

An Improved Two-Dimensional Calculation Method for PCB Winding Loss of Planar Transformers

Jinjiang LI, Binzhi XIE, Wei CHEN, and Qingbin CHEN

Abstract—Accurate winding-loss calculation is critical for the design of printed circuit board (PCB) planar transformers, especially for considering the edge effects of the PCB winding. In conventional winding-loss modeling, the winding is uniformly divided into a limited number of segments due to computational constraints. However, this approach fails to accurately represent the nonuniform current distribution within the winding under high-frequency operating conditions. To address this issue, an improved winding segmentation algorithm is proposed, based on an iterative procedure. Each PCB turn of the winding is modeled as multiple parallel-connected conductors with nonuniform dimensions, and the segmentation size is adaptively determined according to the actual current distribution. An 8-turn planar transformer is used as a sample, and the winding-loss calculation results are validated against those obtained from two-dimensional (2-D) finite element method (FEM). The results indicate that the proposed algorithm reduces the maximum winding-loss calculation error from 2.9% to 0.22% compared with conventional uniform segmentation methods, in which each turn is divided into eight segments for calculation. The experiment results show that the winding AC resistance is lower than 8% up to 1 MHz, which verifies that the proposed method is correct and flexible.

Index Terms—Improved winding segmentation, optimisation design, planar magnetic components, winding loss model.

I. INTRODUCTION

HIGH-frequency switched-mode power supplies (SMPS) are widely used in modern power electronic systems due to their high efficiency and compact size. As power density continues to increase, planar magnetic components have become a key technology due to their low profile, improved thermal performance, and favourable electromagnetic compatibility. However, under high-frequency operation, pronounced skin, proximity, and edge effects lead to highly nonuniform current distribution in printed circuit board (PCB) windings, resulting in a significant increase in winding losses. Accurate winding-

loss modeling is therefore essential for the optimal design of planar transformers in high-frequency SMPS [1]–[3].

Traditional winding-loss calculation methods were primarily based on one-dimensional (1-D) Dowell analytical models [4], [5]. Based on the classical Dowell model, [6]–[10] demonstrated that skin and proximity effects in multilayer windings can be treated independently using Bessel functions, which can improve loss prediction accuracy. Nevertheless, these 1-D models are primarily applicable to round conductors and assume a uniform current distribution across the conductor width, which is parallel to the direction of the external magnetic field, and this is different from the real situation. Therefore, they limit the accuracy of planar transformer applications.

To address these limitations, finite element method (FEM) has been extensively used to investigate current density and loss distribution in planar windings. FEM results in [11] showed that, at high frequencies, the current tends to concentrate near the edges of rectangular conductors, a phenomenon known as the edge effect. Based on this observation, [12] introduced fitted parameters into 1-D models to account for edge effects and improve prediction accuracy. However, such approaches rely heavily on numerical fitting and lack generality. Subsequently, several 2-D winding-loss models were proposed, considering magnetic field distribution and edge effects [13]–[15]. Although these models improved accuracy, they either depend strongly on numerical computation or require extensive simulation data, resulting in increased computational complexity and limited practical applicability.

An effective alternative is to segment the winding into multiple subconductors and calculate the loss of each segment individually. In [16], a 2-D winding-loss model based on uniform winding segmentation was presented, which significantly enhanced estimation accuracy. However, uniform segmentation cannot accurately capture the highly nonuniform current distribution in planar windings at high frequencies.

To address the above limitations, this paper proposes an iterative improvement of the winding segmentation algorithm for PCB windings in planar transformers. In Part II, a winding-loss calculation method for parallel-connected planar windings is introduced, and a PCB winding-loss model considering edge effects is derived. In Part III, an improved iterative segmentation algorithm is developed, in which the winding is modeled as multiple parallel-connected conductors with nonuniform widths. The segmentation density is adaptively adjusted according to the distribution of magnetic field strength along the winding, enabling a more accurate representation of the

Manuscript received January 23, 2026; revised March 20, 2026; accepted April 13, 2026. Date of publication September 30, 2026; date of current version July 13, 2026. This work was supported in part by National Natural Science Foundation of China under the grant 51407032 and the Natural Science Foundation of Fujian Province under the grant 2019J01251. (Corresponding author: Qingbin Chen.)

All authors are with the Magnetics and Power Conversion Laboratory, the College of Electrical Engineering and Automation, Fuzhou University, Fuzhou 350108, China (e-mail: leejj@fzu.edu.cn; 1363128747@qq.com; fzdxcwh@qq.com; cqib@fzu.edu.cn).

Digital Object Identifier 10.24295/CPSSSTPEA.2026.00017

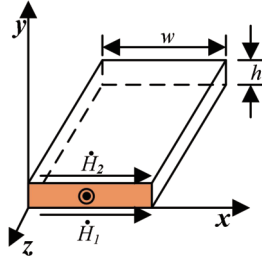


Fig. 1. 1-D Dowell model.

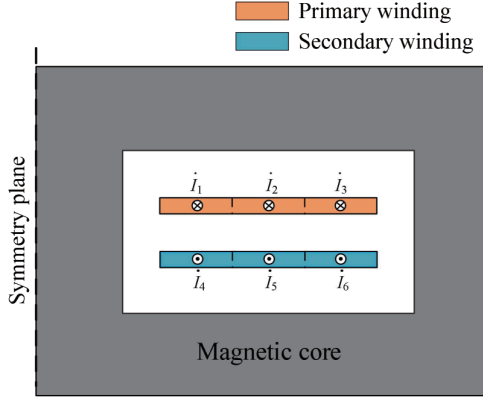


Fig. 2. Each coil winding is evenly segmented into three parallel parts.

actual current distribution. In Part IV, the effectiveness of the proposed algorithm is verified through experimental measurements on an 8-turn planar transformer and 2-D FEM simulations, demonstrating improved accuracy with the same number of segments.

II. BASIC PRINCIPLES OF WINDING LOSSES AND LIMITATIONS OF EXISTING UNIFORM SEGMENTATION METHODS

A. Traditional Calculation Methods for Winding Losses

In Fig. 1, the theoretical calculation of winding losses is based on the Dowell model, which employs a 1-D eddy-current formulation. In this model, the original three-dimensional (3-D) electromagnetic field problem is reduced to a 1-D form and analysed using Maxwell's equations. By calculating the magnetic field intensities $\dot{H}_1$ and $\dot{H}_2$ on both sides of a conductor layer, the current density distribution within a rectangular copper-foil winding can be expressed as

$$\dot{J}_z(y) = \frac{d\dot{H}_x(y)}{dy} = \frac{\dot{H}_1 \cosh[\alpha(h-y)] - \dot{H}_2 \cosh(\alpha y)}{\sinh(\alpha h)} \quad (1)$$

The corresponding expression for calculating the loss in a rectangular copper-foil winding under sinusoidal current excitation at an arbitrary frequency is given by

$$P = \frac{MLT \cdot w}{2\sigma} \int_0^h |\dot{J}_z(y)|^2 dy \quad (2)$$

where σ denotes the electrical conductivity of the copper foil,

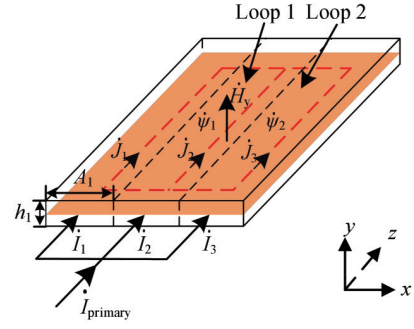


Fig. 3. Each variable of the primary winding conductor in Fig. 2.

MLT is the mean length per turn of the winding, h is the copper thickness, and w is the winding width.

For the analysis of current distribution in parallel-connected conductors, a planar transformer with a primary-to-secondary turn ratio of 1:1 is considered, as shown in Fig. 2. For simplicity, both the primary and secondary windings are each segmented into three parallel-connected conductors of equal width.

Since the winding thickness is much smaller than its lateral width, longitudinal segmentation cannot accurately capture the local current crowding and magnetic field distortion within the winding. In contrast, the current and magnetic field distributions along the horizontal (transverse) direction of a single winding exhibit strong symmetry. Therefore, by utilizing the structural symmetry of the transformer, a unified transverse segmentation strategy can be applied to all windings, which effectively reduces the computational cost in engineering analysis.

[17] presents a generalised modeling approach for calculating the losses of parallel PCB windings. Taking the primary winding shown in Fig. 3 as an example, the currents in the equivalent parallel-connected sub-conductors of each turn are calculated. A rectangular dashed contour passing through the centre points of two adjacent equivalent parallel sub-conductors is assumed as the closed loop. Since three turns are connected in parallel through the sub-conductors, there are three unknown currents. Accordingly, two independent voltage-loop equations and one node current equation, based on Kirchhoff's laws, are required to solve for the currents in the three parallel-connected subconductors.

The current equation is given by

$$\dot{I}_1 + \dot{I}_2 + \dot{I}_3 = \dot{I}_{\text{primary}} \quad (3)$$

And the corresponding voltage-loop equations for Loops 1 and 2 can be expressed as

$$\begin{cases} \frac{\dot{J}_1}{\sigma} - \frac{\dot{J}_2}{\sigma} = j2\pi f \dot{\psi}_1(\dot{I}_1, \dot{I}_2) \\ \frac{\dot{J}_2}{\sigma} - \frac{\dot{J}_3}{\sigma} = j2\pi f \dot{\psi}_2(\dot{I}_2, \dot{I}_3) \end{cases} \quad (4)$$

where the conductor length is assumed to be 1 m, $\dot{I}_{\text{primary}}$ denotes the total primary current, $\dot{\psi}_1$ and $\dot{\psi}_2$ represent the magnetic fluxes enclosed by the assumed closed loops, and $\dot{J}_j$ denotes the current density of the j -th parallel conductor. The relationship

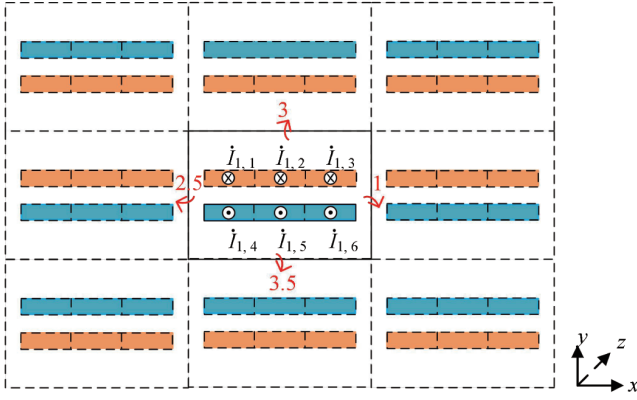


Fig. 4. Mirror image model of a transformer.

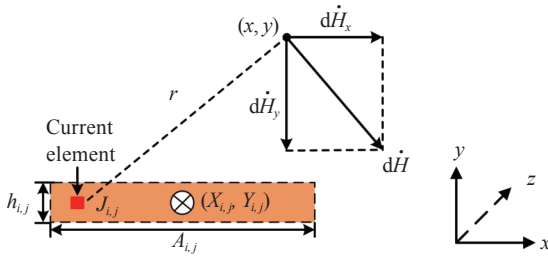


Fig. 5. Magnetic field intensity of a small conductor at an arbitrary point in space.

between current density and conductor geometry is given by

$$\dot{J}_j = \frac{\dot{I}_j}{h_j A_j}, \quad j = 1, 2, 3 \quad (5)$$

where h_j and A_j denote the thickness and width of the j -th conductor, respectively.

For magnetic cores with high permeability, the core boundary can be eliminated by applying the image method, which is based on the uniqueness theorem of electromagnetic fields. This method replaces the magnetic core boundary with an equivalent image winding, thereby removing the core boundary from the analytical model. The final form of this modified model is shown in Fig. 4.

The resulting magnetic field distribution produced by one part of a single turn is illustrated in Fig. 5.

According to Biot-Savart's law, the magnetic field intensity at an arbitrary point within the core window can be expressed as

$$\dot{\mathbf{H}} = \dot{H}_x(x, y) \cdot \mathbf{e}_x + \dot{H}_y(x, y) \cdot \mathbf{e}_y \quad (6)$$

where the horizontal and vertical field components are given by

$$\dot{H}_x(x, y) = \sum_{j=1}^6 \sum_{i=1}^9 \int_{Y_{i,j}-h_{i,j}/2}^{Y_{i,j}+h_{i,j}/2} \int_{X_{i,j}-A_{i,j}/2}^{X_{i,j}+A_{i,j}/2} \frac{\dot{I}_{i,j}}{2\pi h_{i,j} A_{i,j}} \frac{y-Y}{(x-X)^2 + (y-Y)^2} dXdY \quad (7)$$

$$\dot{H}_y(x, y) = \sum_{j=1}^6 \sum_{i=1}^9 \int_{Y_{i,j}-h_{i,j}/2}^{Y_{i,j}+h_{i,j}/2} \int_{X_{i,j}-A_{i,j}/2}^{X_{i,j}+A_{i,j}/2} \frac{\dot{I}_{i,j}}{2\pi h_{i,j} A_{i,j}} \frac{x-X}{(x-X)^2 + (y-Y)^2} dXdY \quad (8)$$

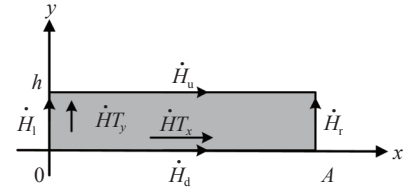


Fig. 6. 2-D winding loss model for a part of a single-turn rectangular conductor.

Fig. 6 shows the 2-D winding loss model for a part of a single-turn rectangular conductor, where $\dot{H}_u$ and $\dot{H}_d$ denote the magnetic field intensities at the upper and lower boundaries, respectively, $\dot{H}_l$ and $\dot{H}_r$ denote those at the left and right boundaries, and $\dot{H}_T$ and $\dot{H}_x$ represent the horizontal and vertical components of the magnetic field intensity within the conductor.

Based on the magnetic field intensities at the four boundaries of each conductor, the current density distribution within a parallel conductor can be expressed as

$$\begin{aligned} jT_z(x, y) &= \frac{d\dot{H}_T(x)}{dx} - \frac{d\dot{H}_T(y)}{dy} \\ &= \alpha \frac{\dot{H}_d \cosh[\alpha(h-y)] - \dot{H}_u \cosh(\alpha y)}{\sinh(\alpha h)} \\ &\quad - \alpha \frac{\dot{H}_l \cosh[\alpha(A-x)] - \dot{H}_r \cosh(\alpha x)}{\sinh(\alpha h)} \end{aligned} \quad (9)$$

Accordingly, the PCB winding loss of a planar transformer, considering edge effects, can be calculated as

$$P = \sum_{j=1}^{N_{\text{parallel}}} \frac{MLT}{2\sigma} \int_0^{h_{i,j}} \int_0^{A_{i,j}} |jT_{z-j}(x, y)|^2 dx dy \quad (10)$$

where jT_x denotes the current density at an arbitrary point within the conductor, and N_{parallel} denotes the total number of equivalent parallel-connected conductors within the winding.

When the transformer operates under non-1:1 turns-ratio conditions, the proposed method can still effectively calculate winding loss, since key parameters such as the core and winding structures, and the current excitation in the model, can be clearly determined.

For non-sinusoidal current conditions, the primary and secondary currents can be decomposed into a DC component and harmonic components using Fourier decomposition. Based on the orthogonality of winding losses at different frequencies, the DC loss and the harmonic losses can be calculated separately and then summed to obtain the total winding loss:

$$P_T = I_0^2 \cdot R_{dc} + I_1^2 \cdot R_{1dc} + I_2^2 \cdot R_{2dc} + \dots + I_k^2 \cdot R_{kac} \quad (11)$$

where I_k represents the root mean square (RMS) value of the DC component and the k -th harmonic component of the primary current obtained through Fourier decomposition, and R_{kac} is the equivalent AC resistance of the winding at the k -th harmonic frequency referred to the primary side.

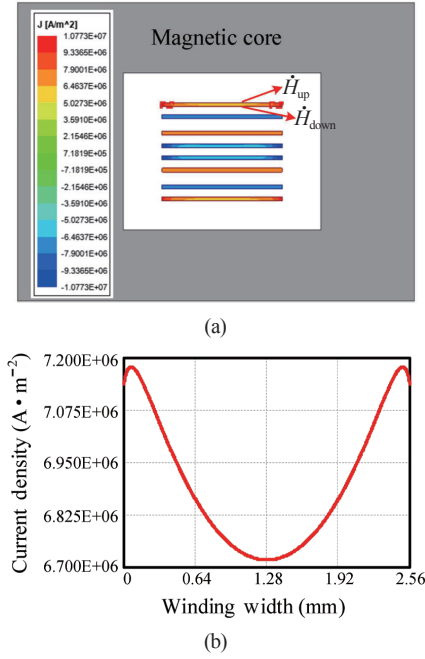


Fig. 7. Current-density characteristics of planar transformer windings. (a) 2-D Simulation model. (b) Current density distribution along the winding width.

B. Effect of Edge Phenomena on Winding Segmentation

At high frequencies, skin and proximity effects cause the magnetic field parallel to the winding thickness direction to induce eddy currents near conductor edges, resulting in current crowding. This edge-induced current crowding, in turn, leads to a pronounced deviation between 1-D analytical results and 2-D numerical simulations.

Under high-frequency conditions, due to the existence of a gap between the planar transformer winding and the magnetic core, the edge effect causes the magnetic field intensity near the winding edges to be significantly higher than that in the central region. Moreover, the magnetic field distribution near the edges is highly distorted with large gradient variations, leading to significant current crowding. In contrast, the central region of the winding is less affected by edge effects, with a relatively stable magnetic field distribution and a more uniform current density distribution. As shown in Fig. 7, the current density is mainly concentrated in the edge regions of the winding, while the current density in the central region is significantly lower.

Given this distribution difference, the traditional uniform segmentation strategy cannot capture the local variations in the magnetic field. As shown in Fig. 8, when the number of segments is small, the segmentation size in edge regions becomes too large, leading to insufficient accuracy in loss calculation and an inability to capture local high-loss regions. Although increasing the number of uniformly segmented conductors can improve calculation accuracy, excessively fine segmentation in the central region can introduce redundant segments, increase the number of iterations, and reduce the algorithm's computational efficiency. Therefore, uniform segmentation is not suitable for practical engineering applications, and an improved winding segmentation strategy is required to achieve both high

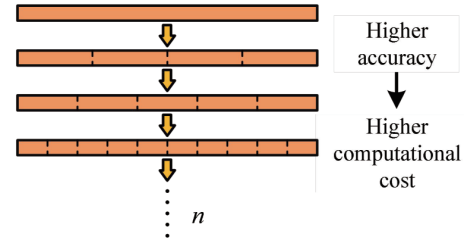


Fig. 8. Method for segmenting uniform windings.

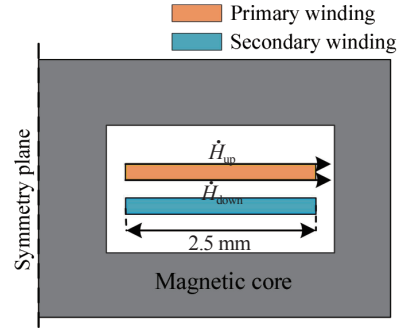


Fig. 9. Two-turn planar transformer.

accuracy and computational efficiency.

III. IMPROVED ITERATIVE WINDING SEGMENTATION ALGORITHM

A. Magnetic Field Analysis at the Winding Boundaries

To analyze the effect of current magnitude variation on the magnetic field, a simulation model is established using a planar transformer with a 1 : 1 primary-to-secondary turns ratio, as shown in Fig. 9. In this model, identical sinusoidal current excitations are applied to both the primary and secondary windings. For subsequent analysis, the magnetic field strength along the upper and lower boundaries of the primary winding is extracted.

The corresponding simulation results are presented in Fig. 10. It can be observed that increasing the excitation current leads to a proportional increase in the magnetic field magnitude, while the distribution pattern remains nearly unchanged. This indicates that the boundary magnetic field distribution is independent of the current amplitude. Furthermore, FEM simulations reveal that the magnetic field distribution varies with frequency, which may result in different winding segmentation outcomes at different frequencies. However, when the winding segmentation results obtained at the fundamental frequency are adopted for higher harmonic frequencies, the resulting calculation accuracy remains acceptable, thereby reducing the overall computation time.

To ensure the accuracy of the computational model, the magnetic field intensity along the same boundary of the parallel subconductors within each turn should exhibit a linear distribution with a small variation in magnitude. The uniform segmentation algorithm clearly fails to satisfy this criterion. Therefore, an improved winding segmentation algorithm based

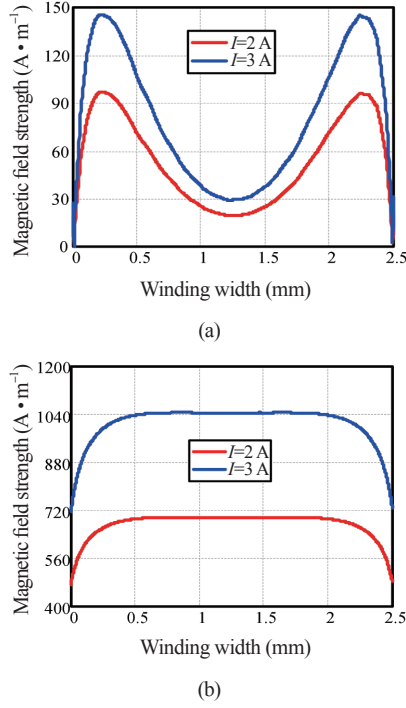


Fig. 10. Magnetic field strength distribution along the primary winding. (a) Upper edge (H_{up}). (b) Lower edge (H_{down}).

on an iterative procedure is proposed, which adaptively adjusts the width of the parallel subconductors in each turn according to the magnetic field distribution, thereby fulfilling the requirements for accurate loss calculation.

B. Iterative Winding Segmentation Algorithm

As shown in Fig.7, under high-frequency conditions, the current density near the winding edges fluctuates significantly, while the distribution in the central region is relatively smooth. Therefore, the winding segmentation strategy should be consistent with this nonuniform distribution characteristic. As illustrated in Fig.10, the magnetic field variation along the upper boundary of the winding is highly consistent with the current density distribution, whereas the magnetic field along the lower boundary cannot effectively represent the current crowding and distortion characteristics. Therefore, the magnetic field at the upper boundary is selected as the segmentation reference, thereby effectively improving the accuracy of the winding loss calculation.

Based on the foregoing analysis, the proposed method first defines the core geometry, number of turns, and operating frequency of the planar transformer. A uniformly segmented PCB winding is then adopted as the initial segmentation, and the current distribution among the parallel subconductors is obtained via the formulations presented in Section II. From these results, the magnetic field intensity at the upper boundary of a specific winding turn is subsequently derived.

The winding is then resegmented using the improved iterative algorithm proposed in this work. The algorithm proceeds as follows: (1) the winding segmentation scheme is determined

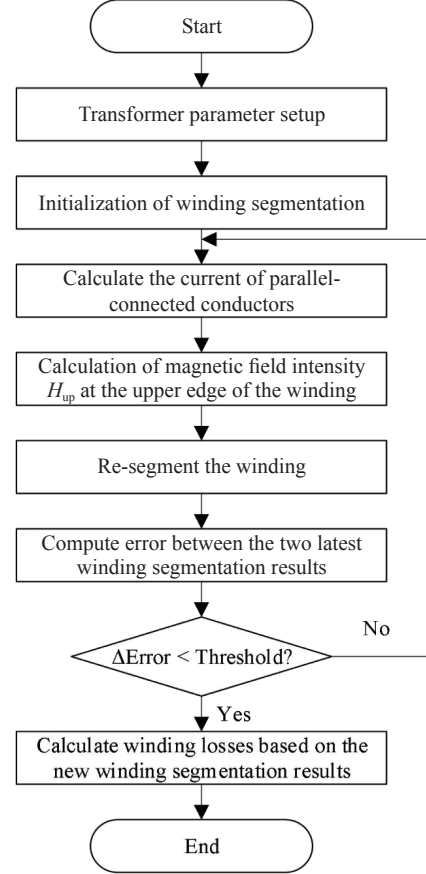


Fig. 11. Basic flowchart of the improved winding segmentation algorithm for PCB windings in planar transformers

based on the boundary magnetic field distribution; (2) the resulting segmentation is validated against a predefined accuracy criterion; (3) the error between two consecutive segmentation results is evaluated to assess iterative convergence. If the error falls below a predefined threshold, the iteration is terminated; otherwise, the latest segmentation result is adopted as the initial condition for the subsequent iteration. The detailed implementation procedure is elaborated upon below, and the overall algorithm flow is illustrated in Fig.11.

1) Extremum Value Identification

First, a Cartesian coordinate system is established at the intersection of the magnetic core pillar and the core window. Then, let d_1 denote the horizontal coordinate of an extremum on the magnetic field intensity distribution curve. Next, the curve is uniformly sampled at n discrete points, and an extremum is identified when the condition (12) is satisfied. To prevent false detection arising from numerical or computational errors, an extremum is deemed valid only if it is consistently greater than or smaller than several neighboring points on both sides.

$$(\dot{H}_{i-1} < \dot{H}_i \& \& \dot{H}_i > \dot{H}_{i+1}) \parallel (\dot{H}_i < \dot{H}_{i-1} \& \& \dot{H}_i < \dot{H}_{i+1}), \quad i = 2, 3, \dots, n-1 \quad (12)$$

2) Slope Calculation

The derivative of the magnetic field intensity curve is com-

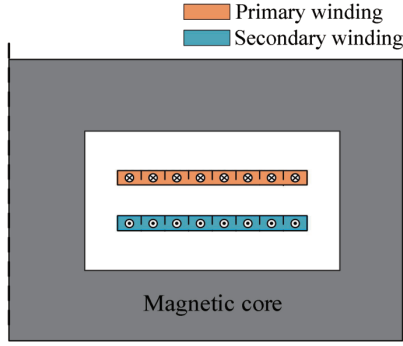


Fig. 12. Two-turn planar transformer.

TABLE I
LENGTHS OF CONDUCTORS IN THE LAST FEW ITERATIONS

Small conductor number	Lengths / mm		
1	0.03	0.02	0.02
2	0.15	0.14	0.13
3	0.10	0.13	0.14
4	0.92	0.91	0.91
5	0.92	0.91	0.91
6	0.10	0.13	0.14
7	0.15	0.14	0.13
8	0.03	0.02	0.02

puted to represent the slope of the curve. Assuming the winding width is w , the slope matrix k satisfies the condition (13).

$$k_i = \frac{\dot{H}_{i+1} - \dot{H}_i}{\frac{w}{n}}, i = 1, 2, \dots, n-1 \quad (13)$$

3) Slope Extremum Value Identification

The extrema of the slope matrix k are then determined, and their horizontal coordinates are denoted as d_2 . Similar to step 1, an extremum is considered valid only if it is simultaneously greater than or smaller than several adjacent slope values.

$$(k_{i-1} < k_i \text{ \& \& } k_i > k_{i+1}) \parallel (k_i < k_{i-1} \text{ \& \& } k_i > k_{i+1}), i = 2, 3, \dots, n-2 \quad (14)$$

4) Segmentation Point Refinement

The sets d_1 and d_2 are merged and sorted in ascending order to form a new set d_3 . The magnetic field intensity differences between adjacent points in d_3 are evaluated, and an additional division point d_4 is inserted between the two points exhibiting the largest field difference to form the set d_5 . Then, the distances between consecutive points in d_5 are recalculated to generate the set d_6 , where the elements of d_6 correspond to the width of each segmented subconductor.

5) Iteration and Convergence

The segmentation error tolerance is defined according to the design requirements. As shown in Fig. 7(b), the current density

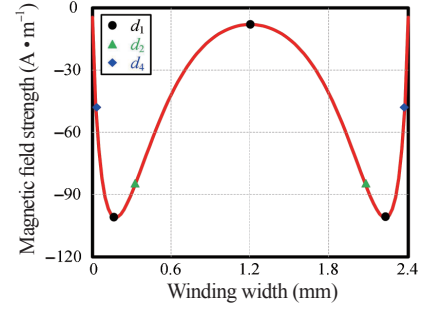


Fig. 13. Schematic diagram of PCB winding resegmentation results.

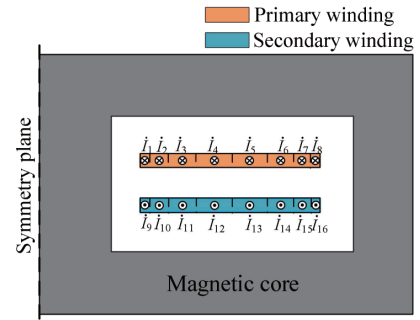


Fig. 14. Results of PCB winding resegmentation.

is primarily concentrated near the winding edges; therefore, accurate segmentation in these regions is particularly critical. By comparing the current distributions and segmentation results obtained from two consecutive iterations, the algorithm is terminated once the difference falls below the predefined tolerance.

To illustrate the algorithm, the two-turn planar transformer shown in Fig. 12 is considered as an example. The primary and secondary windings are initially segmented into eight parallel-connected conductors of equal width, which serve as the initial segmentation. The core window width and height are 3.5 mm and 1.9 mm, respectively, while the winding width and thickness are 2.4 mm and 0.2 mm. The operating frequency is set to 100 kHz.

Based on the proposed algorithm, the winding is iteratively resegmented, and the results obtained after several iterations are summarised in Table I. As shown, the segmentation converges within the last three iterations, with negligible differences that fall within the specified error tolerance. Consequently, the algorithm can be terminated at this point. Both the primary and secondary PCB windings are ultimately segmented into eight parallel-connected conductors with unequal widths, as illustrated in Figs. 13 and 14.

IV. EXPERIMENTAL VERIFICATION

To validate the accuracy of the proposed improved winding segmentation algorithm and its associated winding loss model, both numerical simulations and experimental measurements were carried out. Specifically, the current distribution of the equivalent parallel-connected subconductors and the average magnetic field intensity along the conductor boundaries, as obtained via the proposed algorithm, were compared against the



Fig. 15. Prototype of the planar transformer.

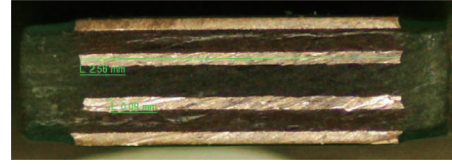


Fig. 17. Winding cross-section of one-quarter planar transformer prototype.

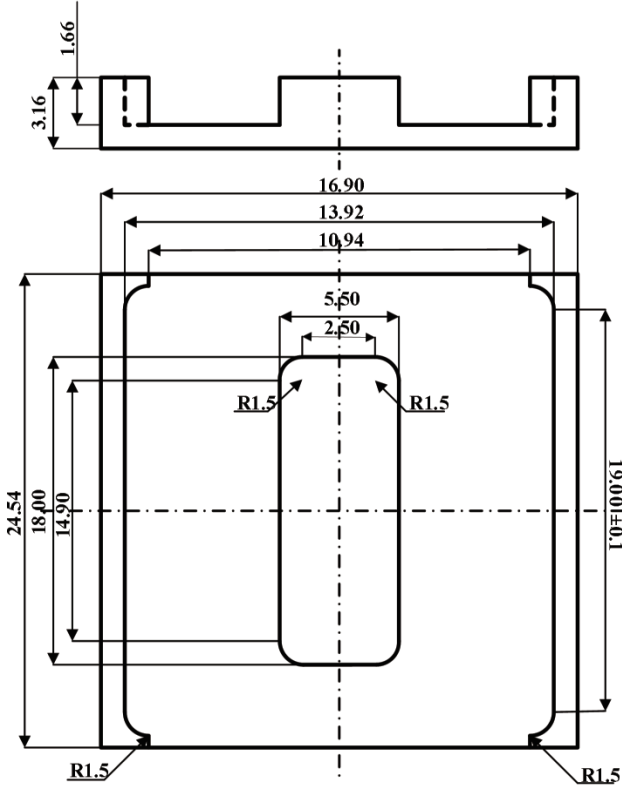


Fig. 16. Dimensions of the magnetic core. (unit: mm)

results from FEM simulations. In addition, the AC resistance of a planar transformer was experimentally measured across a wide frequency range and compared with analytical predictions and simulation results, thereby evaluating the accuracy of the winding loss model over a wide range of frequencies.

A. Verification of the Improved Winding Segmentation Algorithm

An 8-turn planar transformer prototype was fabricated for experimental validation, as shown in Fig. 15. The transformer consists of two identical four-layer PCB boards connected in series. Both the primary and secondary windings comprise four layers connected in series, resulting in a primary-to-secondary turn ratio of 1 : 1. An interleaved winding arrangement was adopted to enhance coupling performance. The magnetic core material is DMR95 from the ECW16.8 series manufactured by the DMEGC. The detailed dimensions of the magnetic core are illustrated in Fig. 16. The core pillar adopts a racetrack-shaped geometry, which effectively reduces the risk of insulation damage associated with sharp corners in rectangular cores.

To obtain accurate geometric parameters for modeling, the PCB winding of the planar transformer was sectioned, and

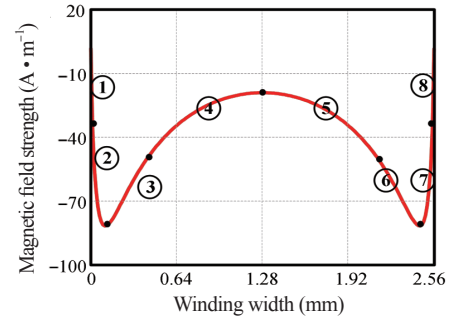


Fig. 18. Results of the improved winding segmentation.

TABLE II
RESULTS OF IMPROVED WINDING SEGMENTATION

Small conductor number	Lengths / mm
1	0.02
2	0.11
3	0.32
4	0.83
5	0.83
6	0.32
7	0.11
8	0.02

one-quarter of the winding cross-section was examined using an industrial microscope, as shown in Fig. 17. The measured conductor width, thickness, core window dimensions, and winding interconnection details were used as inputs to the MATLAB-based segmentation algorithm.

The main parameters were set as follows. The width of each winding layer was 2.56 mm, and the copper thickness was 0.08 mm. Each winding layer was initially segmented into eight equal-width segments, each 0.32 mm in width. The operating frequency was set to 100 kHz, and a sinusoidal current with a peak value of ± 1.414 A was applied to each winding layer. The core window height and width were 3.32 mm and 4.21 mm, respectively. After several iterations, the segmentation results converged, as shown in Fig. 18 and Table II.

To validate the correctness of the segmentation theory, the converged segmentation results were substituted back into the algorithm to recalculate the current distribution of the equivalent parallel-connected conductors and the average magnetic field intensity along their four boundaries. These results were then compared with those obtained from FEM simulations.

A 2-D FEM model of the transformer, shown in Fig. 19, was

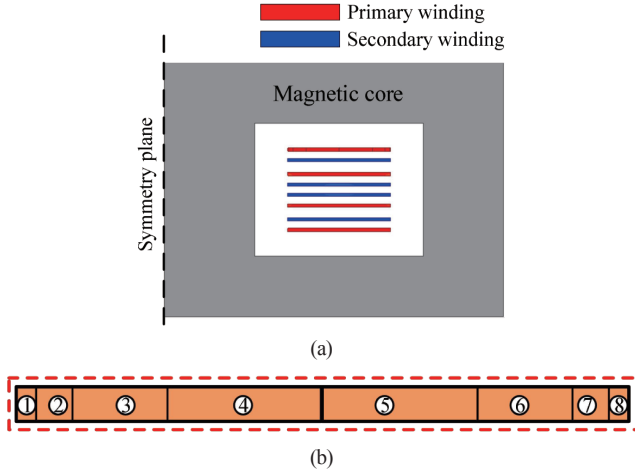


Fig. 19. FEM setup for improved winding-segmentation simulation. (a) 2-D simulation model. (b) Segmentation of half-turn windings.

TABLE III
CURRENT COMPARISON FOR SMALL CONDUCTORS

Small conductor number	Result	
	Calculation / A	Simulation / A
1	0.011	0.011
2	0.063	0.063
3	0.181	0.181
4	0.451	0.452
5	0.451	0.452
6	0.181	0.181
7	0.063	0.063
8	0.011	0.011

constructed. Sinusoidal currents with an amplitude of ± 1.414 A were applied to both the primary and secondary windings, and the simulation frequency was set to 100 kHz. Using the segmentation results obtained from the proposed algorithm, the first winding layer was segmented into eight parallel-connected conductors with unequal widths, while all other parameters were kept consistent with the MATLAB model.

The current distribution among the eight parallel-connected conductors in the first winding layer and the average magnetic field intensity along each conductor boundary were extracted and compared with the FEM results. The comparison, shown in Table III and Fig. 20, indicates close agreement between the analytical and simulation results. Both the current distribution and the boundary magnetic field intensities are accurately predicted, thereby validating the correctness of the proposed winding segmentation algorithm.

B. Verification Experiment of the Loss Model With the Improved Algorithm

Using the segmentation results obtained with the improved algorithm, each winding turn was segmented into eight parallel-connected conductors with unequal widths. For comparison,

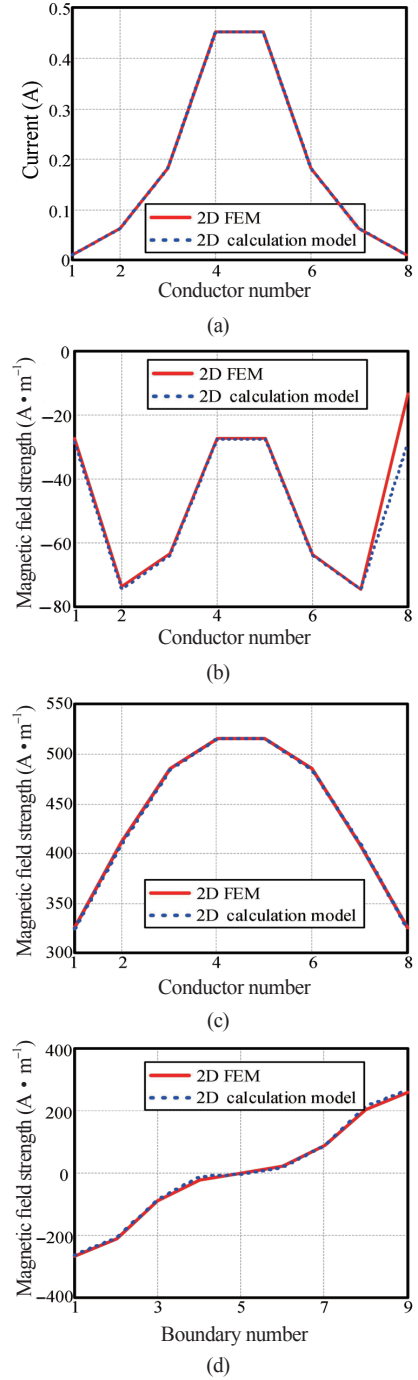


Fig. 20. Comparison results between the 2-D FEM and the proposed 2-D calculation model. (a) Current distribution of small conductors. (b)–(d) Magnitude of average magnetic field strength at the (b) upper, (c) lower, and (d) side boundaries.

two additional segmentation schemes were considered, in which each turn was uniformly segmented into 8 and 10 equal-width conductors, respectively. These segmentation schemes were applied to the winding loss model described in Section II, with all other parameters remaining unchanged.

The AC resistance of the planar transformer in the frequency range from 100 kHz to 1 MHz was calculated using the 2-D winding loss model. The calculated results for the uniform

TABLE IV
IMPEDANCE ANALYSER RAC MEASUREMENT RESULTS

Frequency/kHz	Uniform /mΩ		Improved 8-segment/mΩ
	8-segment	10-segment	
100	44.4	44.7	44.7
200	44.8	45.1	45.1
300	45.3	45.7	45.7
400	45.8	46.4	46.4
500	46.5	47.0	47.1
600	47.2	47.7	48.0
700	48.0	48.6	49.0
800	48.8	49.5	50.0
900	49.8	50.5	51.1
1000	50.8	51.5	52.3

TABLE V
PLANAR TRANSFORMER AC RESISTANCE SIMULATION RESULTS

Frequency/kHz	Simulation result/mΩ	
	2-D	3-D
100	44.7	44.8
200	45.1	46.2
300	45.7	47.5
400	46.3	48.7
500	47.1	49.8
600	47.9	50.9
700	48.9	51.9
800	49.9	52.7
900	51.1	53.6
1000	52.3	54.3

segmentation and the improved segmentation schemes are summarised in Table IV. For comparison, the 2-D and 3-D AC resistance values of the transformer obtained by FEM are listed in Table V. Among them, the 2-D simulation model is shown in Fig. 7(a), and the 3-D simulation model is shown in Fig. 21.

Fig. 22(a) compares the calculated and simulated AC resistance values over the extended frequency range up to 1 MHz, and Fig. 22(b) presents the relative error between the analytical models and the FEM simulation results. In contrast, comparison with FEM simulations shows that the improved segmentation model achieves a calculation error of less than 0.22% up to 1 MHz, as illustrated in Fig. 22(b). In contrast, increasing the number of uniformly segmented conductors to ten reduces the error slightly, but its accuracy still does not match that of the eight-segment improved segmentation scheme. Moreover, the improved eight-segment algorithm requires less computational time than the ten-segment uniform scheme, despite incorporating an additional segmentation step.

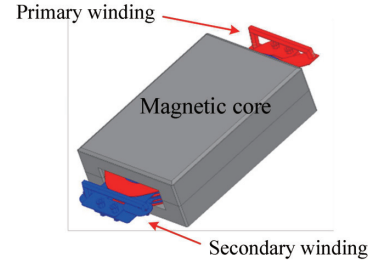
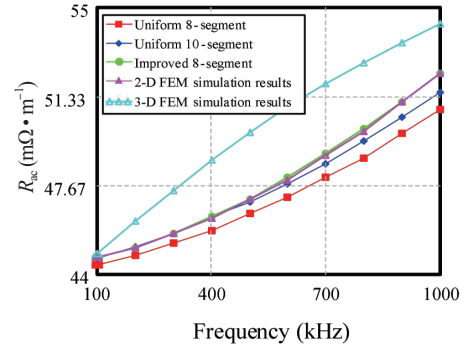
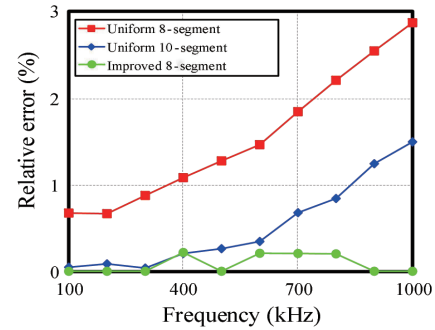


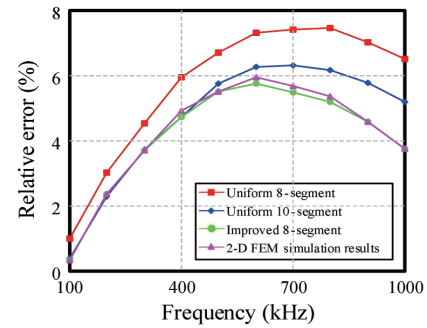
Fig. 21. 3-D simulation model.



(a)



(b)



(c)

Fig. 22. Comparison of AC resistance performance for various segmentation models at different frequency. (a) AC resistance ratio of different models at different frequencies. (b) Relative error of different models vs. the 2-D FEM simulation results model at different frequencies. (c) Relative error of different models vs. the 3-D FEM simulation results model at different frequencies.

Fig. 22(c) shows the relative error between the 2-D model results and the 3-D simulation results. It can be observed from Fig. 22(c) that the relative error between the proposed im-

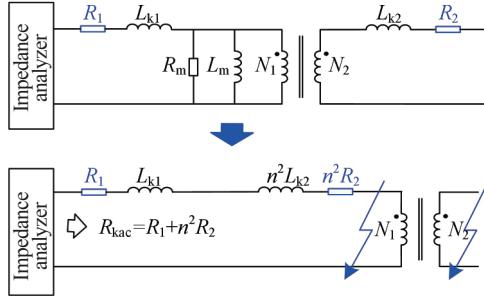


Fig. 23. Principle of the secondary short-circuit measurement method.



Fig. 24. Measurement diagram of (a) entire resistance and (b) lead resistance.

proved winding segmentation loss model and the 3-D finite element simulation results is within 6% up to 1 MHz. Although its accuracy in quantitative loss calculation is lower than that of the 3-D model, the 2-D model significantly improves computational efficiency at the cost of reducing one-dimensional spatial information, making it more suitable for the preliminary topology selection and loss mechanism analysis of planar transformers.

In the FEM simulations, a sinusoidal current with a peak value of 1.414 A was applied to the primary winding, and the corresponding winding loss was used to determine the equivalent AC resistance of the transformer. To further validate the analytical model, experimental measurements were conducted using a Wayne Kerr WK6500B precision impedance analyser, which provides a maximum operating frequency of 120 MHz and a basic measurement accuracy of 0.05%.

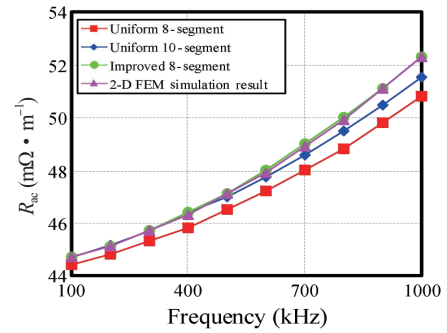
During the measurements, the secondary winding of the transformer was short-circuited, and the primary winding was connected to the impedance analyser, as illustrated in Fig. 23. Under this condition, the magnetic flux in the core is negligible, and core losses can be ignored. Consequently, the measured impedance primarily corresponds to the winding's AC resistance. Since the connecting leads exhibit inherent resistance during the measurement process, the resistance for small-signal measurement was obtained by subtracting the lead resistance from the total measured resistance. The physical measurement setup is shown in Fig. 24, and the measured AC resistance values are summarised in Table VI.

Fig. 25(a) compares AC resistance of the planar transformer obtained from the uniform segmentation model, the improved segmentation model, the FEM simulations, and the experimental measurements over the frequency range from 100 kHz to 1 MHz. The corresponding relative error with respect to the experimental measurements is shown in Fig. 25(b).

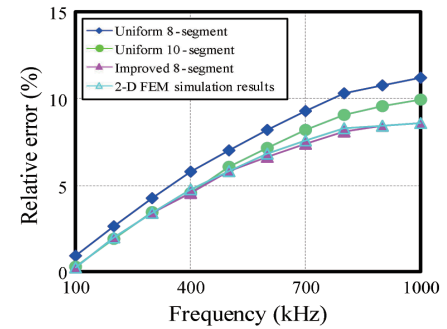
The results demonstrate that the winding loss model based

TABLE VI
IMPEDANCE ANALYSER RAC MEASUREMENT RESULTS

Frequency/kHz	Resistance/mΩ		Small-signal measured/mΩ
	Entire	Lead	
100	50.4	5.6	44.8
200	52.0	6.0	46.0
300	53.8	6.5	47.3
400	55.8	6.9	48.6
500	57.4	7.4	50.0
600	59.4	8.0	51.4
700	61.5	8.6	52.9
800	63.6	9.2	54.4
900	65.5	9.7	55.8
1000	67.4	10.2	57.2



(a)



(b)

Fig. 25. AC-resistance comparison among different calculation models. (a) AC resistance ratio of different models at different frequencies. (b) Relative error of different models vs. the small-signal model at different frequencies.

on the improved segmentation strategy exhibits significantly better agreement with both experimental measurements and FEM simulations than the uniform segmentation model. Although the relative error with respect to experimental measurements reaches approximately 8% at 1 MHz, this discrepancy can be attributed to the limitations of the 2-D model.

During the 2-D simulation, due to dimensional limitations and computational simplifications, the winding is idealized as a symmetric structure fully enclosed by the magnetic core, resulting in more regular magnetic boundary conditions and

a relatively uniform magnetic field distribution. However, in practical planar transformer structures, the winding is typically fabricated in a racetrack configuration, in which the magnetic core window can only partially surround the winding conductors, leaving some edge regions directly exposed to the air domain. In addition, the actual winding contains 3-D structures such as interlayer vias and terminal leads. These factors alter the local magnetic field distribution, further affecting winding losses. Since these 3-D structures and non-ideal factors cannot be accurately reproduced in the 2-D model, a discrepancy inevitably arises between the simulation and experimental results.

Overall, the experimental and numerical results confirm that the proposed improved winding segmentation algorithm offers superior accuracy and computational efficiency for analysing high-frequency winding losses in planar transformers.

V. CONCLUSION

This paper has presented an improved winding segmentation algorithm for accurate winding-loss calculation in PCB planar transformers.

At high-frequency operating conditions, nonuniform current distribution caused by skin and proximity effects limits the accuracy of conventional uniform winding segmentation methods. To overcome this issue, an iterative winding segmentation algorithm was proposed, in which the segmentation is adaptively determined based on the distribution of magnetic field strength along the winding boundaries.

An 8-turn planar transformer was used as the case study in this work. By comparing the current distribution of the equivalent parallel conductors and the average magnetic field strength along the winding boundaries, as obtained from 2-D FEM simulation and the improved winding segmentation algorithm, the accuracy of the proposed algorithm was validated. Furthermore, by comparing the winding-loss calculation results and the corresponding relative errors obtained from the conventional uniform winding segmentation loss model, the improved winding segmentation loss model, FEM simulation and small-signal measurement, the results demonstrate that the improved model not only achieves higher calculation accuracy with the same number of segments but also provides superior accuracy with a reduced number of segments, thereby confirming the effectiveness of the proposed method.

REFERENCES

- [1] R. Yu, T. Chen, P. Liu, and A. Q. Huang, "A 3-D winding structure for planar transformers and its applications to LLC resonant converters," in *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 9, no. 5, pp. 6232–6247, Oct. 2021.
- [2] H. Liu, K. Sun, Z. Yi, G. Cao, and S. Lu, "Modeling and evaluation of winding loss for high-voltage planar transformer," in *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 13, no. 1, pp. 904–920, Feb. 2025.
- [3] L. Pniak, L. Quéval, B. Revol, J.-S. N. Teu, C. Gautier, and O. Béthoux, "AC resistance and leakage inductance estimation for planar transformers with parallel-connected windings," in *IEEE Transactions on Power Electronics*, vol. 38, no. 1, pp. 728–738, Jan. 2023.
- [4] P. L. Dowell, "Effects of eddy currents in transformer windings," in *Proceedings of the Institution of Electrical Engineers*, vol. 113, no. 8, pp. 1387–1394, Aug. 1966.
- [5] J. A. Ferreira, "Improved analytical modeling of conductive losses in magnetic components," in *IEEE Transactions on Power Electronics*, vol. 9, no. 1, pp. 127–131, Jan. 1994.
- [6] Z. Ouyang, O. C. Thomsen, and M. A. E. Andersen, "Optimal design and tradeoffs analysis for planar transformer in high power DC-DC converters," in *Proceedings of the 2010 International Power Electronics Conference*, Sapporo, Japan, 2010, pp. 3166–3173.
- [7] J. A. Ferreira, "Appropriate modelling of conductive losses in the design of magnetic components," in *Proceedings of 21st Annual IEEE Conference on Power Electronics Specialists*, San Antonio, USA, 1990, pp. 780–785.
- [8] B. Y. Liu and W. Chen, "An improved high-frequency loss model for round-wire windings of magnetic components," in *Proceedings of the CSEE*, vol. 39, no. 9, pp. 2795–2803, 2019.
- [9] X. Nan and C. R. Sullivan, "An improved calculation of proximity-effect loss in high-frequency windings of round conductors," in *Proceedings of IEEE 34th Annual Conference on Power Electronics Specialist*, Acapulco, Mexico, 2003, pp. 853–860, vol. 2.
- [10] F. Robert, J. Sprooten, P. Mathys, J.-P. Schauwers, and B. Velaerts, "Eddy current losses in SMPS transformers round wire windings: A semi-analytical closed-form formula," in *Proceedings of 2005 European Conference on Power Electronics and Applications*, Dresden, Germany, 2005, p. 9.
- [11] N. Dai and F. C. Lee, "Edge effect analysis in a high-frequency transformer," in *Proceedings of 1994 Power Electronics Specialist Conference*, Taipei, Taiwan, China, 1994, pp. 850–855, vol. 2.
- [12] F. Robert, P. Mathys and J.-P. Schauwers, "A closed-form formula for 2-D ohmic losses calculation in SMPS transformer foils," in *IEEE Transactions on Power Electronics*, vol. 16, no. 3, pp. 437–444, May 2001.
- [13] F. Robert, P. Mathys, B. Velaerts, and J.-P. Schauwers, "Two-dimensional analysis of the edge effect field and losses in high-frequency transformer foils," in *IEEE Transactions on Magnetics*, vol. 41, no. 8, pp. 2377–2383, Aug. 2005.
- [14] X. Wang, L. Wang, L. Mao, L. Yi, and S. Yang, "Calculation method of winding loss in high frequency planar transformer," in *Proceedings of 2016 International Conference on Electrical Systems for Aircraft, Railway, Ship Propulsion and Road Vehicles & International Transportation Electrification Conference (ESARS-ITEC)*, Toulouse, France, 2016, pp. 1–5.
- [15] D. Ahmed, L. Wang, M. Wu, L. Mao, and X. Wang, "Two-dimensional winding loss analytical model for high-frequency multilayer air-core planar inductor," in *IEEE Transactions on Industrial Electronics*, vol. 69, no. 7, pp. 6794–6804, Jul. 2022.
- [16] T. Tian, W. Chen, and S. Lin, "Two-dimensional calculation model of the edge effect and losses in high-frequency transformer foils," in *Proceedings of 2021 IEEE 1st International Power Electronics and Application Symposium (PEAS)*, Shanghai, China, 2021, pp. 1–6.
- [17] X. K. Mao and W. Chen, "Modeling and analysis of AC losses in parallel PCB windings of high-frequency planar transformers for switching power supplies," in *Proceedings of the CSEE*, vol. 26, no. 2, pp. 167–173, 2006.



Jinjiang Li was born in Quanzhou, China, in 2001. He received the B.S. degree in electrical engineering and automation from Fuzhou University, China, in 2023. He is currently pursuing a master's degree at Fuzhou University, Fuzhou. His research interest includes high-frequency magnetic technology.



Binzhi Xie was born in Quanzhou, China, in 1998. He received the B.S. degree in electrical engineering and automation from Fuzhou University, China, in 2021. He received the master's degree in electrical engineering and automation from Fuzhou University, Fuzhou, China, in 2024. His research interest includes high-frequency magnetic technology.



Wei Chen received the M.S. and Ph.D. degrees in electrical engineering from Fuzhou University, Fuzhou, China, in 1987 and 1990, respectively. He was a Senior Visiting Professor with the Centre for Power Electronics Systems, Virginia Tech, Blacksburg, VA, USA, from 1996 to 1998. He was a Research and Development Manager at Delta Electronics Company Ltd. and the Delta Power Electronics Centre in Shanghai, China, from 1999 to 2008. He has authored or coauthored more than 80 technical articles. He held more than 40 approved patents from China and the USA. His research interests include power conversion, high-frequency magnetic technology, EMI debug and solution, wireless power transfer, electromagnetic field analysis, and applications. Dr. Chen is a fellow of the China Power Supply Society.



Qingbin Chen was born in Quanzhou, China. He received the B.E. and Ph.D. degrees in electrical engineering from Fuzhou University, Fuzhou, China, in 2007 and 2012, respectively. He worked as a visiting scholar at the University of Florida in the USA from 2017 to 2018. He is now a professor and Ph.D. advisor at Fuzhou University. His research encompasses high-frequency magnetic technology, EMC diagnosis and suppression technology, and wireless power transfer. Dr. Chen served as the Vice Chairperson and Deputy Secretary General of the Magnetic Component Specialty Committee of CPSS (China Power Supply Society).