

An Improved Double-2D Electromagnetic Simulation Methodology for Winding Loss Analysis of High-Frequency Magnetic Components

Kaining FU, Jiaqing YE, Jiangtao TU, Yujie LAN, and Wei CHEN

Abstract—Precise and efficient winding loss calculation is critical for optimizing high-frequency magnetic components. While three-dimensional (3D) finite element method (FEM) provides accuracy, its computational demands limit practical applications. Conventional two-dimensional (2D) FEM, though computationally efficient, fails to balance calculation accuracy and efficiency. This paper develops a double-2D electromagnetic simulation methodology achieving 3D-equivalent accuracy through magnetic potential equivalence in the air gap location. The proposed double-2D approach can be applied to common core structures such as EE, EC, and others: (1) Synthesis modeling for symmetric structures. (2) Magnetic potential compensation for complex asymmetric structures. A deviation of less than 5% from the measured results is achieved over the frequency range of 10 kHz to 1 MHz, and a 130-fold acceleration was achieved, resolving the accuracy–efficiency trade-off in 3D FEM simulations. The proposed double-2D simulation technique provides practical significance for high-frequency magnetic device design, particularly in GaN-based power delivery.

Index Terms—Double-2D, FEM, magnetic component, magnetic potential equivalence, winding loss.

I. INTRODUCTION

WITH the rapid development of semiconductor devices, power converters toward high-frequency operation and high-power density, the optimization of losses in magnetic components (e.g., transformers and inductors) has emerged as a critical challenge for improving overall system efficiency. Winding loss, including both skin effect and proximity effect losses, significantly increases at high-frequency range, directly impacting device temperature rise, operational reliability, and system energy efficiency. Consequently, the development of high-

precision and computationally efficient loss analysis/optimization methodologies has become paramount for next-generation high-performance magnetic component design [1]–[2].

The evaluation methods of the magnetic component can be categorized into measurement, calculation, and simulation. The impedance measurement contains both the winding loss and core loss in the test results. While short-circuiting the magnetic core with an auxiliary winding eliminates core loss (by reducing the magnetizing flux to zero), it also introduces the auxiliary winding's own loss into the measured winding loss [3]. For the analytical methods, such as Dowell's model [4], Tourkhani method [5], and the equivalent area method [6], are widely used in winding loss calculations due to their high computational efficiency. However, the aforementioned analytical methods exhibit certain limitations. First, they assume a uniform magnetic field distribution, neglecting the geometric irregularities of actual windings (e.g., circular conductors), which necessitates the use of equivalent diameters to mitigate errors. Second, they only consider two-dimensional (2D) planar magnetic fields but neglect the vortex path in the conductor's axial (Z) direction.

FEM simulations can accurately resolve the complex electromagnetic field distributions in winding loss calculations. However, their engineering applications present significant computational demands [7]. These challenges include: First, three-dimensional (3D) modeling requires meshing with millions of grid elements (e.g., a Litz wire model in 1 MHz operation may require over 5 million mesh elements), resulting in single simulation durations extending to tens of hours on a 32-core high-performance computing (HPC) platform, coupled with memory requirements exceeding hundreds of gigabytes [8]. Such demands severely limit practicality in multi-parameter optimization studies [9]. Second, high-frequency refined modeling faces limitations: Submillimeter-level meshing in the skin depth direction (e.g., skin-layer elements $\leq 65 \mu\text{m}$ for 1 MHz copper conductors) is often compromised in real-world engineering due to computational constraints, introducing 10%–30% errors in field strength predictions. Despite advances in computational power, these solvers remain efficient only for specific 3D transformer configurations. They are unsuitable for versatile model generation and redefinition due to computational time and convergence issues. Although 2D FEM simulation has efficiency advantages in winding loss calculation, its dimensional reduction and high frequency adaptability limit its

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applicability in precision design and high frequency scenarios [10]–[14].

Magnetic components with one-dimensional (1D) or 2D field distributions can be accurately modeled using analytical expressions or numerical methods, respectively [15]. However, practical magnetic components frequently exhibit 3D field distributions, with toroidal and EE-shaped cores representing the most common configurations. Although the double-2D simulation methodology was proposed in [16]–[18] to overcome the above limitations, [18] proposed a 2D modeling to improve magnetic energy calculation accuracy for E-cores through reluctance equivalence. However, they ignored the influence of magnetic potential on the magnetic field distribution inside the core window of the magnetic components. The main contribution of this paper is to propose an improved double-2D simulation methodology, which further ensures the same magnetic potential as that a full 3D simulation. The proposed approach can account for 3D magnetic field distribution effects while utilizing computationally efficient 2D solvers.

This paper is structured as follows. Section II analyzes key discrepancies between conventional 2D model and actual 3D model, systematically identifying error sources inherent in 2D winding loss simulations. Section III introduces the fundamental principles of the double-2D simulation methodology, and proposes two kinds of the implementations for calculating winding loss aiming to different structural characteristics of magnetic components. Section IV conducts a comparative analysis of the accuracy and computational efficiency among 3D, 2D, and the proposed double-2D FEM simulation results. Section V provides the conclusions of this paper.

II. ERROR SOURCE ANALYSIS OF TRADITIONAL 2D SIMULATION

While 3D FEM simulations resolve the accuracy limitations of classical winding loss models (e.g., Dowell, Bessel-function approaches), they incur prohibitively high computational costs. Complex 3D magnetic component simulations consume more computational resources, resulting in impractical time requirement for design optimization processes. To mitigate these challenges, both industrial engineers and researchers frequently adopt simplified 2D models as substitutes for 3D simulations. Nevertheless, this approach introduces critical limitations, as single 2D representations fundamentally fail to capture the 3D field distribution characteristics of actual magnetic components.

A. Difference Between 3D and 2D Model

Fig. 1 illustrates a representative case using an EE-core structure, where the 3D magnetic component model is reduced to a single 2D model. Obviously, it can only describe the magnetic field distribution in the X - Y plane while entirely neglecting orthogonal field components in the Y - Z plane. The inherent discrepancy in field distributions between these orthogonal planes causes the 2D approximation to fundamentally fail in reflecting actual 3D flux distribution, consequently introducing substantial errors in winding loss

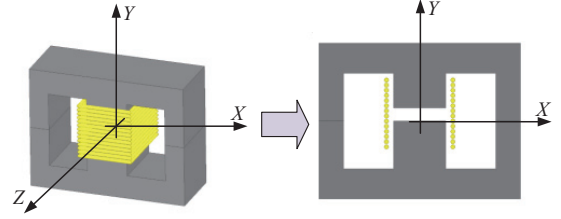


Fig. 1. Difference between 3D and 2D model.

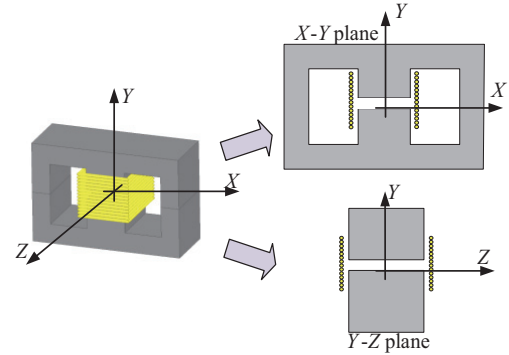


Fig. 2. Traditional 2D simulation methodology.

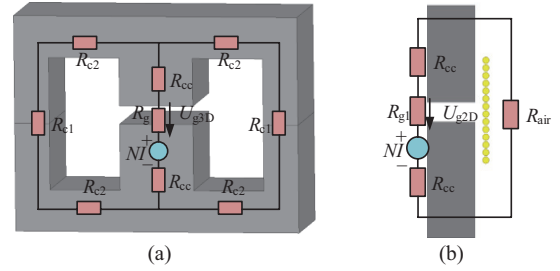


Fig. 3. Equivalent magnetic circuit model of a magnetic component with an air gap. (a) 3D model. (b) 2D model in Y - Z plane.

calculation.

To accurately calculate the winding loss, especially in the high frequency range, the magnetic field distribution both in X - Y and Y - Z planes should be considered simultaneously, as is shown in Fig. 2.

Based on the winding loss of 2D models in X - Y and Y - Z planes, the total winding losses can be expressed as:

$$P = P_{X-Y} \cdot L_{X-Y} + P_{Y-Z} \cdot L_{Y-Z} \quad (1)$$

where P_{X-Y} and P_{Y-Z} denote per-unit-length losses in the X - Y and Y - Z planes, with L_{X-Y} and L_{X-Z} representing corresponding winding lengths.

While the orthogonal planes (X - Y and Y - Z) in Fig. 2 provide intuitive visualization of magnetic core and winding geometries, this modeling approach introduces significant discrepancies in field characterization, particularly in the Y - Z plane configuration.

Fig. 3(a) and (b) illustrate the corresponding equivalent magnetic circuit models in 3D view and the Y - Z plane, respectively. The structural incompleteness of the Y - Z plane model (due

TABLE I
SPECIFICATIONS OF INDUCTOR

Core type	Number of turns	Core material	Wire diameter	Permeability	Air-gap length
EE28	14	PC95	0.6 mm	3300	1.8 mm

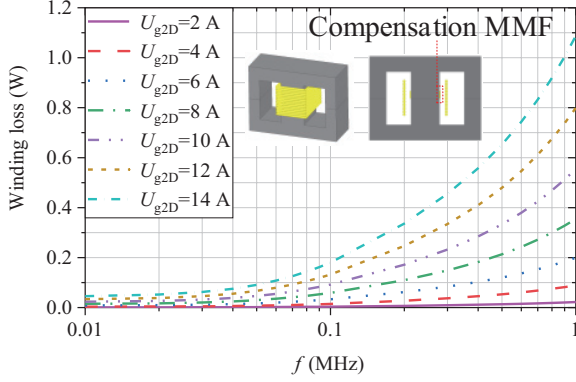


Fig. 4. Impact of magnetic potential drop on winding loss.

to the omission of the core side column) fundamentally alters flux circulation path. Specifically, the magnetic flux no longer circulates through the side column to the center column. It should be noticed that the X - Y / Y - Z splitting is an electromagnetic field plane decomposition rather than a magnetic circuit parallel connection. The 2D model shown in Fig. 3(b) omits magnetic reluctance information of the core side columns (R_{c2} - R_{c1} - R_{c2}). Therefore, the 2D simulation model fails to capture the parallel connection between R_{air} and R_{c2} - R_{c1} - R_{c2} , and only the series relationship of the magnetic reluctances R_{cc} - R_{g1} - R_{cc} - R_{air} is considered in the FEM analysis. In this scenario, the air reluctances satisfy the relationship: $R_{g1} \ll R_{air}$. Consequently, the major part of the magnetic potential drop is on the air reluctance R_{air} , with only a minor part is on the air gap reluctance R_{g1} . As a result, the magnetic potential drop U_{g2D} across the air gap reluctance R_{g1} in Y - Z plane is significantly lower than U_{g3D} .

B. Influence of Magnetic Potential Drop on Winding Loss

As shown in Table I, the inductor uses an EE28 core with a round conductor of 0.6 mm diameter wound in 14 turns. The core is made of PC95 ferrite material and features a 1.8 mm air gap in the central limb. An equivalent ampere-turn or called as compensation magnetomotive force (MMF) is applied to the air gap to alter the magnetic potential drop U_{g2D} , as shown in Fig. 5, in order to demonstrate its influence on winding loss. The principle of this implementation is discussed in Section III-B.

The air-gap magnetic potential drop dominates the magnetic field distribution inside the core window, which is the decisive factor for accurate winding loss simulation. As shown in the Fig. 4, the compensation air-gap magnetic potential U_{g2D} has a decisive impact on the winding loss simulation of the 2D model. At the same frequency, the winding loss increases with U_{g2D} , since U_{g2D} directly determines the magnetic field intensity at the

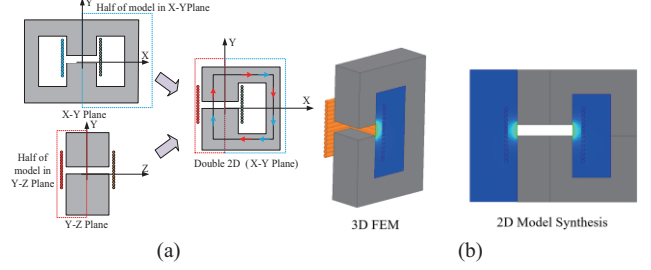


Fig. 5. 2D Model synthesis technique. (a) Synthesis scheme. (b) Magnetic equivalent.

air gap, which dominates the eddy current loss of the winding.

The model in Fig. 3(b) omits the side column, which causes the magnetic potential drop (U_{g2D}) to be lower than that in the practical magnetic component. The uncompensated 2D model ($U_{g2D} \ll U_{g3D}$) seriously underestimates the winding loss, especially at high frequencies. Only when U_{g2D} is accurately compensated to the real 3D air-gap magnetic potential U_{g3D} can the 2D model achieve the same simulation accuracy as the 3D model, which fully verifies the necessity of magnetic potential compensation.

III. PROPOSED DOUBLE-2D SIMULATION METHODOLOGY

The calculation of winding losses in magnetic components is significantly influenced by the distribution of the magnetic field. While 2D simulation offers superior computational efficiency compared to 3D simulation, it compromises accuracy as a result of model simplification. This simplification causes an inadequate representation of the magnetic field distribution in 2D simulation results, leading to significant errors in winding loss calculations.

To achieve an optimal balance between modeling accuracy and computational efficiency, this paper proposes a novel double-2D simulation approach implemented in a 2D eddy-current solver for precise winding loss calculation. As analyzed in Section II, the size of magnetic potential drop in the air gap significantly affects the magnetic field distribution. To establish the same magnetic field distribution as that in 3D simulations, it is essential to perform magnetic potential compensation in the air gap. Given the diversity of the magnetic core structure, the proposed double-2D simulation technique can be further classified into 2D model synthesis technique and magnetic potential compensation technique.

A. 2D Model Synthesis Technique

In this paper, the EE-core is used as the research object. As illustrated in Fig. 2, conventional 2D simulation model in Y - Z plane neglects the contribution of side columns on the main flux, leading to significant inaccuracies in representing the magnetic field distribution. To accurately construct the actual magnetic flux path in a 2D simulation model, this paper proposes a 2D model synthesis technique, as shown in Fig. 5. As shown in Fig. 5(a), the synthetic model is constructed in the X - Y plane, consisting of two components: half of the model in the X - Y plane and the other half in the Y - Z plane. It should be

noted that the Y - Z plane half of the model can be combined with the X - Y plane half to form a closed magnetic circuit, which is essential for calculating the winding loss in the Y - Z plane. When the Y - Z plane model is mapped onto the X - Y plane, the two half-models (from the X - Y and Y - Z planes) can be combined to construct a closed magnetic circuit for the synthetic model. Thus, the proposed synthetic model incorporates two components of winding loss: half from the X - Y plane and half from the Y - Z plane. Fig. 5(b) illustrates the comparison of magnetic field distribution between 3D FEM simulation and proposed doubled-2D simulation. It can find that there is the same magnetic field distribution. Therefore, the proposed 2D Model synthesis technique can accurately represent the magnetic field distribution of magnetic components as well as winding loss characteristics.

The winding losses of a half model of per unit length in X - Y and Y - Z plane can be calculated separately and named as P_{X-Y} and P_{Y-Z} . Then, the total winding losses of the whole magnetic component can be represented as:

$$P_{\text{loss}} = 2P_{X-Y} \cdot L_{X-Y} + 2P_{Y-Z} \cdot L_{Y-Z} \quad (2)$$

where L_{X-Y} and L_{Y-Z} denote the winding length in their respective orthogonal planes.

The proposed 2D synthesis follows three fundamental electromagnetic principles: orthogonal magnetic field superposition (decomposing 3D winding loss into linearly independent contributions from X - Y and Y - Z planes), airgap magnetic potential equivalence (enforced via a closed synthetic magnetic circuit to match the full 3D model), and loss superposition (summing the contributions from the two orthogonal planes).

Compared with 3D simulation technique, the 2D model synthesis technique significantly reduces memory resources and simulation time, enabling the simulation of winding losses of magnetic components to be calculated on standard computers. Compared with the traditional 2D simulation method, the proposed synthesis technique can construct the same magnetic field distribution as the 3D magnetic components, making the winding loss simulation results more accurate and reliable.

Since the synthesis model is constructed on the Cartesian coordinate system, the proposed technique is consequently only applicable to EE-type core structures that accommodate rectangular windings, and is unsuitable for EC-type cores requiring circular windings. Besides, the proposed 2D model synthesis technique is applicable only to magnetic components with a single air gap in center column. For the magnetic components with distributed air gaps or air gaps in side columns or the circular windings, the proposed 2D model synthesis technique is invalid. In these cases, the magnetic potential compensation technique should be used for calculating the winding losses.

B. Magnetic Potential Compensation Technique

As analyzed in Section II, the magnetic potential drop in the air gap affects the magnetic field distribution within the

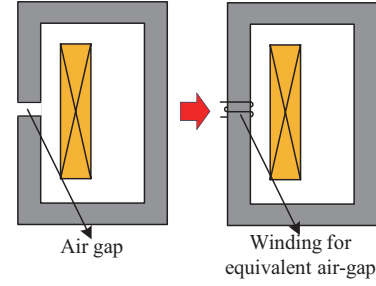


Fig. 6. Equivalent substitution diagram for air gap and its equivalent winding.

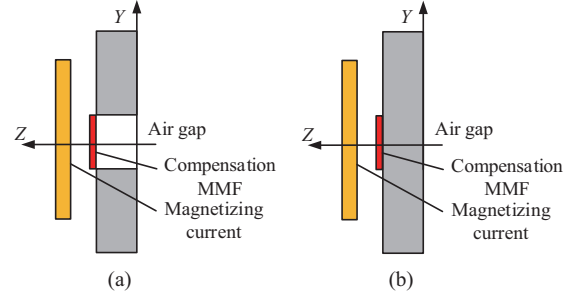


Fig. 7. Concept of magnetic potential compensation technique. (a) Add compensation MMF. (b) Remove air gap.

core window of the magnetic component. Therefore, a general double-2D simulation technique applying the magnetic potential compensation technique is proposed.

For gapped inductors, since the magnetic core exhibits high permeability, the MMF generated by the magnetizing winding is almost entirely borne by the air gap. Compared with air-gap-free transformers, it is clear that the opposing MMF originally borne by the secondary winding of the latter is completely dropped across the air gap of the gapped inductor. In other words, the air-gap magnetic voltage drop can be equivalent to the opposing MMF induced by the winding current. Based on this equivalent substitution concept, an equivalent air-gap winding can be designed. The opposing MMF from the current in this winding replaces the air-gap magnetic voltage drop, also referred to as the compensation MMF. The constructed winding for equivalent air-gap (WEAG) is wound at the position of the original air gap, with its width matching that of the air gap, as illustrated in Fig. 6.

As shown in Fig. 7(a), the magnetic potential drop across the air gap can be adjusted by introducing a compensation MMF at the air gap position, as illustrated in Fig. 7(b). Then, the air gap can be removed. Therefore, when the magnetic potential drop of the air gap in a 2D model differs from that in the actual magnetic component, a compensation current can be added at the air gap position to compensate the magnetic potential drop. This technique ensures consistency in the magnetic potential drop between 2D model and the actual magnetic component.

A comparison of the magnetic flux distribution between the magnetic component with an air gap and the component with applied magnetic potential compensation is presented in Fig. 8.

Visually, the two distributions are almost identical. In order

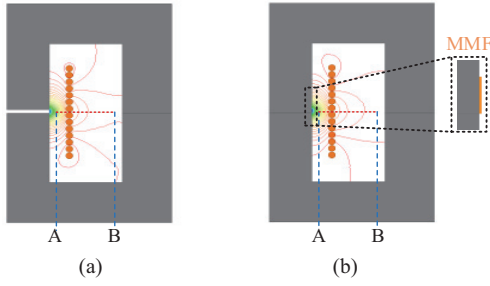


Fig. 8. Magnetic flux distribution of components. (a) Magnetic flux distribution in a component with an air gap. (b) Magnetic flux distribution with magnetic potential compensation applied.

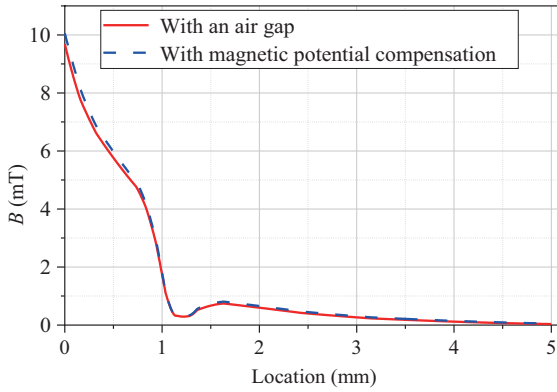


Fig. 9. Magnetic flux density along observation lines (AB).

to compare the magnetic field amplitude quantitatively, a profile was extracted along the dotted red line marked AB in Fig. 8. The results, shown in Fig. 9, demonstrate that the magnetic field exhibits the same variation trend along line AB for both cases.

In order to obtain the magnitude of the compensation current, it is necessary to determine the equivalent target magnetic potential drop in the air gap. In other words, the magnetic potential drop across the air gap of the actual magnetic component needs to be determined.

For single air-gap magnetic components, its magnetic potential drop is approximately equal to the magnetizing ampere-turns (NI). For multi-air-gap magnetic components, the magnetic potential drop of each air gap can be calculated via 3D magnetostatic FEM simulation. Since proposed method targets magnetic potential drop rather than air gap reluctance, which avoids solving air gap reluctance and thus do not need to separately consider the 3D flux effect of the air gap in the calculation process. Fig. 10 illustrates the influence of frequency on magnetic potential drop under eddy current solver. It can be vividly observed that the magnetic potential drop will not be influenced by the eddy current effect as frequency increases.

Within FEM simulation solvers, the eddy current solver can simultaneously account for the effect of eddy current and magnetic field, at the cost of significant computational resources, which can reflect the effect of parameters varying with frequency. However, the magnetostatic solver only considers the effect of magnetic field without eddy current. Since the eddy current effect does not affect the magnetic potential dis-

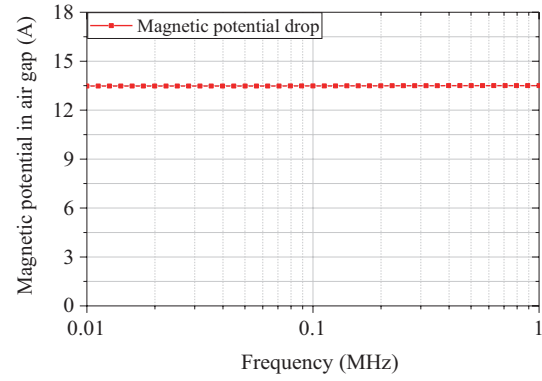


Fig. 10. Variation of magnetic potential drop in air gap versus frequency.

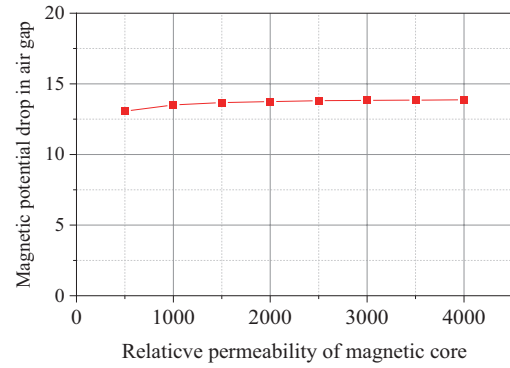


Fig. 11. Influence of permeability on magnetic potential drop of the air gap.

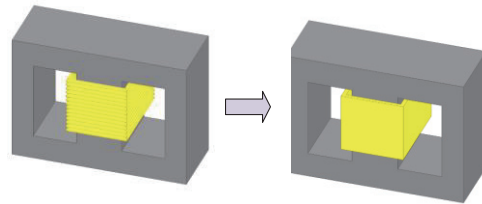


Fig. 12. Simplification of winding in magnetostatic solver.

tribution, the magnetostatic solver will be selected for higher computational efficiency.

The permeability of the magnetic core varies with frequency. Fig. 11 illustrated the variation tendency of magnetic potential drop versus permeability. Because the magnetic core permeability substantially exceeds that of air, the change in permeability caused by frequency variation does not affect the magnetic potential drop across the air gap. Therefore, the magnetic potential drop across the air gap does not be changed significantly with the variation of frequency.

The shape of winding structure only affects the winding loss characteristics instead of magnetic potential distribution. Therefore, the winding structure of magnetic components in the simulation can be simplified as shown in Fig. 12, where the ampere-turn and the height of winding remain the same.

Table II shows comparison results under different FEM solvers and model simplification methods. The simulation results demonstrate that the magnetic potential drop across

TABLE II
COMPARISON RESULTS OF MAGNETIC POTENTIAL DROP

	Magnetostatic solver	Eddy current (1 MHz)	Magnetostatic solver (simplified winding model)
magnetic potential drop/A	13.297	13.483	13.563
simulation time	28:56	34:12	00:56

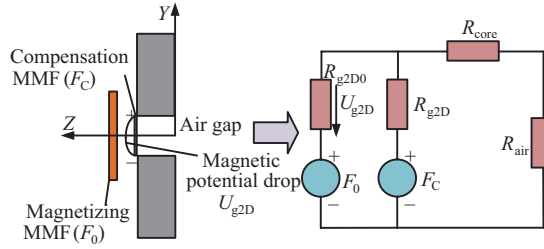


Fig. 13. Equivalent magnetic circuit model in Y - Z plane.

the air gap remains unaffected by eddy current and winding structure. Furthermore, the simplified winding structure can significantly improve the computational efficiency while maintaining accuracy.

Based on the 3D magnetostatic solver, the actual magnetic potential U_{g3D} drop across air gap can be obtained. Subsequently, the required compensation current can be determined by the magnetic circuit analysis. The 2D simulation model in Y - Z plane is used for the research object. Fig. 13 shows its corresponding equivalent magnetic circuit model, where the compensation MMF F_C is used to adjust the magnetic potential drop U_{g2D} of the air gap to match the 3D reference value.

The expression of magnetic potential drop across the air gap U_{g2D} for the 2D model in Y - Z plane can be given as:

$$U_{g2D} = F_0 \frac{R_{g2D0}}{(R_{core} + R_{air}) // R_{g2D} + R_{g2D0}} + F_C \frac{R_{g2D0} // (R_{core} + R_{air})}{(R_{core} + R_{air}) // R_{g2D} + R_{g2D0}} \quad (3)$$

where U_{g2D} is the magnetic potential drop in the air gap, F_0 and F_C are ampere-turn generated by the magnetizing current and compensation current, respectively.

The magnetic resistance R_{core} , R_{air} , R_{g2D0} , and R_{g2D} , are determined by the physical structure of the magnetic core and remain unaffected by the value of ampere-turn force F_C generated by compensation current. Thus, the expression in (3) can be further simplified as:

$$U_{g2D} = F_0 \cdot A_0 + F_C \cdot A_1 \quad (4)$$

where A_0 and A_1 are the coefficients of magnetic resistance.

In order to determine the coefficients of A_0 and A_1 , two compensation MMFs values, F_{C1} and F_{C2} are applied in the simulation to obtain the corresponding magnetic potential drops

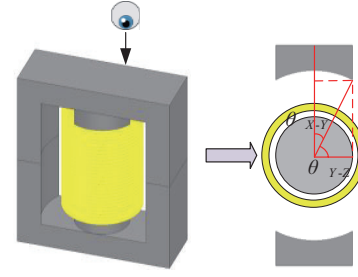


Fig. 14. Cross-sectional view of the magnetic components.

U_{g2D1} and U_{g2D2} . They satisfy the following relationship:

$$\begin{cases} U_{g2D1} = F_0 \cdot A_0 + F_{C1} \cdot A_1 \\ U_{g2D2} = F_0 \cdot A_0 + F_{C2} \cdot A_1 \end{cases} \quad (5)$$

According to (5), the coefficients A_0 and A_1 can be calculated as:

$$\begin{cases} A_0 = \frac{F_{C1} \cdot U_{g2D2} - F_{C2} \cdot U_{g2D1}}{F_0 \cdot (F_{C1} - F_{C2})} \\ A_1 = \frac{U_{g2D2} - U_{g2D1}}{F_{C2} - F_{C1}} \end{cases} \quad (6)$$

Based on the obtained target magnetic potential drop U_{g3D} via 3D simulation, the required compensation MMF F_C can be deduced as:

$$F_C = \frac{F_{C1}(U_{g2D2} - U_{g3D}) + F_{C2}(U_{g3D} - U_{g2D1})}{U_{g2D2} - U_{g2D1}} \quad (7)$$

According to (7), the compensation MMF F_C assigned in the double-2D simulation can be calculated.

Note that (2) is strictly applicable to magnetic components with rectangular windings in the central column, such as those with EE-core. For magnetic components with an EC-core, the winding in the central column is circular. The winding distribution across regions in the double-2D model must be defined by the magnetic component's cross-section, as shown in Fig. 14. In such cases, the proposed double-2D simulation technique should be conducted in cylindrical coordinate, with the winding loss given as:

$$P = 4(\theta_{X-Y} \cdot \frac{P_{X-Y}}{2\pi} + \theta_{Y-Z} \cdot \frac{P_{Y-Z}}{2\pi}) \quad (8)$$

Based on the above analysis, Fig. 15 presents the computational flowchart for determining the compensation MMF corresponding to the magnetic potential compensation technique.

The magnetic potential equivalence principle is universally applicable to rectangular/circular air-gap geometries and multi-gap structures, with no additional corrections or adjustments required.

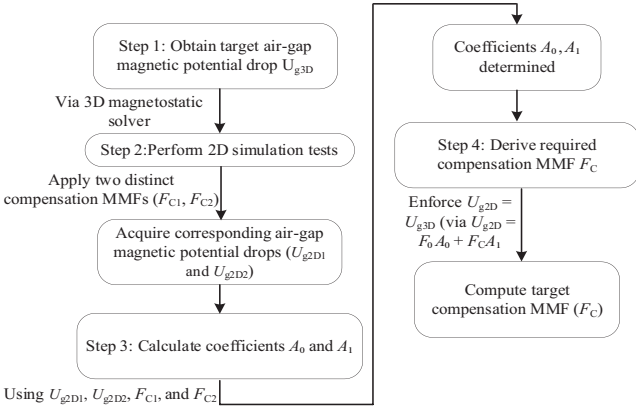


Fig. 15. Procedure for deriving compensation MMF for magnetic potential compensation technique.

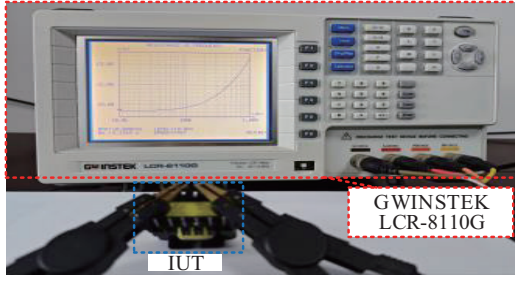


Fig. 16. Test platform for AC resistances.

IV. EXPERIMENTAL VERIFICATION

To validate the accuracy and computational efficiency of the proposed double-2D simulation technique, 3D simulation results serve as the reference benchmark. The magnetic components with EE-core and EC-core structures are selected as the research objects. Due to their structure characteristics, the winding losses are calculated using corresponding simulation strategy, i.e., 2D model synthesis technique and magnetic potential compensation technique. Besides, the magnetic components with distributed air gaps are further considered to verify the accuracy of the proposed double-2D simulation technique.

Using impedance analyzer to measure loss from the input terminals of the gapped inductor contains both the winding loss and the core loss. Hence, it is important to eliminate the influence of core loss on the measurement results. A winding loss measurement proposed in [19] can be adopted. Fig. 16 illustrates the test platform for AC resistances. Since the main-flux in the core is negligible, the measured loss of the inductor is winding loss without core loss.

The AC resistance R_{test1} measured from the test ports meets the following relationship:

$$R_{test1} = R_m + R_s \left(\frac{N_1}{N_2} \right)^2 = R_m + R_s^* \quad (9)$$

where R_s^* is the equivalent resistance of R_s transformed to the

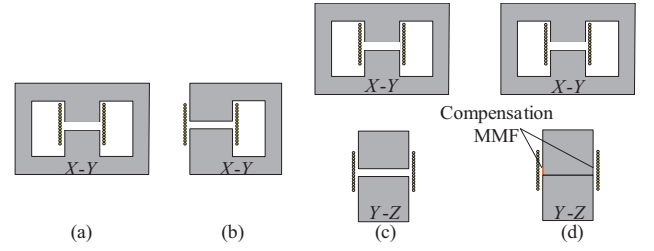


Fig. 17. EE-core inductor. (a) Traditional 2D. (b) 2D model synthesis. (c) 2D without magnetic potential compensation. (d) 2D with magnetic potential compensation.

primary side. Therefore, it is only necessary to subtract R_s^* from R_{test1} to obtain the value of R_m .

In this paper, a single-turn winding using a thickness of 0.065 mm copper foil is selected for winding equivalent air-gap (WEAG). All the simulations are conducted by Ansys maxwell with 3D and 2D solvers. The computational platform employed in this study is a Dell T640 server, with the following specifications: CPU: Intel Xeon Gold 5218 (2 units), RAM: 256 GB.

The selected frequency range (10 kHz–1 MHz) is based on the switching frequency range of typical switched-mode power supplies. At frequencies below 10 kHz, the skin effect and proximity effect have a negligible impact on winding losses. Consequently, the discrepancy between the AC resistance R_{ac} and DC resistance R_{dc} is minimal, eliminating the need for FEM simulation tools. At frequencies above 1 MHz, winding losses increase significantly, posing greater challenges to the optimization of core and winding designs. Furthermore, the research focus of this paper is inductors with lumped or distributed air gaps, which typically employ Mn-Zn ferrite materials with a cutoff frequency generally below 1 MHz.

It should be noticed that the single-layer winding structure is applied in the verification of EE-core with lumped air gap, EE-core with distributed air gaps, and EC-core with lumped air gap. Because multi-layer windings may introduce additional inconsistencies derived from practical winding processes, such as inter-layer gaps, winding tightness variation, and slight deviations in conductor positions. These manufacturing-induced uncertainties would bring extra errors to winding loss simulation. Therefore, single-layer windings are adopted to eliminate irrelevant process disturbances.

A. Inductors With EE-Core and a Lumped Gap

In order to verify the accuracy and computational efficiency of the proposed 2D model synthesis technique, the inductor with EE-core is used. The detail specifications the same as that listed in Table I. Fig. 17 depicts the corresponding simulation models for four approaches.

As demonstrated in Fig. 18, the proposed 2D model synthesis and magnetic potential compensation techniques, agree well with the measured results in the frequency range from 10 kHz to 1 MHz. There is a significant discrepancy between the results of 2D simulations without magnetic potential compensation and the experimental measurements. This is because the 2D model in the Y-Z plane without magnetic compensation yields

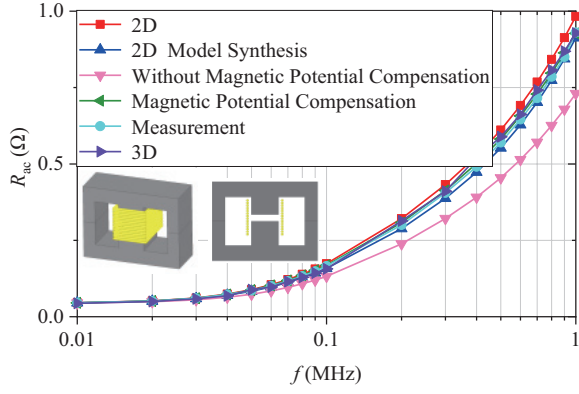


Fig. 18. Winding losses comparison results of magnetic components with a lumped air gap.

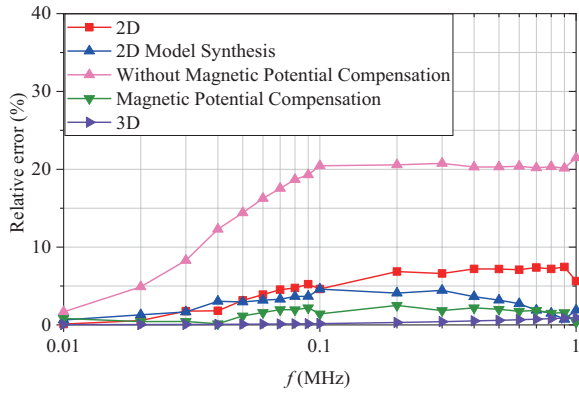


Fig. 19. Relative error of simulation results of magnetic components with a lumped air gap.

a small U_{g2D} value, which in turn leads to an underestimation of winding losses.

In order to calculate the relative error, the measured winding loss results are as the basement. Fig. 19 presents the relative errors between each simulation model—the traditional 2D model, 2D synthesized model, and 2D models (both with and without magnetic potential compensation)—and the experimentally measured winding loss results, over the frequency range of 10 kHz to 1 MHz.

When frequency exceeds 0.1 MHz, the traditional 2D method's relative error exceeds 5%. However, the proposed double-2D techniques—2D model synthesis and magnetic potential compensation—maintain high accuracy from 10 kHz to 1 MHz, with a maximum error below 5%. Although the error from the 2D model synthesis is higher than that from the magnetic potential compensation method, it remains below 5% across the frequency range of 10 kHz to 1 MHz. In contrast, the double-2D model without magnetic potential compensation demonstrates an error exceeding 20% across the frequency range from 100 kHz to 1 MHz. All simulation results are obtained for the designed air gap length (1.8 mm).

Table III shows the comparison results of simulation time and the amount of mesh. It can be found that both the 2D model synthesis and magnetic potential compensation techniques can significantly enhance the computational efficiency compared

TABLE III
CONSUMPTION OF SIMULATION TIME UNDER LUMPED AIR-GAP

Simulation method	2D	2D model synthesis	Magnetic potential compensation technique	3D-eddy current solver
Consumed Time	00:05	00:59	00:54	4:24:50
Mesh	14327	11633	10330	3927019

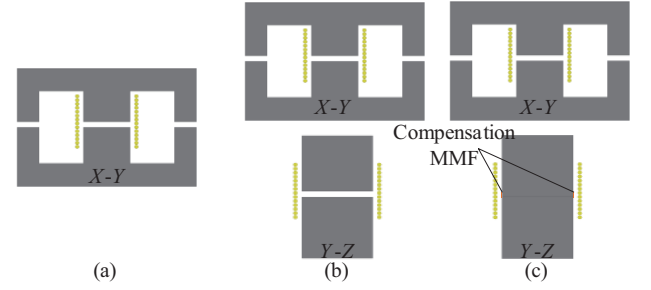


Fig. 20. EE-core inductor with distributed air gaps. (a) Traditional 2D. (b) 2D without magnetic potential compensation. (c) 2D with magnetic potential compensation.

with the 3D-eddy current solver simulations. The reason why 3D FEM simulation results are presented in these figures is to demonstrate that 3D FEM can accurately predict winding losses, but at the expense of increased simulation time. On the contrary, the proposed double-2D simulation technique can simultaneously maintain both computational efficiency and accuracy.

B. Inductors With EE-Core and Distributed Air Gaps

In practical designs, the air gaps in inductors increase fringe losses. Replacing them with distributed air gaps significantly mitigates these losses. As analyzed in Section III, the proposed 2D model synthesis technique is applicable only to magnetic components with EE-core and a single air gap. However, for the component with air gaps existing in both the side columns and the central column (1 mm gap length, other specifications matching Table I). For this configuration, winding losses were calculated using magnetic potential compensation technique. Fig. 20 depicts the corresponding simulation models for three approaches

Fig. 21 shows the winding losses from simulation results of an inductor with distributed air gaps. It can be found that the proposed magnetic potential compensation technique achieves high accuracy by ensuring the air-gap magnetic potential drop matches the measured results. Without proper compensation, such as the uncompensated simulation results, there is a larger discrepancy compared with measured results.

As shown in Fig. 22, which benchmarks against measured results, the proposed magnetic potential compensation technique exhibits a relative error below 5% across the entire frequency range from 10 kHz to 1 MHz. In contrast, both the traditional 2D method and the double-2D method (without magnetic compensation technique) indicate that the discrepancy remains below 10% only within the frequency range of 10 kHz

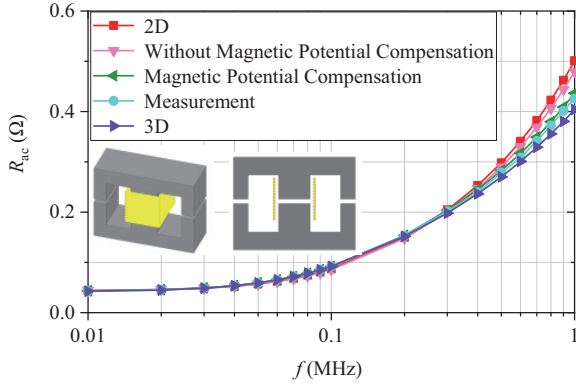


Fig. 21. Winding losses simulation in magnetic components with distributed air gaps.

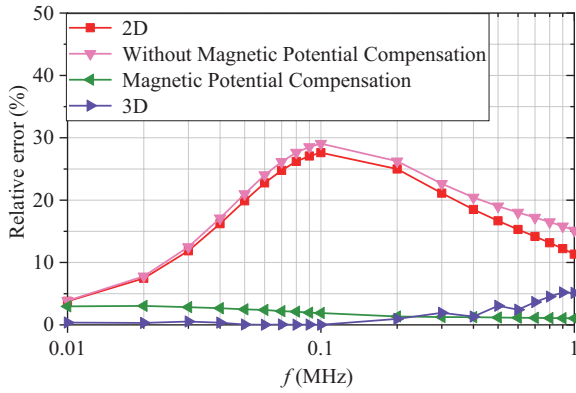


Fig. 22. Relative error in distributed air gaps magnetic Components: Simulation Results.

TABLE IV
CONSUMPTION OF SIMULATION TIME UNDER DISTRIBUTED AIR-GAP

Simulation method	2D	Magnetic potential compensation technique	3D-eddy current solver
Consumed Time	00:16	01:06	39:41:53
Mesh	9838	5703	4123946

to 25 kHz. In the higher frequency range, these methods cannot be applied because the relative error exceeds 20%.

Table IV shows the simulation time consumption for distributed air gaps. It can be found that implementing the magnetic potential compensation technique significantly improves the computational efficiency while maintaining equivalent accuracy.

C. Inductors With EC-Core and a Lumped Air Gaps

It has been verified that the proposed 2D model synthesis technique achieves high accuracy in winding loss calculation. However, this method is inapplicable to EI-core (due to edge column air gaps) or EC-core (which have circular windings). Furthermore, the proposed magnetic potential compensation technique has been verified on the EE-core. This technique is

TABLE V
SPECIFICATIONS OF INDUCTOR USING EC-CORE

Core -type	Number of turns	Core material	Wire diameter	Permeability	Air gap length
EC2834	28	PC44	0.6 mm	2300	1.0 mm

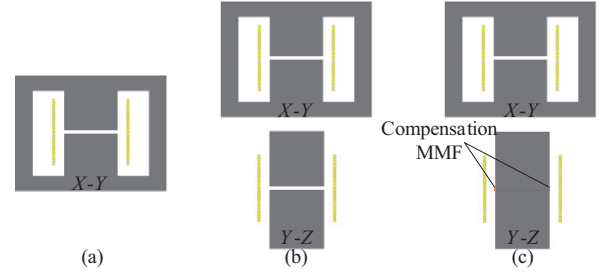


Fig. 23. EC2834-core inductor. (a) Traditional 2D. (b) 2D without magnetic potential compensation. (c) 2D with magnetic potential compensation.

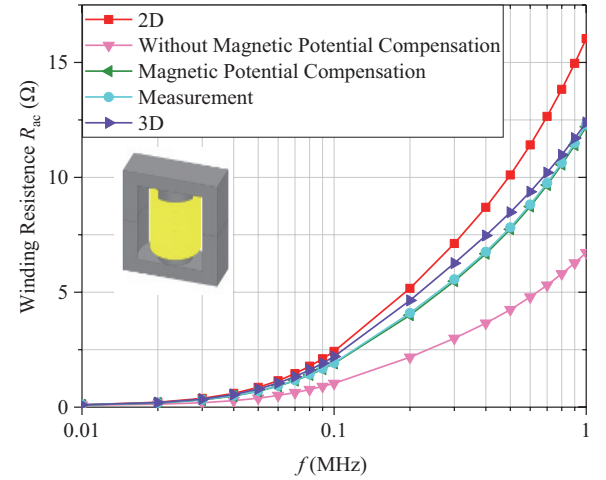


Fig. 24. Winding loss simulation in EC2834-core inductor.

further validated on EC-core magnetic components, with their specifications listed in Table V.

For the EC2834-core inductor, according to Fig. 14, the θ_{X-Y} is 30.35° , and the θ_{Y-Z} is 59.65° . Fig. 23 depicts the corresponding simulation models for three approaches. Fig. 24 presents the winding loss simulation results. The proposed technique demonstrates accuracy comparable to measured results, with Fig. 25 showing the relative error versus frequency. The simulation result verifies the feasibility of the double-2D simulation approach.

Similarly, Table VI compares the simulation consumed time and mesh count for EC-core model. The proposed technique significantly reduces winding loss computation time relative to 3D eddy-current solver benchmarks.

As the proposed double-2D method is fully implemented based on standard 2D electromagnetic simulations, it does not require any special mesh generation or boundary condition settings:

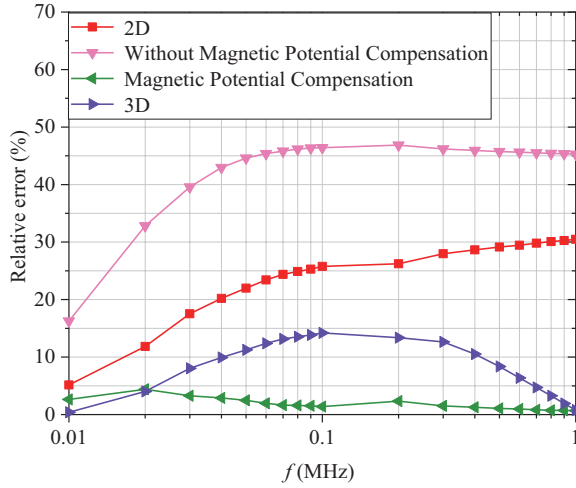


Fig. 25. Relative error in EC2834-core magnetic components: Simulation results.

TABLE VI
CONSUMPTION OF SIMULATION TIME UNDER EC-CORE

Simulation method	2D	Without compensation	Magnetic potential compensation technique	3D-eddy current solver
Consumed time	14:46	24:30	34:23	76:15:30
Mesh	113442	70280	170898	5012254

Mesh generation: The built-in adaptive meshing function in finite-element simulation software is fully sufficient to automatically refine key regions (windings, air gap) and ensure satisfactory simulation accuracy. No customized or extra-strict mesh rules are needed.

Boundary conditions: We only adopt the conventional balloon boundary that is widely used in 2D magnetic component simulations, which is a default and universal setting in commercial software.

V. DISCUSSION

The double-2D method is suitable for most magnetic integration structures including planar transformers and matrix transformers, while it has clear and manageable inherent limitations for complex magnetic integration structures.

If the Planar transformers feature standard planar symmetric structures and clear orthogonal field components (X - Y / Y - Z planes), the proposed double-2D method can be applied. The magnetic potential equivalence principle can be directly applied to calibrate the air-gap magnetic field, making this method particularly well-suited for planar magnetic integration structures, but the proposed 2D model synthesis technique cannot be applied due to the complex structure.

For matrix transformers with regular symmetric multi-core arrays, the method remains applicable. We only need to perform orthogonal plane decomposition for each sub-magnetic core and

implement magnetic potential compensation for each air gap separately. The core principle of magnetic potential equivalence still holds without additional corrections.

The method has two inherent topological constraints:

1) It relies on orthogonal 2D plane decomposition, so it is less friendly to extremely asymmetric or non-orthogonal magnetic integration structures with no clear decomposable planes.

2) The proposed 2D model synthesis technique is limited to some simple core structures; only the magnetic potential compensation technique can be used for complex integrated structures, which requires accurate extraction of 3D magnetic potential drops for multiple gaps.

VI. CONCLUSION

This paper proposes a double-2D simulation technique for winding loss analysis in magnetic components. The key conclusions are:

1) The air-gap magnetic potential drop significantly influences winding loss characteristics.

2) The 2D model synthesis technique can only be applicable to the EE-core with an air gap in the central column.

3) The proposed magnetic potential compensation approach significantly expands applicability while maintaining superior accuracy (within 5% deviation) compared to 2D model synthesis technique.

4) The double-2D simulation technique successfully combines computational efficiency (reducing simulation time by 99.2% relative to 3D-FEM benchmarks) with high accuracy through innovative 2D models decomposition strategy.

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