynamic model for two objects.

The current sources P_1 and P_2 are the internal heat generation (power loss) of each object. The terms R_{11} and R_{22} are the diagonal elements of the matrix $[R_{th_{max}}]$ (Equation 3.9); R_{12} and R_{22} are the crossed elements from the same matrix, representing the interaction between objects. Finally, C_{th1} and C_{th2} are the thermal capacitance of both objects. By solving this circuit analogy, the generic formulation of the temperature rise as a function of time for N objects is:

$$\begin{pmatrix} \Delta T_{max1}(t) \\ \Delta T_{max2}(t) \\ \vdots \\ \Delta T_{maxN}(t) \end{pmatrix} = [R_{th_{max}}] \cdot \begin{pmatrix} P_1 \\ P_2 \\ \vdots \\ P_N \end{pmatrix} \cdot \left([I] - \begin{pmatrix} e^{\frac{t}{R_{11} \cdot C_{th1}}} & 0 & \dots & 0 \\ 0 & e^{\frac{t}{R_{22} \cdot C_{th2}}} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & e^{\frac{t}{R_{NN} \cdot C_{thN}}} \end{pmatrix} \right) \quad (B.2)$$

Where the terms $R_{11}, R_{22}, \dots, R_{NN}$ correspond to the diagonal elements from the matrix $[R_{th_{max}}]$. $[I]$ is the identity matrix. The terms $C_{th1}, C_{th2}, \dots, C_{thN}$ are equal to the thermal capacitance of each object and correspond to:

$$C_{th_{ij}}|_{i=j} = c_i \cdot \rho_i \cdot Vol_i \quad (B.3)$$

Where c_i , ρ_i and Vol_i are the specific heat $[\frac{J}{K \cdot kg}]$, the material density $[\frac{kg}{m^3}]$ and the volume $[m^3]$ of each object, respectively.

However, if neglecting the non-diagonal terms, like in this case, the dynamic behavior of an object is only considered if it has internal heat generation (i.e. power losses), as shown in Figure B.4.a. The other objects would reach their steady-state temperature, given by

$[R_{th}]$, but a sudden temperature step would be seen in those objects, without the natural transition.

To add the dynamic thermal interaction between objects, the complete capacitance matrix is required, as explained next. This has an increase of the computational cost of the model as a counterpart.

Figure B.4: Circuit representation of the dynamic models based on capacitance matrices for three objects and their temperature over time if one of them has internal heat generation: a) Diagonal matrix, b) Complete matrix with simplifications, c) Complete matrix without simplifications.

B.2.3 Complete capacitance matrix

In order to model the dynamic interactions between objects, the complete capacitance matrix must be considered, according to the next equation:

$$\begin{pmatrix} \Delta T_{max_1}(t) \\ \Delta T_{max_2}(t) \\ \vdots \\ \Delta T_{max_N}(t) \end{pmatrix} = [R_{th_{max}}] \cdot \begin{pmatrix} P_1 \\ P_2 \\ \vdots \\ P_N \end{pmatrix} \cdot \left([I] - \begin{pmatrix} e^{\frac{t}{R_{11} \cdot C_{th_{11}}}} & e^{\frac{t}{R_{21} \cdot C_{th_{21}}}} & \dots & e^{\frac{t}{R_{N1} \cdot C_{th_{N1}}}} \\ e^{\frac{t}{R_{12} \cdot C_{th_{12}}}} & e^{\frac{t}{R_{22} \cdot C_{th_{22}}}} & \dots & e^{\frac{t}{R_{N2} \cdot C_{th_{N2}}}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{\frac{t}{R_{1N} \cdot C_{th_{1N}}}} & e^{\frac{t}{R_{2N} \cdot C_{th_{2N}}}} & \dots & e^{\frac{t}{R_{NN} \cdot C_{th_{NN}}}} \end{pmatrix} \right) \quad (B.4)$$

Where the thermal capacitance for the non-diagonal terms –if there is only conduction between objects i and j – is approximated to:

$$C_{th_{ij}}|_{i \neq j} \approx \frac{C_{th_{ii}} + C_{th_{jj}}}{2} \quad (B.5)$$

Adding all the terms to the capacitance matrix gives the most accurate dynamic thermal behaviour among the described options. However, this would considerably

increase the computational cost of the model. Furthermore, the terms $C_{th_{ij}}$ corresponding to the conductive heat transfer between objects i and j are the most relevant. The convection and radiation between objects can be neglected, since their temperature and the surrounding temperature are in the same range of values.

As a qualitative example, Figure B.4.b shows the effect of neglecting the $C_{th_{ij}}$ terms of separated objects. It can be seen that the dynamic behaviour of the green and yellow objects is not affected by this simplification, but the temperature of the blue object rises slowly compared to the more complete model (Figure B.4.c). However the difference is minimal. As a general rule to get a good compromise between accuracy and computational cost, considering the dynamic $C_{th_{ij}}$ terms corresponding to the conductive heat transfer between objects i and j is sufficient.

The decision on which dynamic model should be used is a trade-off between the accuracy of the dynamic evolution of the temperature and the computational cost.

Proposed $P_{core} - f - B - T$ formulation

Core losses are usually provided graphically by manufacturers, first as a function of frequency and flux density, and then as a magnitude dependent on temperature, for given combinations of frequency and flux density. This set of information is limited and not enough to accurately define the parameters of the Steinmetz equation and considering the temperature dependence at the same time. A formulation is proposed in this annex to account for this dependence, given the limited information provided by the ferrite manufacturers.

C.1 $P_v - f - B$ relationship

Core losses, P_{core} , are usually provided per volume unit, $P_v = P_{core}/V$. The classic approach is obtaining the coefficients k_{fe} , α and β of the Steinmetz equation from the corresponding graph (Figure C.1):

$$P_v = k_{fe} \cdot f^\alpha \cdot B^\beta \quad (\text{C.1})$$

To do so, a system of equations is solved by taking points from two curves of that graph [101]. However, α and β are not constant for every curve, as depicted in Figure C.1, where β_1 , β_2 and β_3 are different, indicating that each curve has a different slope. The parameter α is related to the ‘separation’ between curves and β to their slope. Assuming that α and β are constant would imply considering that the curves are equally separated proportionally to frequency and parallel, but this assumption differs from the actual core losses and cannot be generalized for a wide range of frequency and flux density.

In numerical design techniques, as optimization processes, a generalized equation of the core losses is desirable. In this way, core losses can be computed for any frequency or flux density avoiding the previous mentioned problem. There are two options: making a set of $P_v - B$ measurements –which is an expensive and time consuming method, but more accurate– or developing analytical approaches by means of a curve fitting of the data provided by the manufacturers [101]. By using the second option, experimental measurements are not required but it is constraint by the available information.

In this appendix, the second option is suggested. Instead of using the classical Steinmetz equation, a more generic formulation of the core losses is proposed according to the next expression:

$$P_v = C(f) \cdot B^{\beta(f)} \quad (\text{C.2})$$

Figure C.1: Core losses per volume curves as a function of magnetic flux for a given temperature and three frequencies ($f_1 > f_2 > f_3$) of certain commercial soft ferrite: data from manufacturers in solid lines; fitted curves in dashed lines. α and β are related to the separation between curves and their slope, respectively.

Where $C(f)$ and $\beta(f)$ are function of the frequency and can be expressed as (C.3) and (C.4), respectively.

$$C(f) = c_{ka} \cdot f^{c_{kb}} + c_{kc} \quad (C.3)$$

$$\beta(f) = c_{ba} \cdot f^2 + c_{bb} \cdot f + c_{bc} \quad (C.4)$$

The coefficients c_{ka} , c_{kb} , c_{kc} , c_{ba} , c_{bb} and c_{bc} are obtained by fitting the data from ferrite manufacturers. An application example is shown in Figure C.1. It can be seen that by means of this formulation, the computed core losses show a proper matching with the provided data regardless the ‘slope’ of the curves (i.e. the difference of slope from curves for the frequencies f_2 and f_3 is more pronounced, but the fitted curves are matching with the data).

At this point, a generalized core losses formulation as a function of the flux density and the frequency, for a specific temperature, is proposed. The next step is providing the temperature dependence.

C.2 $P_v - T$ relationship

In order to account for the temperature dependence of power losses, certain coefficients must be added to (C.2). In some manufacturer datasheets and in the research from [63], [64], this dependence is assumed as a second order polynomial for given magnetic flux density and frequency pair according to:

$$P_v = (k_{fe} \cdot f^\alpha \cdot B^\beta) \cdot (c_{x2} \cdot T^2 + c_{x1} \cdot T + c_{x0}) \quad (C.5)$$

Nevertheless, due to the asymmetry of the actual curves, this equation is suitable just for one range: below the temperature at which the core losses are minimized or above it

(Figure C.2). By adding a degree on that equation, a better fitting can be obtained for the whole range of temperatures. So, (C.6) is proposed instead:

$$P_v = C(f) \cdot B^{\beta(f)} \cdot (c_{t3}(B) \cdot T^3 + c_{t2}(B) \cdot T^2 + c_{t1}(B) \cdot T + c_{t0}(B)) \quad (C.6)$$

Since it is desirable to have an equation as general as possible, coefficients c_{t0} , c_{t1} , c_{t2} , c_{t3} and c_{t4} can be expressed as a function of the flux density. The dependence of these coefficients on the frequency is avoided due to the limited available data from manufacturers and for simplicity.

Depending on the available information from the manufacturer, the fitting expression for the coefficients c_{t0} , c_{t1} , c_{t2} , c_{t3} and c_{t4} as a function of the flux density can be either a first order polynomial:

$$c_t|_{0,1,2,3}(B) = c_{tb} \cdot B + c_{ta} \quad (C.7)$$

if the $P_v - T$ curves are given for only two values of B. Or a second order polynomial:

$$c_t|_{0,1,2,3}(B) = c_{tc} \cdot B^2 + c_{tb} \cdot B + c_{ta} \quad (C.8)$$

if the curves are given for more than two values of B.

Notice that the fitting expressions are reliable in the same range that data is provided. An extrapolation out of that range is not as accurate as interpolation. For example, if $P_v - T$ curves are given for $T = [0, 120]^\circ\text{C}$, the resulting fitting is no longer valid over 120°C or below 0°C . Also, the correlation between this proposed fitting and the actual core losses of a magnetic component is better with a higher amount of data, provided either by manufacturers or obtained by experimental measurements.

As a summary, the generalized equations (C.2) and (C.6) allow the computation of core losses for a wide range of frequency, magnetic flux density and temperature, what is convenient for numerical design processes such as optimizations. The accuracy of this approximations depends on the available data.

Figure C.2: Core losses per volume as a function of temperature curves for four combinations of frequency and magnetic flux of certain commercial soft ferrite: data from manufacturers in dots, fitted curves in solid lines.

References: thermal models

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