

Article

Modeling of the Resistance of a Coil Module with Ferrite and Shielding for Wireless Power Transfer in Electric Vehicles

Timo Lämmle *  and Nejila Parspour 

Institute of Electrical Energy Conversion, University of Stuttgart, Pfaffenwaldring 47, 70569 Stuttgart, Germany

* Correspondence: timo.laemmle@iew.uni-stuttgart.de

Abstract

For optimal design of a wireless power transfer system, knowing the electrical resistance of the coil modules is essential. For vehicle applications, a flux guidance and aluminum shielding are used in addition to the winding. Beyond considering the individual components, an understanding of the interactions between them is also necessary to determine the coil module resistance. This paper first provides the theoretical background of the relevant loss effects in litz wire and coil modules. Existing approaches are then analyzed for their applicability for the application considered in this work. Based on this analysis, a practical method is proposed that systematically combines established approaches from the literature to calculate the electrical resistance of a coil module while accounting for the effects of both flux guidance and shielding. In contrast to existing approaches, which primarily address individual components or effects, the proposed method enables the resistance of the entire coil module to be determined within a unified modeling framework. The method is verified by comparison with measured results, showing good agreement with a maximum deviation of 2%. The presented approach supports the efficient design of wireless power transfer systems and substantially reduces the need for repeated coil module prototyping and resistance measurements.

Keywords: wireless power transfer; inductive charging; wireless charging; inductive power transfer; electric vehicles; resistance modeling; ferrite; shielding; skin effect; proximity effect



Academic Editor: Peter Van den Bossche

Received: 28 July 2026

Revised: 30 September 2026

Accepted: 6 October 2026

Published: 15 October 2026

Copyright: © 2026 by the authors.

Published by MDPI on behalf of the

World Electric Vehicle Association.

Licensee MDPI, Basel, Switzerland.

This article is an open access article

distributed under the terms and

conditions of the [Creative Commons](https://creativecommons.org/licenses/by/4.0/)[Attribution \(CC BY\)](https://creativecommons.org/licenses/by/4.0/) license.

1. Introduction

The development of wireless power transfer (WPT) systems for charging electric vehicles (EVs) is the focus of numerous current research projects, which have been steadily increasing in number for many years [1–10]. By eliminating physical contact, the technology is robust against all weather conditions, is low-wear, and offers the user a higher level of comfort [11,12]. In addition, a future autonomous driving or automated parking concept also requires an automated charging solution without any user interaction [13–16].

Power is transferred from the Ground Assembly (GA) to the Vehicle Assembly (VA) via an alternating magnetic field as schematically shown in Figure 1. The GA is the infrastructure component that is installed on the ground and connected to the grid. Due to the integration of the VA in the underbody of the vehicle, suitable magnetic field shielding is required for safety reasons. This ensures that the applicable limit values are not exceeded inside and around the vehicle [17,18]. In addition, the efficiency of WPT can be increased by shielding the vehicle components, which are usually made of steel, near the VA with aluminum shielding [19]. In order to counteract the field-weakening effect of the shielding in the vehicle and also to increase the magnetic coupling to the infrastructure coil, a

flux guidance made of ferrite is used. This material is characterized by a high relative permeability and at the same time a low electrical conductivity [20]. The setup consisting of winding, flux guide, and shielding is referred to as a coil module in this paper.

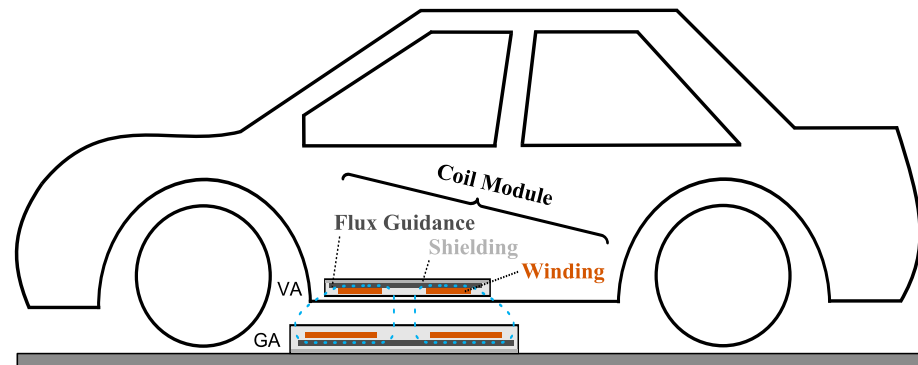


Figure 1. Schematic setup of a WPT system in an EV with the Ground Assembly (GA) on the ground and the Vehicle Assembly (VA) in the car underbody. Each coil module consists of winding, flux guidance, and shielding.

An optimal design of the WPT system requires precise knowledge of the resistance of the coil modules. In addition, a thorough understanding of the loss effects is needed to identify optimization potential. Special high-frequency litz wires are used for the windings, which consist of a large number of insulated individual strands. These in turn are twisted in bundles on several levels to reduce the losses caused by skin and proximity effects. In addition to the complex litz wire structure, the winding losses are influenced by the flux guidance and shielding. Furthermore, losses occur within the flux guidance and shielding themselves. Consequently, when determining the electrical resistance of a coil module, all loss effects and interactions between the components must be considered.

One possibility of determining the electrical resistance of a coil module is its assembly and measurement. However, this requires the procurement of the individual components and the assembly of the coil module. Investigating design variations or conducting parameter studies requires mechanical adjustments. As a result, the process can be both time-consuming and cost-intensive. Therefore, a modeling method for the resistance of a coil module is required taking into account flux guidance and shielding effects. Furthermore, the method should reduce the need for repeated coil module prototyping and resistance measurements.

Several studies have already addressed this general topic with different methodological and physical focuses. Without claiming to provide a comprehensive literature review, selected representative contributions can be grouped as follows:

- Analytical approaches to determine the effects of eddy current in litz wires or windings, as presented in [21–29];
- Numerical approaches like the Finite-Element Method (FEM), as presented in [30];
- Measurement-based approaches for characterizing litz wires and windings [31];
- Combined analytical, numerical, and measurement-based modeling approach for the resistance of a winding with flux guidance as presented in [32];
- Analysis of the influence of conductive materials on the resistance of a winding as presented in [25,28,33].

These studies provide important foundations and address relevant aspects of the problem, as the detailed analysis of these approaches in Section 4 shows. However, each approach covers only parts of what is required for the application considered here. A holistic method for modeling the resistance of a coil module that simultaneously considers flux guidance and shielding effects has not yet been introduced. Existing approaches typically

focus on individual loss mechanisms, isolated components, or conductor characterization. A methodology that combines these approaches into a unified workflow for determining the resistance of an entire EV-WPT coil module including litz-wire effects, ferrite influence, and shielding losses is currently lacking.

In the end, a practical method is key while achieving high accuracy. To achieve this, the effort of costly and time-consuming measurement setups with coil module prototyping as well as complex and time-consuming simulation models should be reduced. Therefore, this gap is addressed in this paper.

In Section 2, the boundary conditions for a WPT system for use in an EV are specified, and then the analytical equations for the loss calculation are derived in Section 3. After analyzing the loss mechanisms and discussing various approaches from the literature (Section 4), a practical method for modeling of the resistance of a coil module is proposed, applied, and finally compared with measured values for verification (Section 5).

2. Boundaries for WPT in EV

Several standardization documents currently provide boundary conditions and design guidelines for the development of WPT systems for EVs. Examples for this are the classification of the systems into different WPT power classes for one maximum input power each as listed in Table 1 and three Z-classes depending on the ground clearance of the vehicle as listed in Table 2.

Table 1. WPT power classifications in [34].

WPT-Class	WPT 1	WPT 2	WPT 3
Range of Input Volt-Amps	0 to 3.7 kVA	0 to 7.7 kVA	0 to 11.1 kVA

Table 2. Definition of the Z-classes in [34].

Z-Class	Z1	Z2	Z3
VA Coil			
Ground Clearance Range	100 mm to 150 mm	140 mm to 210 mm	170 mm to 250 mm

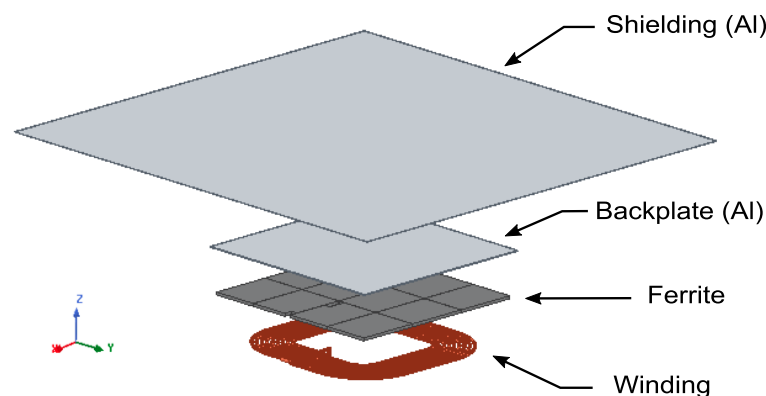
For the VA, reference designs for the setup of a coil module are given in SAE J2954 [34], IEC 61980-3 [35] and GB/T 38775.7 [36], whereby the first two documents are harmonized and are used as a reference for the analyses in this paper. The nominal frequency for power transfer is defined as 85 kHz in [34,35]. In this paper, analyses and verification of the proposed method are carried out using the VA coil module of the WPT3/Z2 class from SAE J2954 as an example.

Figure 2 shows the setup of the coil module schematically as an exploded view consisting of winding, ferrite, backplate and shielding. The aluminum backplate forms the top of the coil housing. An additional aluminum plate is centrally positioned directly on the backplate to simulate the shielding in the vehicle and to take into account the influence of the vehicle underbody.

The mechanical parameters for the setup of the coil module, such as the dimensions of the components, their distances from each other, the length of the winding, and the number of turns, are listed in Table 3.

Table 3. Mechanical parameters of the WPT3/Z2 coil module setup [34].

Parameter	Value
Winding Size (x, y, $\varnothing$)	320 × 320 mm, 5 mm
Ferrite Size (x, y, z)	334 × 334 × 5 mm
Backplate Size (x, y, z)	350 × 350 × 2 mm
Shielding Size (x, y, z)	800 × 800 × 1 mm
Distance Winding-Ferrite	1 mm
Distance Ferrite-Backplate	5.6 mm
Distance Backplate-Shielding	0 mm
Winding Length	7.8 m
Number of turns	9

**Figure 2.** Exploded view of the WPT3/Z2 circular VA coil module from SAE J2954 [34].

3. Fundamentals

This section provides an overview of the loss mechanisms that occur in conductors carrying alternating current. The starting point is the frequency-independent DC resistance R_{DC} of a conductor with length l and cross-sectional area A . It can be calculated from the conductor geometry and the electrical conductivity σ .

$$R_{DC} = \frac{l}{\sigma A} \quad (1)$$

In addition to DC losses, alternating currents cause eddy-current losses as a result of skin and proximity effects. The skin effect leads to a frequency-dependent concentration of current at the conductor surface. As a result, the current distribution becomes non-uniform while the total current remains constant. The proximity effect originates from magnetic fields produced by neighboring current-carrying conductors. These fields induce circulating eddy currents that further distort the current distribution. Both effects increase the effective conductor resistance and associated losses.

Figure 3a schematically shows the cross-section of a circular conductor, including the definition of the skin depth δ and the external magnetic field strength $\vec{H}_{ext}$, with the latter being responsible for the proximity effect. To reduce the additional losses caused by skin and proximity effects, special litz wires are used at higher frequencies. Litz wires consist of a large number of insulated strands that are twisted into bundles. As illustrated in Figure 3b, the strands interact through the internal magnetic field strength $\vec{H}_{int}$. Consequently, the internal proximity effect introduces an additional loss component that must be considered.

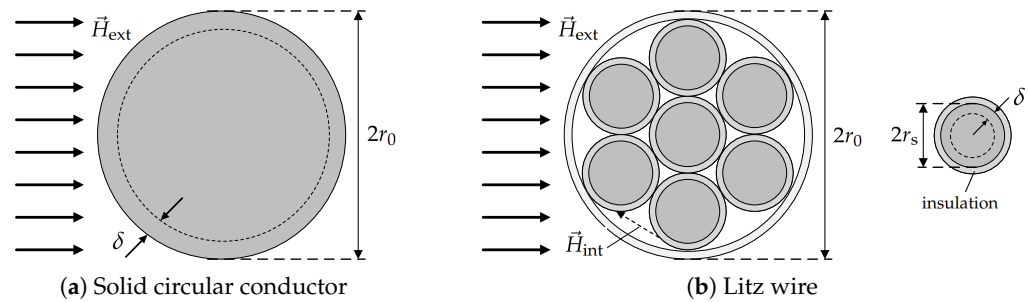


Figure 3. Schematic illustration of the cross-section including the eddy current effects for a circular conductor and a litz wire.

Based on the explanations provided in [37–39], the skin and proximity effects are discussed in Sections 3.1 and 3.2. These fundamentals are subsequently used to derive the loss calculations for a circular conductor in Section 3.3 and for an ideal litz wire in Section 3.4.

Due to the superposition of several physical effects, the analytical calculation of conductor losses is highly complex and therefore requires simplifying assumptions. The derivations presented in this work follow a quasi-static approach and neglect displacement currents. Furthermore, cylindrical coordinates and infinitely long circular conductors in the z -direction are assumed. Under these conditions, the relevant physical quantities can be considered constant along the conductor without loss of accuracy [40]. Quantities marked with a hat like $\hat{H}$ denote complex phasor amplitudes (peak values), whereas loss calculations are generally expressed using RMS quantities.

3.1. Skin Effect

The skin depth δ defines the distance from the edge of the conductor at which the current density has dropped to $1/e$ of the amplitude value and is dependent on the frequency f and the properties of the material like the electrical conductivity σ and the relative permeability μ_r [41].

$$\delta = \sqrt{\frac{1}{\sigma \pi f \mu_0 \mu_r}} \quad (2)$$

The amplifying effect of the current displacement to the edge with increasing frequency leads to a frequency-dependent increase in losses due to the skin effect P_{skin} . This additional loss component can be calculated using the frequency-dependent skin factor $F_s(f)$, the DC resistance and the squared RMS current I_{rms} [41]. The calculation of $F_s(f)$ depends on the geometry and is given for a circular conductor in (10) and for a litz wire in (18).

$$P_{\text{skin}} = R_{\text{DC}} I_{\text{rms}}^2 (F_s(f) - 1) \quad (3)$$

When analyzing the skin effect, only the radial dependence and thus the component of the current density in the z -direction has to be taken into consideration [32]. With the modified Bessel functions with zero order (J_0) and first order (J_1), a sinusoidal excitation results in the current characteristic in (4) for the skin effect within a circular conductor with radius r_0 and peak current phasor $\hat{I}$.

$$\hat{j}(r) = \frac{\hat{I}}{2\pi r_0^2} \frac{\alpha r_0 J_0(\alpha r)}{J_1(\alpha r_0)} \quad (4)$$

The procedure for solving the resulting differential equations using Kelvin or Bessel functions is derived and described in detail in [37,42]. The used auxiliary variable α can be calculated via the skin depth δ from (2).

$$\alpha = \frac{1+j}{\delta} = (1+j)\sqrt{\sigma\pi f\mu_0\mu_r} \quad (5)$$

3.2. Proximity Effect

The additional losses due to the external proximity effect $P_{\text{ext,prox}}$ can be calculated using the peak of the external magnetic field $\hat{H}$, the conductor length l , its conductivity σ , and the frequency-dependent proximity factor D_s [37]. The calculation of $D_s(f)$ depends on the geometry and is given for a circular conductor in (11) and for a litz wire in (21).

$$P_{\text{ext,prox}} = \frac{l}{\sigma} \hat{H}^2 D_s(f) \quad (6)$$

For the influence on the current density, an unpowered circular conductor in an alternating magnetic field is considered. Compared to the skin effect, the resulting current density due to the external proximity effect is also dependent on the direction of the external magnetic field amplitude $\hat{H}$. Therefore, the current density $\hat{j}$ is not rotationally symmetrical to the center of the conductor, and the calculation in (7) requires cylindrical coordinates (r, ϕ) [41].

$$\hat{j}(r, \phi) = \frac{2\hat{H}}{r_0} \frac{\alpha r_0 J_1(\alpha r)}{J_0(\alpha r_0)} \sin(\phi) \quad (7)$$

3.3. Loss Calculation for Circular Conductor

The individual loss components can be determined separately according to [31,39,43–45] due to the orthogonality of the loss components. This simplifies the calculation of the total losses. For a solid circular conductor, the total conductor losses P_{cond} can therefore be expressed as the sum of three loss components:

$$P_{\text{cond}} = P_{\text{DC}} + P_{\text{skin}} + P_{\text{ext,prox}} \quad (8)$$

The frequency-independent DC-losses P_{DC} can be calculated with the squared RMS current I_{rms} and the ohmic resistance R_{DC} from (1) for a circular conductor with radius r_0 .

$$P_{\text{DC}} = R_{\text{DC}} \cdot I_{\text{rms}}^2 \quad \text{with} \quad R_{\text{DC}} = \frac{l}{\sigma\pi r_0^2} \quad (9)$$

Integrating the current density from (4) over the conductor cross-section results in the frequency-dependent losses caused by the current flow in the conductor. The skin factor F_s can be used to determine the additional loss component due to the skin effect P_{skin} according to (3).

$$P_{\text{skin}} = R_{\text{DC}} I_{\text{rms}}^2 (F_s - 1) \quad \text{with} \quad F_s = \frac{1}{2} \text{Re} \left\{ \alpha r_0 \frac{J_0(\alpha r_0)}{J_1(\alpha r_0)} \right\} \quad (10)$$

The calculation of the losses $P_{\text{ext,prox}}$ due to the external proximity effect is carried out according to (6) with the proximity factor D_s .

$$P_{\text{ext,prox}} = \frac{l}{\sigma} \hat{H}^2 D_s \quad \text{with} \quad D_s = 2\pi \text{Re} \left\{ \alpha r_0 \frac{J_1(\alpha r_0)}{J_0(\alpha r_0)} \right\} \quad (11)$$

3.4. Loss Calculation for an Ideal Litz Wire

The same losses as in a solid circular conductor occur in a litz wire. The structure with several individual strands adds losses $P_{\text{int,prox}}$ due to the internal proximity effect, as shown in Figure 3b.

$$P_{\text{litz}} = P_{\text{DC}} + P_{\text{skin}} + P_{\text{int,prox}} + P_{\text{ext,prox}} \quad (12)$$

The DC-losses of a litz wire can be calculated from its structure with the number of strands N_s , the radius of the individual strand r_s and the length l [32].

$$P_{\text{DC}} = R_{\text{DC}} \cdot I_{\text{rms}}^2 = \frac{l}{\sigma_{\text{Cu}} N_s \pi r_s^2} \cdot I_{\text{rms}}^2 \quad \text{with} \quad R_{\text{DC}} = \frac{l}{\sigma_{\text{Cu}} N_s \pi r_s^2} \quad (13)$$

The copper fill factor k_{Cu} corresponds to the ratio of the electrically conductive copper area to the total cross-sectional area of the litz wire, resulting in an adapted electrical conductivity.

$$\sigma_{\text{litz}} = k_{\text{Cu}} \sigma_{\text{Cu}} \quad \text{with} \quad k_{\text{Cu}} = \frac{N_s \pi r_s^2}{\pi r_0^2} \quad (14)$$

The complex structure of the litz wire, consisting of individual strands, bundles, the type of twisting and connection technology, makes it difficult to determine the loss components. To simplify this, the two borderline cases of twisting are considered: ideally twisted litz wire and litz wire with individual strands arranged in parallel. The latter has the highest losses, as the parallel individual strands have no influence on the generation and impact of the eddy current effects, whereas ideal twisting leads to the lowest losses [32,41].

According to [31], the equations for an equivalent circular conductor with the adapted conductivity from (14) can be approximately adopted for the litz wire with individual strands arranged in parallel. There are no losses due to the internal proximity effect, and the losses due to the skin effect $P_{\text{par,skin}}$ can be calculated according to (10) with the normalized skin parameter of a litz wire with parallel individual strands x_0 .

$$x_0 = \alpha r_0 = (1 + j) r_0 \sqrt{\sigma_{\text{litz}} \pi f \mu_0 \mu_r} \quad (15)$$

$$P_{\text{par,skin}} = R_{\text{DC}} \cdot I_{\text{rms}}^2 \cdot \left(\frac{1}{2} \text{Re} \left\{ x_0 \frac{J_0(x_0)}{J_1(x_0)} \right\} - 1 \right) \quad (16)$$

For an ideally twisted litz wire, the losses due to skin and proximity effects at the bundle level can be neglected, so that only the losses occurring at the strand level need to be considered. The current is divided equally between all individual strands in this case, so that the losses can be calculated as the sum of the losses of all individual strands. This results in the skin losses $P_{\text{id,skin}}$ for an ideally twisted litz wire with the skin factor at strand level $F_{s,\text{strand}}$ and the normalized skin parameter of an ideally twisted litz wire x_s .

$$P_{\text{id,skin}} = N_s \cdot R_{\text{DC}} N_s \cdot \left(\frac{I_{\text{rms}}}{N_s} \right)^2 \cdot (F_{s,\text{strand}} - 1) = R_{\text{DC}} I_{\text{rms}}^2 (F_{s,\text{strand}} - 1) \quad (17)$$

$$F_{s,\text{strand}} = \frac{1}{2} \text{Re} \left\{ x_s \frac{J_0(x_s)}{J_1(x_s)} \right\} \quad \text{with} \quad x_s = \alpha r_s = (1 + j) r_s \sqrt{\sigma_{\text{Cu}} \pi f \mu_0 \mu_r} \quad (18)$$

The internal H_{int} and external magnetic field strength H_{ext} are required to calculate the losses due to the proximity effect. The internal magnetic field strength depends on the position of a strand within the conductor. The magnitude of the internal magnetic field strength $\hat{H}_{\text{int}}(r)$ can be calculated based on the excitation current. Assuming an ideal litz wire, this current is distributed uniformly among all strands.

$$\hat{H}_{\text{int}}(r) = \frac{\hat{I}r}{2\pi r_0^2} \quad (19)$$

The losses $P_{\text{id,int,prox}}$ of an ideally twisted litz wire can be determined by the internal proximity effect when inserted in (6). The proximity factor at strand level $D_{\text{s,strand}}$ becomes zero for ideal twisting at all geometry levels [38].

$$P_{\text{id,int,prox}} = N_s \frac{l}{\sigma_{\text{Cu}}} \left(\frac{\hat{I}}{2\pi r_0} \right)^2 2\pi \text{Re} \left\{ x_s \frac{J_1(x_s)}{J_0(x_s)} \right\} = \frac{1}{2} R_{\text{DC}} N_s k_{\text{Cu}} \text{Re} \left\{ x_s \frac{J_1(x_s)}{J_0(x_s)} \right\} \hat{I}^2 \quad (20)$$

$$D_{\text{s,strand}} = 2\pi \text{Re} \left\{ x_s \frac{J_1(x_s)}{J_0(x_s)} \right\} \quad (21)$$

Assuming ideal twisting at the bundle level, the eddy currents arising there cancel each other out, so that the calculation of losses due to the external proximity effect can take place at the individual strand level. These are the same for each strand and can therefore be added up for the total losses of the litz wire. This results in the losses due to the external proximity effect $P_{\text{par,ext,prox}}$ for a litz wire with individual strands arranged in parallel and $P_{\text{id,ext,prox}}$ for the ideally twisted litz wire.

$$P_{\text{par,ext,prox}} = \frac{l}{\sigma_{\text{litz}}} \hat{H}_{\text{ext}}^2 2\pi \text{Re} \left\{ x_0 \frac{J_1(x_0)}{J_0(x_0)} \right\} \quad (22)$$

$$P_{\text{id,ext,prox}} = N_s \frac{l}{\sigma_{\text{Cu}}} \hat{H}_{\text{ext}}^2 2\pi \text{Re} \left\{ x_s \frac{J_1(x_s)}{J_0(x_s)} \right\} \quad (23)$$

4. Modeling of the Resistance in a Coil Module with Ferrite and Shielding

This section presents a method for the determination of the resistance of an entire coil module consisting of a winding, a ferrite layer and aluminum shielding. In the first section, the loss effects in a coil module are analyzed in order to gain a precise understanding of the individual components. For the calculation of the different resistance components, the appropriate method for the application considered in this paper is selected and explained in the following sections after discussing various approaches. Finally, the presented modeling method is summarized in a schematic workflow.

4.1. Analysis of the Loss Mechanisms in the Coil Module

A coil module used in a vehicle requires both a ferrite layer for flux guidance and an aluminum shielding. This results in various loss effects, as illustrated in Figure 4. The different effects can be considered separately due to the orthogonality. However, the decomposition of the loss terms does not necessarily mean that there is no electromagnetic coupling between the physical components.

The three loss components of DC, skin, and internal proximity effect shown at the bottom are all dependent only on the structure and properties of the litz wire. It is therefore useful to consider them together as presented in Section 4.2. By arranging the conductors in a winding with several turns, they influence each other due to the external proximity effect. In addition, the coil module contains soft magnetic materials with a high relative permeability, such as ferrite. It also contains electrically conductive materials, such as aluminum. Both influence the external H-field and thus the loss component caused by the external proximity effect in the litz wire.

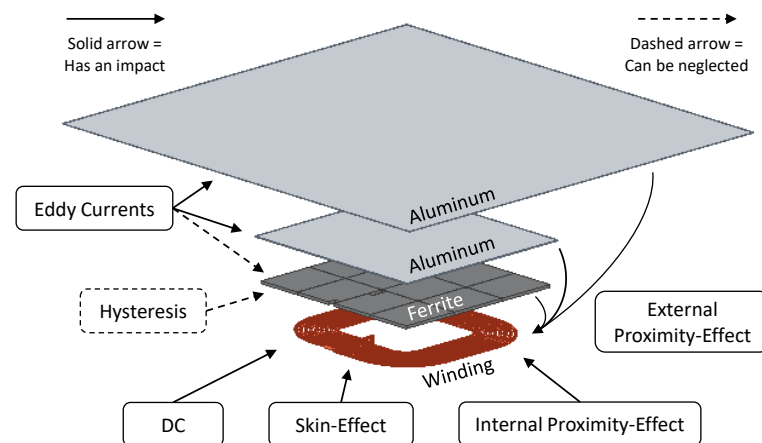


Figure 4. Overview of the loss effects in a coil module with ferrite and shielding.

Beside the loss mechanisms in the litz wire, the effects in ferrite and in aluminum have to be analyzed as well. The alternating magnetic field generates eddy currents in electrically conductive components and thereby creates losses. When measuring the electrical resistance of the entire coil module, for example with an LCR meter connected to both ends of the winding, these losses are inherently included in the measurement. Therefore, they must be taken into account in the modeling.

The core losses in the ferrite are mainly composed of hysteresis and eddy current losses [46]. Due to soft magnetic ferrite with its narrow hysteresis curve, core losses can be reduced by selecting a material that matches the operating frequency [20,47]. In addition, the conductivity is very low, so the eddy current losses in the ferrite can be neglected [43]. Therefore, this arrow is shown with a dashed line in Figure 4. A low-signal measurement can be performed using an LCR meter, as described in Section 5.4. In this case, both the excitation current in the winding and the resulting flux density in the ferrite are very low. As a result, the hysteresis losses can be neglected for the application considered here [48] and are therefore shown in dashed lines. In power-transfer operation, higher coil currents result in a higher magnetic flux density in the ferrite material. Under these conditions, hysteresis losses may become relevant and must be taken into account [49–51].

In summary, the loss components of (12) need to be extended by the component of the eddy currents in the aluminum shielding R_{eddy} . The total electrical resistance of a coil module R_{module} is the sum of its individual resistance components.

$$R_{\text{module}} = \underbrace{R_{\text{DC}} + R_{\text{skin}} + R_{\text{int,prox}}}_{\text{section 4.2}} + \underbrace{R_{\text{ext,prox}}}_{4.3} + \underbrace{R_{\text{eddy}}}_{4.4} \quad (24)$$

Various options for determining the individual resistance components are analyzed in the following sections as marked in (24) and a suitable method is identified for the application being considered here.

4.2. Determination of the Resistance Considering DC, Skin and Internal Proximity Effects

The three resistance components associated with DC, the skin effect and the internal proximity effect depend only on the structure of the litz wire. Consequently, these three contributions are considered jointly in the following. The DC resistance of the litz wire can be determined in the first step via (1) and the copper fill factor from (14). Various methods for determining the resistance components due to the skin and internal proximity effects can be found in the literature. Detailed approaches are presented in [23–25] to analyze the influence of the twisting of the litz wire on these loss components. With the subdivision of the loss components caused by the transverse and longitudinal magnetic field, ref. [26]

presents an alternative approach. Due to the twisting, the individual strands in the litz wire cross-section do not have the same direction vector as the overall litz wire. This angle creates an additional component for the internal magnetic field, which must be taken into account in the loss calculation [27,28].

With all these approaches mentioned here as excerpts, a deep understanding of the loss effects within the litz wire can be achieved. However, the analytical approach requires considerable effort when the entire litz wire structure is taken into account. In addition, the deviation increases with the number of strands [29].

Numerical methods are an alternative to analytical calculations. In [30], for example, it was shown that the effects of skin and proximity effects within the litz wire can be calculated approximately using the Finite-Element Method (FEM). This was shown for a sufficiently thin strand diameter, whereby the skin effect within the strand is negligible, and for a low number of strands (≤ 80). However, the simulation effort increases exponentially with the number of strands. It further increases when multiple twisting levels and several bundles are considered. For the application considered here, in which a winding has several turns and is not rotationally symmetric, a numerical method is therefore not suitable.

Another possibility is the measurement-based characterization of an existing litz wire. For each real litz wire, its behavior for a certain frequency range lies between that of an ideally twisted litz wire and that of a litz wire with parallel strands. These two extreme scenarios can each be described analytically according to the description in Section 3.4. By introducing the parameter λ_{skin} , the behavior of a real litz wire can be characterized. The resistance of the real litz wire R_{litz} is then described by a linear combination of the resistance of the ideally twisted litz wire R_{id} and the resistance of the litz wire with parallel strands R_{par} [31].

$$R_{\text{litz}} = \lambda_{\text{skin}} R_{\text{id}} + (1 - \lambda_{\text{skin}}) R_{\text{par}} = R_{\text{DC}} + R_{\text{skin}} + R_{\text{int,prox}} \quad (25)$$

An ideal litz wire has $\lambda_{\text{skin}} = 1$ and a litz wire with parallel strands has $\lambda_{\text{skin}} = 0$. A suitable test setup for determining λ_{skin} is described in Section 5.1, in which the litz wire is placed between two metal bars in a groove. By doing so, it is ensured that the influence of magnetic stray fields from the environment is excluded and only the losses of the conductor considering DC, skin and internal proximity effects are determined.

The determination of R_{id} and R_{par} can be carried out using the equations derived in Section 3.4, whereby the frequency-dependent resistance change has to be considered instead of the power. Both the DC component and the resistance components considering skin and internal proximity effects are included in R_{id} . Therefore, the correlations in (17) and (20) can be summarized in (26). The twisting increases the effective length of the strands compared to the litz wire. This increase in length, expressed by the factor m_{twist} , can be calculated based on the litz wire design and the stranding structure [23,52]. It is taken into account via the adjusted resistance $R_{\text{DC,twist}} = m_{\text{twist}} R_{\text{DC}}$.

$$R_{\text{id}} = R_{\text{DC,twist}} \left(\frac{1}{2} \text{Re} \left\{ x_s \frac{J_0(x_s)}{J_1(x_s)} \right\} + N_s k_{\text{Cu}} \frac{1}{2} \text{Re} \left\{ x_s \frac{J_1(x_s)}{J_0(x_s)} \right\} \right) \quad (26)$$

A litz wire with strands arranged in parallel is represented by an equivalently sized solid circular conductor. Therefore, the increase in length does not need to be considered. Consequently, there are no losses due to the internal proximity effect, and therefore R_{par} is calculated based on (16).

$$R_{\text{par}} = R_{\text{DC}} \frac{1}{2} \text{Re} \left\{ x_0 \frac{J_0(x_0)}{J_1(x_0)} \right\} \quad (27)$$

This procedure with the characterization by measurement considers the exact structure of a real litz wire, including the effects of twisting and potential twisting errors during production. For a real litz wire, the contacting method further influences the losses caused by the skin and internal proximity effects, as shown in [30,53]. This influence due to the contacting of the litz wire is included here in the measurement-based characterization. It is recommended to carry out the characterization of the litz wire with the same contacting method as in the later setup.

4.3. Determination of the Resistance Considering the External Proximity Effect

This resistance component does not depend on the magnetic field generated by the conductor itself. Instead, it depends on the external field generated by neighboring windings and influenced by surrounding materials.

The difference in the resulting magnetic field outside a conductor is negligible between a litz wire and an equivalently sized solid circular conductor [32,54]. In both cases, the field depends only on the current. Regarding the determination of the losses due to the external proximity effect, a rotationally symmetrical distribution of the current density in the litz wire can be assumed [32]. Therefore, the litz wire can be considered as a solid circular conductor for this analysis.

The internal magnetic field strength H_{int} equals zero at the center of a conductor with a round cross-section. This property can be used to determine the external magnetic field strength H_{ext} . In an FEM simulation model, a polyline is placed at the center of the conductor along its winding path [32]. The magnetic field strength extracted along this polyline corresponds to the desired field component H_{ext} .

By dividing the polyline into n segments with length Δl , the magnitude of the magnetic field strength H_{ext} can be calculated at multiple points. The sum of the losses of the individual segments results in the total losses of a coil due to the external proximity effect. Taking into account the selected current excitation in the simulation model, the calculation of the resistance component due to the external proximity effect is performed using (23). The polyline with its length does not correspond to the real length of the litz wire. Therefore, the increase in length m_{twist} has to be considered due to twisting so that the actual length of the wire exposed to the external field is taken into account.

$$R_{\text{ext,prox}} = \frac{N_s m_{\text{twist}} D_{s,\text{strand}}}{\sigma_{\text{Cu}} I^2} \sum_1^n |H_{\text{ext}}|^2 \Delta l \quad (28)$$

With regard to losses due to the external proximity effect, the electrical behavior of a real litz wire in an external magnetic field can be assumed to be ideal, as long as the length exposed to the field is many times longer than the pitch [31]. Therefore, a measurement-based characterization of the real litz wire is not necessary at this point.

4.4. Determination of the Resistance Considering Eddy Currents

The influence of electrically conductive materials on the electrical resistance of the winding has already been discussed in [25,33]. Measurements in [28] show an additional loss component when the material is approached, whereas the eddy currents reduce the modeled losses in the litz wire due to the field attenuation. This emphasizes the importance of considering the resistance component R_{eddy} due to the eddy currents in (24).

The proposed method for this is to calculate the eddy current losses using an FEM simulation. For this purpose, the complex peak phasor amplitude of the induced current density $\hat{j}$ is determined in the material. The eddy current losses P_{eddy} can then be calculated by integrating the resulting current density distribution over the material volume.

$$P_{\text{eddy}} = \frac{1}{2\sigma} \int_V |\hat{j}|^2 dV \quad (29)$$

This simulated power loss can then be converted into the equivalent resistance component R_{eddy} using the RMS excitation current in the simulation model.

$$R_{\text{eddy}} = \frac{P_{\text{eddy}}}{I_{\text{rms}}^2} \quad (30)$$

With this new approach for considering the shielding influence, all resistance components can be determined and the modeling method is complete.

4.5. Summary of Modeling Method

To summarize the modeling method presented, Figure 5 shows the procedure for determining the resistance of a coil module with ferrite and aluminum shielding via the three different pathways. The novelty of the proposed method is not a new analytical loss equation, but the systematic combination of analytical, measurement-based, and FEM-based approaches into a single workflow that allows the resistance of a complete coil module, including ferrite and conductive shielding, to be determined with low experimental effort.

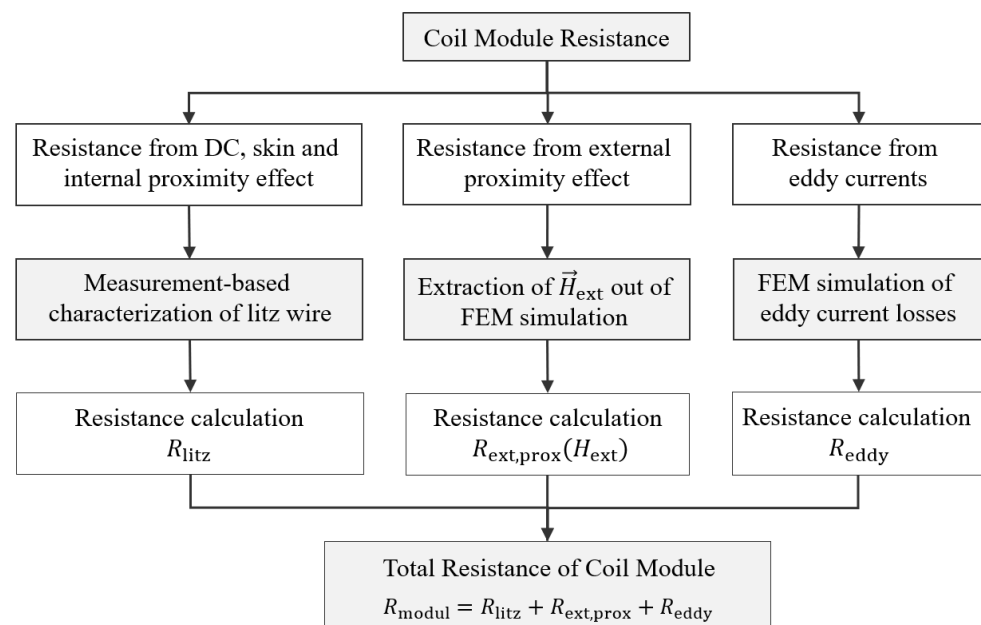


Figure 5. Schematic workflow of the presented modeling method for calculation of the resistance of a coil module with ferrite and aluminum shielding.

5. Results

The new methodology is applied in this section using the example of the circular coil from Figure 2. Three scenarios are considered here: the winding only; the winding with ferrite; and the entire coil module consisting of winding, ferrite, and shielding. After calculating the resistance components R_{litz} , $R_{\text{ext,prox}}$ and R_{eddy} , the modeled values are then compared with the measurement results to verify the presented method.

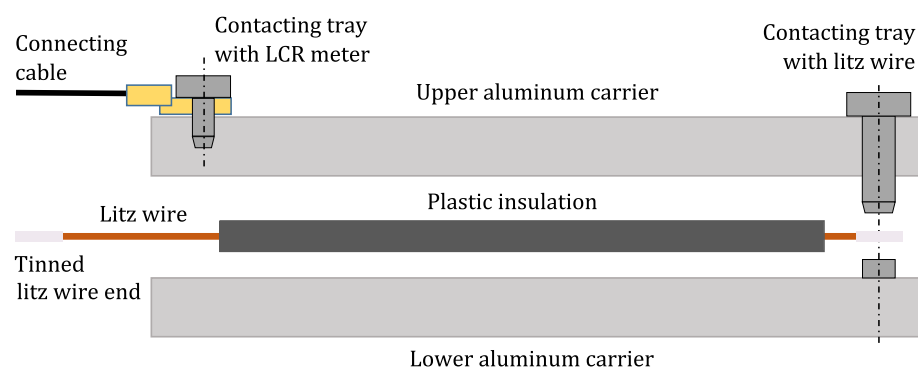
The properties of the litz wire used for this are shown in Table 4. The litz wire has a total diameter of 5 mm and is constructed from 1176 individual strands with a diameter of 0.1 mm. The insulation of the individual strands has grade 1.

Table 4. Properties of the litz wire used for coil module setup.

Cross Section	Number of Strands	Strand Diameter	Insulation
Ø 5 mm	1176	0.1 mm	Grade 1

5.1. Calculation of R_{litz}

For determining the losses due to DC, skin and internal proximity effects, an experimental setup is required in which the litz wire under investigation can be characterized without the influence of stray magnetic fields from the environment. For this purpose, a tube is formed from two joined aluminum carriers, each containing a groove. The test sample is inserted into the tube and positioned centrally by plastic spacers, which also provide electrical insulation. The test setup based on [31] is shown schematically in Figure 6. The aluminum carriers have a length of 1200 mm and the groove diameter is 8 mm. A screw at the end of the holder creates electrical contact between the carrier and the test sample. At the other end of the carrier, a screwed-on cable and the tinned end of the litz wire form the connection points for an LCR measuring device.

**Figure 6.** Test setup for characterizing a litz wire to determine the losses due to DC, skin effect and internal proximity effect based on [31].

The real part of the measured impedance represents the total resistance of the test sample R_{litz} and the test setup R_{setup} consisting of the aluminum carriers, screw connection and connection cable.

$$\text{Re}\{Z_{\text{meas}}\} = R_{\text{setup}} + R_{\text{litz}} \quad (31)$$

The frequency-dependent resistance of the experimental setup can be determined by a separate measurement using a solid circular conductor. Its resistance can be calculated analytically using the equations from Section 3.3. Based on the litz wire properties listed in Table 4, the resistances R_{id} for the ideally twisted litz wire and R_{par} for the litz wire with parallel strands can be calculated analytically using (26) and (27). Once R_{litz} is known from (31), the parameter λ_{skin} can be determined for the considered litz wire using (25). At a frequency of 85 kHz, this results in $\lambda_{\text{skin}} = 0.61$.

This allows the calculation of the resistance due to DC, skin and internal proximity effects of the coil module winding in the next step. Using the winding length of 7.8 m from Table 3 results in $R_{\text{litz}} = 33.2 \text{ m}\Omega$. This value remains constant for all three scenarios considered here. The corresponding resistance component depends only on the mechanical structure of the litz wire and is not influenced by surrounding components or neighboring conductors in the winding.

5.2. Calculation of $R_{\text{ext,prox}}$

An FEM simulation is performed to determine the resistance component due to the external proximity effect $R_{\text{ext,prox}}$. For this purpose, the coil module from Figure 2 is built

in ANSYS Maxwell 2018 R2 using the material parameters listed in Table 5 for electrical conductivity σ and relative permeability μ_r .

Table 5. Properties of the materials used in the FEM simulation in ANSYS Maxwell.

Material	Electrical Conductivity σ	Relative Permeability μ_r
Copper	58 MS/m	0.999991
Ferrite	0.01 S/m	2200
Aluminum	38 MS/m	1.000021

The magnetic field distribution is calculated with the “Eddy Current Solver” using adaptive mesh refinement and a convergence criterion of 1%. For current injection, the two ends of the winding are connected together and short-circuited. Here, a sinusoidal current with $I_{\text{rms}} = 1$ A is defined as a “stranded current” in the winding, which means that the current density is constant across the conductor cross-section.

As described in Section 4.3, the external magnetic field strength H_{ext} can be extracted using the polyline in the center of the conductor. For the three considered scenarios, Figure 7 shows the course of H_{ext} over the length of the winding from the inside to the outside.

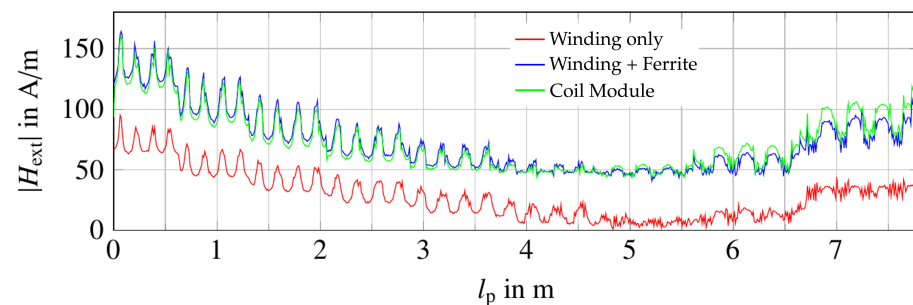


Figure 7. Course of the external magnetic field strength $|H_{\text{ext}}|$ along the polyline in the center of the conductor from the inside to the outside for winding only, winding with ferrite, and coil module.

The three courses of H_{ext} show that the external magnetic field strength reaches its maximum for the inner winding, then decreases successively before increasing again for the outermost winding. In addition, local peaks occur in the curve areas due to the bending radius. Using the relationship in (28), the resistance components can be calculated as $R_{\text{ext,prox}} = 1.5$ m Ω considering the winding only and $R_{\text{ext,prox}} = 7.6$ m Ω for both the combination of winding plus ferrite and for the entire coil module.

The effect of adding a flux guidance with ferrite on the external magnetic field strength can be seen in the rise from the red curve to the blue curve. Due to its high permeability, ferrite amplifies the entire magnetic field, resulting in a higher resistance value due to the external proximity effect $R_{\text{ext,prox}}$. Due to the closed ferrite layer, the aluminum shielding of this coil module has very little influence on the external magnetic field strength in the center of the conductor, leading to the same value for $R_{\text{ext,prox}}$ as without shielding.

5.3. Calculation of R_{eddy}

To calculate the resistance component R_{eddy} , the eddy current losses P_{eddy} of the two aluminum plates of the entire coil module from Figure 2 are calculated in the FEM simulation from Section 5.2 according to (29). For high accuracy, a suitable meshing must be ensured so that the mesh is sufficiently fine, especially in areas of high power loss density. To achieve this, three dummy layers are inserted underneath the surface of the aluminum on the side facing the magnetic field, spaced 0.1 mm apart. As a result of this action, the

mesh becomes finer in areas of higher eddy current density. Using the relationship in (30), the value of $R_{\text{eddy}} = 9.4 \text{ m}\Omega$ is calculated using the current excitation with $I_{\text{rms}} = 1 \text{ A}$.

5.4. Comparison of Modeling and Measurement

After calculating the individual resistance components, the comparison with the measured values is made in this section. For this purpose, three different scenarios are considered to verify the presented modeling method with the winding only, the combination of winding and ferrite, and the entire coil module consisting of winding, ferrite and shielding.

Starting from the left, Table 6 shows the three resistance components R_{litz} , $R_{\text{ext,prox}}$ and R_{eddy} . The sum of these values results in the modeled resistance R_{model} . Next to it the measured value R_{meas} and last the deviation ΔR between these values are listed. The Wayne Kerr 6500B impedance analyzer (Wayne Kerr Electronics Pvt. Ltd., Ghaziabad, India) was used as LCR meter for the measurements. The litz wire ends are tinned and do not have cable lugs, analogous to the characterization of the litz wire in the test setup in Figure 6.

Prior to calibration and measurements, the impedance analyzer was allowed to reach thermal steady-state conditions. Calibration was subsequently performed in accordance with the manufacturer's recommendations: open-circuit, short-circuit, and high-frequency compensation were carried out using the complete measurement fixture, including the four-terminal Kelvin clips. All measurements were conducted using the "Slow" measurement speed setting to maximize measurement accuracy and with an excitation voltage of 1 V. Only digits that remained stable throughout the measurement process are reported in Table 6. Repeated measurements were not documented as a formal measurement series and therefore no statistical uncertainty or standard deviation is available. This represents a limitation of the presented validation.

Table 6. Calculated resistance components with the presented modeling method and comparison with measurement results.

Coil Type	R_{litz}	$R_{\text{ext,prox}}$	R_{eddy}	R_{model}	R_{meas}	ΔR
Winding only	33.2 m Ω	1.5 m Ω	-	34.7 m Ω	34.1 m Ω	1.8%
Winding + Ferrite	33.2 m Ω	7.6 m Ω	-	40.8 m Ω	40.0 m Ω	2.0%
Coil module	33.2 m Ω	7.6 m Ω	9.4 m Ω	50.2 m Ω	49.8 m Ω	0.8%

The resistance component R_{litz} remains unchanged when ferrite or aluminum is added, as it depends only on the structure of the litz wire and is not influenced by the surrounding components. The situation is different for the resistance component due to the external proximity effect $R_{\text{ext,prox}}$. This resistance component increases when ferrite is used as a result of its field-amplifying effect, which depends on the size of the ferrite layer and the distance between it and the winding. In the shielding, the eddy currents generate an opposing field in the electrically conductive material, which leads to a weakening of the original field. Due to the closed ferrite layer, this effect is not noticeable in the coil module considered here, and the resistance component due to the external proximity effect remains constant.

The shielding does not further increase the resistance component $R_{\text{ext,prox}}$ in this example. However, it causes the resistance component R_{eddy} , which is newly introduced in this paper. Without taking this proportion into account, the deviation between the modeled and measured value would be around 18%, whereas by including it the deviation can be reduced to less than 1%. This shows the need to model the eddy current losses in

coil modules with expansive aluminum shielding and in compact structures with small distances between winding and shielding.

Finally, it should be noted that the presented methodology is verified by comparing the modeled and measured values for the three scenarios in Table 6. Differences of a maximum of 2% can result from deviations in the mechanical structure or measurement inaccuracies during the litz wire characterization.

6. Conclusions

In this paper, a practical method for modeling the electrical resistance of a coil module with flux guidance and shielding was presented. First, the fundamentals of general and frequency-dependent loss mechanisms in a conductor and a litz wire were investigated. Then, the individual loss mechanisms in a coil module with flux guidance and shielding, as well as their interactions, were analyzed in detail. As a result, it was found that the individual resistance components can be determined separately due to their orthogonality. After different approaches were discussed, a suitable calculation method was selected for each resistance component.

The resistance components due to DC resistance, the skin effect, and the internal proximity effect all depend on the mechanical structure of the litz wire and were therefore considered together as R_{litz} . The underlying idea is that the resistance of a real litz wire lies between the two extreme cases of an ideally twisted litz wire and a litz wire with individual strands arranged in parallel. For these two cases, the resistance can be calculated analytically. The resistance of the real litz wire is then modeled as a linear combination of these extreme cases using the quality parameter λ_{skin} , which is determined through a measurement-based characterization of the litz wire.

The resistance component due to the external proximity effect $R_{\text{ext,prox}}$ was determined using an FEM simulation. Here, use is made of the fact that the internal magnetic field strength is zero in the center of a circular conductor, which means that the external magnetic field strength can be evaluated there. With the help of a polyline in the center of the conductor, the external magnetic field strength can be extracted, and the resistance component due to the external proximity effect can be calculated.

The FEM simulation was also used to determine the resistance component due to shielding R_{eddy} . A new approach was presented for this purpose, in which the eddy current losses are simulated and then converted via the current excitation into an equivalent resistance component. Together with the established approaches used for the remaining resistance components, this enables a unified resistance model for complete EV-WPT coil modules including conductive shielding.

The presented modeling method was verified by considering three different scenarios: winding only; the combination of winding with ferrite layer; and, lastly, the entire coil module with winding, ferrite, and shielding. A comparison of the modeled and measured values showed good agreement between the results with deviations of 0.8%, 1.8%, and a maximum of 2%. The investigated coil module was deliberately chosen as a representative test case to validate the proposed methodology and to enable a systematic analysis of the individual loss mechanisms. Nevertheless, the model is not restricted to this specific geometry. Since it is based on the underlying physical loss mechanisms in the litz wire, ferrite, and conductive shielding, it can be applied to other wireless power transfer systems by adapting the corresponding geometric and material parameters. The presented validation demonstrates the accuracy of the proposed methodology for the investigated coil module. Additional validation using further WPT topologies and geometries is a subject for future work.

The primary contribution of this work is the integration of established analytical, numerical, and measurement-based approaches for the determination of individual resistance components into a unified model for the resistance of an entire coil module. In particular, the additional consideration of shielding-induced eddy current losses extends existing approaches and enables accurate resistance prediction for practical EV-WPT coil modules with conductive shielding.

Based on these results, the method is highly suitable to replace the prototyping process of the coil module and the repeated resistance measurements. This makes the design of a WPT system more time and cost efficient. Characterization of litz wires using the dimensionless parameter λ_{skin} also allows an evaluation of the quality of the litz wire structure, the identification of optimization potential, and the possibility of comparing different litz wires for an optimal choice. In addition to the precise analysis of the litz wire properties, the presented method also offers the possibility of splitting the coil module resistance into its individual components and thus enables a targeted optimization of the coil module design with parameter studies.

The presented method has been demonstrated and validated under the investigated small-signal measurement conditions. In principle, the method can also be applied to estimate losses at higher currents and during power transfer. In addition to calculating the losses from the resistance components discussed in this paper, the additional hysteresis losses in the ferrite that occur with increasing current amplitude need to be considered. However, the application of the method under rated-power operating conditions requires further experimental validation.

In the area around the coil module installed in the vehicle, other components consisting of electrically conductive material can influence the resistance and generate losses during power transfer. These effects can also be considered during the design and optimization process of the coil module and shielding for vehicle integration using the presented method.

In summary, modeling the resistance of a coil module with ferrite and shielding allows a prediction of the electrical properties of the coil module. By considering the loss mechanisms, a targeted optimization can be carried out without repeated coil module prototyping and resistance measurement series when using the presented method. This paper thus contributes to the optimal design and accelerated development of a wireless power transfer system.

Author Contributions: Conceptualization, T.L. and N.P.; methodology, T.L. and N.P.; validation, T.L.; investigation, T.L. and N.P.; writing—original draft preparation, T.L.; writing—review and editing, T.L.; visualization, T.L.; supervision, N.P. All authors have read and agreed to the published version of the manuscript.

Funding: This research received no external funding.

Data Availability Statement: The original contributions presented in this study are included in the article. Further inquiries can be directed to the corresponding author.

Conflicts of Interest: The authors declare no conflicts of interest.

References

1. Sagar, A.; Kashyap, A.; Nasab, M.A.; Padmanaban, S.; Bertoluzzo, M.; Kumar, A.; Blaabjerg, F. A Comprehensive Review of the Recent Development of Wireless Power Transfer Technologies for Electric Vehicle Charging Systems. *IEEE Access* **2023**, *11*, 83703–83751.
2. Mahesh, A.; Chokkalingam, B.; Mihet-Popa, L. Inductive Wireless Power Transfer Charging for Electric Vehicles—A Review. *IEEE Access* **2021**, *9*, 137667–137713.
3. Rayan, B.A.; Subramaniam, U.; Balamurugan, S.T. Wireless Power Transfer in Electric Vehicles: A Review on Compensation Topologies, Coil Structures, and Safety Aspects. *Energies* **2023**, *16*, 3084.

4. Parspour, N.; Maier, D.; Böttigheimer, M. Contactless energy transfer for charging electric and hybrid electric vehicles. In Proceedings of the 30th International Electric Vehicle Symposium (EVS 30), Brussels, Belgium, 9–11 October 2017.
5. Elbracht, L.; Noeren, J.; Parspour, N. A wireless multicoil charging system for low-voltage electric vehicle applications. *World Electr. Veh. J.* **2021**, *12*, 172.
6. Li, S.; Mi, C.C. Wireless Power Transfer for Electric Vehicle Applications. *IEEE J. Emerg. Sel. Top. Power Electron.* **2015**, *3*, 4–17.
7. Maier, D.; Parspour, N. Wireless Charging—A Multi Coil System for Different Ground Clearances. *IEEE Access* **2019**, *7*, 123061–123068.
8. Song, K.; Lan, Y.; Zhang, X.; Jiang, J.; Sun, C.; Yang, G.; Yang, F.; Lan, H. A Review on Interoperability of Wireless Charging Systems for Electric Vehicles. *Energies* **2023**, *16*, 1653.
9. Noeren, J.; Parspour, N.; Elbracht, L. An easily scalable dynamic wireless power transfer system for electric vehicles. *Energies* **2023**, *16*, 3936.
10. Triviño, A.; González-González, J.M.; Aguado, J.A. Wireless power transfer technologies applied to electric vehicles: A review. *Energies* **2021**, *14*, 1547.
11. Dimitriadou, K.; Rigogiannis, N.; Fountoukidis, S.; Kotarela, F.; Kyritsis, A.; Papanikolaou, N. Current Trends in Electric Vehicle Charging Infrastructure; Opportunities and Challenges in Wireless Charging Integration. *Energies* **2023**, *16*, 2057.
12. Ramakrishnan, V.; Savio, D.; C, B.; Rajamanickam, N.; Kotb, H.; Elrashidi, A.; Nureldeen, W. A Comprehensive Review on Efficiency Enhancement of Wireless Charging System for the Electric Vehicles Applications. *IEEE Access* **2024**, *12*, 46967–46994.
13. Kaewthep, C.; Venugopal, R.; Ramamurthi, S.; Palani, V.; Rao, C.H.N.N.; Hui, L.T.; Mahmud, S.M.H.; Al Awadh, M. Design and development of a high-efficiency sustainable wireless charging system for autonomous electric vehicles powered by renewable energy sources for remote locations. *Sci. Rep.* **2026**, 25018.
14. Farghly, A.; Mohamed, A.; Awad, H.; Abdelfatah, S.; Alharbi, M.A.; Alqarni, M.; Ali Abdelrazik, S. A Comprehensive Review of Wireless Power Transfer Techniques for Electric Vehicle Charging. *IEEE Access* **2025**, *13*, 199683–199718.
15. Doan, V.; Fujimoto, H.; Koseki, T.; Yasuda, T.; Kishi, H.; Fujita, T. Allocation of Wireless Power Transfer System From Viewpoint of Optimal Control Problem for Autonomous Driving Electric Vehicles. *IEEE Trans. Intell. Transp. Syst.* **2018**, *19*, 3255–3270.
16. Zhang, J.; Tang, T.; Yan, Y.; Qu, X. Eco-driving control for connected and automated electric vehicles at signalized intersections with wireless charging. *Appl. Energy* **2021**, *282*, 116215.
17. Campi, T.; Cruciani, S.; Maradei, F.; Feliziani, M. Magnetic field during wireless charging in an electric vehicle according to standard SAE J2954. *Energies* **2019**, *12*, 1795.
18. Lämmle, T.; Parspour, N.; Mönch, M. Analysis of the Magnetic Field Distribution in Inductive Power Transfer Systems under the Influence of Aluminum Shielding. In *Proceedings of the 5th International Conference on Smart and Sustainable Technologies (SpliTech)*; IEEE: New York, NY, USA, 2020; pp. 1–6.
19. Li, J.; Yin, F.; Wang, L. Transmission efficiency of different shielding structures in wireless power transfer systems for electric vehicles. *CSEE J. Power Energy Syst.* **2020**, *7*, 1247–1255.
20. Rylko, M.S.; Hartnett, K.J.; Hayes, J.G.; Egan, M.G. Magnetic Material Selection for High Power High Frequency Inductors in DC-DC Converters. In *Proceedings of the Twenty-Fourth Annual IEEE Applied Power Electronics Conference and Exposition*; IEEE: New York, NY, USA, 2009; pp. 2043–2049.
21. McDowell, F.M.E.; Damhuis, C.; Lawton, P.A.J.; Lin, F.J.; Covic, G.A. Litz Wire for High-Power Wireless Charging Systems. *IEEE J. Wirel. Power Technol.* **2026**, *1*, 0500213.
22. Wu, M.; Yang, X.; Cui, H.; Chen, W.; Wang, L.; Zhu, L.; Yu, X.; Yan, Z. Modeling of Litz-Wire DD Coil With Ferrite Core for Wireless Power Transfer System. *IEEE Trans. Power Electron.* **2023**, *38*, 6653–6669.
23. Sullivan, C.R.; Zhang, R.Y. Analytical model for effects of twisting on litz-wire losses. In *Proceedings of the IEEE 15th Workshop on Control and Modeling for Power Electronics (COMPEL)*; IEEE: New York, NY, USA, 2014; pp. 1–10.
24. Acero, J.; Carretero, C.; Lope, I.; Alonso, R.; Burdío, J.M. FEA-based model of elliptic coils of rectangular cross section. *IEEE Trans. Magn.* **2014**, *50*, 8401107.
25. Acero, J.; Alonso, R.; Burdío, J.M.; Barragán, L.A.; Carretero, C. A model of losses in twisted-multistranded wires for planar windings used in domestic induction heating appliances. In *Proceedings of the Twenty-Second Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*; IEEE: New York, NY, USA, 2007; pp. 1247–1253.
26. Carretero, C.; Acero, J.; Alonso, R. TM-TE decomposition of power losses in multi-stranded litz-wires used in electronic devices. *Prog. Electromagn. Res.* **2012**, *123*, 83–103.
27. Umetani, K.; Kawahara, S.; Acero, J.; Sarnago, H.; Lucía, Ó.; Hiraki, E. Analytical Formulation of Copper Loss of Litz Wire With Multiple Levels of Twisting Using Measurable Parameters. *IEEE Trans. Ind. Appl.* **2021**, *57*, 2407–2420.
28. Umetani, K.; Acero, J.; Sarnago, H.; Lucía, Ó.; Hiraki, E. Simple Fully Analytical Copper Loss Model of Litz Wire Made of Strands Twisted in Multiple Levels. In *Proceedings of the IEEE Applied Power Electronics Conference and Exposition (APEC)*; IEEE: New York, NY, USA, 2019; pp. 1257–1264.
29. Acero, J.; Hernandez, P.J.; Burdío, J.M.; Alonso, J.; Barragán, L.A. Simple resistance calculation in litz-wire planar windings for induction cooking appliances. *IEEE Trans. Magn.* **2005**, *41*, 1280–1288.

30. Roßkopf, A.; Bär, E.; Joffe, C. Influence of inner skin-and proximity effects on conduction in litz wires. *IEEE Trans. Power Electron.* **2013**, *29*, 5454–5461.
31. Rossmannith, H.; Doebröenti, M.; Albach, M.; Exner, D. Measurement and characterization of high frequency losses in nonideal litz wires. *IEEE Trans. Power Electron.* **2011**, *26*, 3386–3394.
32. Rosskopf, A. Calculation of Frequency Dependent Power Losses in Inductive Systems with Litz Wire Conductors by a Coupled Numeric Approach. Ph.D. Thesis, Friedrich-Alexander-Universität Erlangen-Nürnberg, Erlangen, Germany, 2018.
33. Acero, J.; Alonso, R.; Burdío, J.M.; Barragan, L.A.; Puyal, D. Frequency-dependent resistance in Litz-wire planar windings for domestic induction heating appliances. *IEEE Trans. Power Electron.* **2006**, *21*, 856–866.
34. SAE J2954; Wireless Power Transfer for Light-Duty Plug-in/Electric Vehicles and Alignment Methodology. SAE International: Warrendale, PA, USA, 2022.
35. IEC TS 61980-3; Electric Vehicle Wireless Power Transfer (WPT) Systems—Part 3: Specific Requirements for the Magnetic Field Wireless Power Transfer Systems. International Electrotechnical Commission: Geneva, Switzerland, 2022.
36. GB/T 38775.7; Electric Vehicle Wireless Power Transfer—Part 7: Interoperability Requirements and Testing—Vehicle Side. GB National Standard. Standards Press of China: Beijing, China, 2021.
37. Mühlethaler, J. Modeling and Multi-Objective Optimization of Inductive Power Components. Ph.D. Thesis, ETH Zurich, Zurich, Switzerland, 2012.
38. Biela, J. *Wirbelstromverluste in Wicklungen induktiver Bauelemente*; ETH Zürich: Zurich, Switzerland, 2011.
39. Ferreira, J. *Electromagnetic Modelling of Power Electronic Converters*; Springer Science & Business Media: Berlin/Heidelberg, Germany, 2013.
40. Lucht, P. *Transmission Lines and Maxwell's Equations*; Rimrock Digital Technology: Salt Lake City, UT, USA, 2014.
41. Albach, M. Two-dimensional calculation of winding losses in transformers. In *Proceedings of the IEEE 31st Annual Power Electronics Specialists Conference*; IEEE: New York, NY, USA, 2000; pp. 1639–1644.
42. Lammeraner, J.; Štafl, M. *Eddy Currents*; SNTL: Prague, Czech Republic, 1966.
43. Kazimierczuk, M. *High-Frequency Magnetic Components*; John Wiley & Sons: Hoboken, NJ, USA, 2009.
44. Ferreira, J.A. Analytical computation of ac resistance of round and rectangular litz wire windings. *IEEE Proc. B Electr. Power Appl.* **1992**, *139*, 21–25.
45. Sullivan, C.R. Optimal choice for number of strands in a litz-wire transformer winding. *IEEE Trans. Power Electron.* **1999**, *14*, 283–291.
46. Mühlethaler, J.; Kolar, J.W.; Ecklebe, A. Loss modeling of inductive components employed in power electronic systems. In *Proceedings of the 8th International Conference on Power Electronics-ECCE Asia*; IEEE: New York, NY, USA, 2011; pp. 945–952.
47. Stadler, A. Messtechnische Bestimmung und Simulation der Kernverluste in Weichmagnetischen Materialien. Ph.D. Thesis, Friedrich-Alexander-Universität Erlangen-Nürnberg, Erlangen, Germany, 2009.
48. Joffe, C. Modellbasierter Entwurf und Charakterisierung Eines Induktiven Ladesystems für Elektrofahrzeuge. Ph.D. Thesis, Friedrich-Alexander-Universität Erlangen-Nürnberg, Erlangen, Germany, 2019.
49. Hou, D.; Mu, M.; Lee, F.C.; Li, Q. New high-frequency core loss measurement method with partial cancellation concept. *IEEE Trans. Power Electron.* **2016**, *32*, 2987–2994.
50. Mu, M.; Li, Q.; Gilham, D.J.; Lee, F.C.; Ngo, K.D.T. New core loss measurement method for high-frequency magnetic materials. *IEEE Trans. Power Electron.* **2013**, *29*, 4374–4381.
51. Toshiyuki, K.; Guo, Y.; Shiozaki, K.; Xu, D.; Ngo, K. High current and high frequency planar inductor loss measurement and analysis. In *Proceedings of the Twenty-Eighth Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*; IEEE: New York, NY, USA, 2013; pp. 1689–1692.
52. Research Codes. Available online: <https://ryz.ece.illinois.edu/software.html> (accessed on 17 January 2024).
53. Barth, D.; Leibfried, T.; Cortese, G. Analytical calculation of the frequency-dependent litz wire resistance considering the wire connectors. In *Proceedings of the 21st European Conference on Power Electronics and Applications (EPE'19 ECCE Europe)*; IEEE: New York, NY, USA, 2019; pp. 1247–1253.
54. Klaus, B. Aspekte zur Praxisnahen Auslegung Induktiver Energieübertragungssysteme für Elektrofahrzeuge. Ph.D. Thesis, Karlsruher Instituts für Technologie (KIT), Karlsruhe, Germany, 2018.

Disclaimer/Publisher's Note: The statements, opinions and data contained in all publications are solely those of the individual author(s) and contributor(s) and not of MDPI and/or the editor(s). MDPI and/or the editor(s) disclaim responsibility for any injury to people or property resulting from any ideas, methods, instructions or products referred to in the content.