

A Wideband Modeling Method for Magnetic Components Considering Strong/Weak Near-Field Coupling Characteristics

Xuanran HUANG and Kaining FU

Abstract—Magnetic components in switching power supplies generate severe leakage electromagnetic fields, which not only affect the filtering performance of EMI filters but also may disrupt the normal operation of circuits. This paper first analyzes the mechanism of near-field leakage generated by magnetic components. It investigates the effect of frequency-dependent permeability on the self-parameters of these components and further explores how the self-parameters influence mutual parameters. By integrating both effects on near-field coupling, a wideband near-field coupling model for magnetic components is established, which accounts for the transmission characteristics of both magnetic and electric field coupling. Insertion loss is measured using a network analyzer and converted into Z-parameters to extract the values of equivalent inductance, equivalent capacitance, and equivalent resistance in the self-parameter module and mutual-parameter module. The near-field transmission characteristics of magnetic components under both strong and weak coupling scenarios are tested to verify the accuracy of the proposed wideband near-field coupling model, which exhibits good conformity in the frequency range from 100 kHz to 60 MHz. The results indicate that magnetic field coupling and electric field coupling dominate the near-field transmission characteristics of magnetic components in different frequency bands.

Index Terms—Insertion loss, magnetic component, near-field coupling, wideband model.

I. INTRODUCTION

WITH the significant improvement of switching frequency and power density, switching power supplies are facing severe electromagnetic interference (EMI) problems. Designing EMI filters is an effective means to suppress electromagnetic interference [1]. Common-mode (CM) inductors are key magnetic components in EMI filters, and the leakage electromagnetic fields generated by them will couple with other filter

components to cause near-field coupling effects, affecting the filtering effect of EMI filters.

Research on the near-field coupling of magnetic components mainly focuses on three aspects: near-field coupling mechanism analysis, modeling, and near-field coupling suppression. In terms of near-field coupling mechanism, early research mainly focused on the magnetic field coupling mechanism. Notably, [2]–[3] carried out in-depth investigations into the near-field coupling mechanism of EMI filters. Researches indicated that while the leakage inductance of CM inductors can serve as differential-mode (DM) inductors, but the leakage flux causes near-field coupling. This phenomenon degrades the high-frequency impedance characteristics of the filter components and further compromises the overall filtering performance of the EMI filter. Furthermore, it is found that the permeability of magnetic components decreases rapidly with the increase of frequency, which can result in a large leakage flux. In addition, high-frequency potential variations exist on the windings of magnetic components. As a result, electric field coupling caused by winding potential variations may further degrade the filtering performance of EMI filters [4]–[6]. Therefore, it is necessary to comprehensively consider the influence of electric field and magnetic field coupling on near-field coupling. In terms of near-field coupling modeling, the existing modeling methods for magnetic components can be mainly divided into lumped circuit modeling and finite element simulation modeling. Most of the existing lumped circuit modeling only considers the influence of magnetic field coupling, that is, a mutual inductance model is used to characterize the coupling effect between components. To evaluate the accuracy of the model, [7] proposes to use the insertion loss (IL). It is found that the effective frequency range of the magnetic field coupling model is usually only 150 kHz–10 MHz. In fact, the main reason is that the existing models do not fully consider the comprehensive effect of electric field and magnetic field coupling on near-field coupling [8]. For the acquisition of model parameters, impedance measurement, scattering parameter measurement, finite element simulation and other methods can be used [9]. [10]–[12] propose to use partial element equivalent circuit, finite element simulation and other methods to characterize the EMI transmission characteristics of each filtering component. Although these methods can accurately characterize the near-field coupling behavior of magnetic components, they have problems such

Manuscript received December 3, 2025; revised February 10, 2026; accepted March 13, 2026. Date of publication September 30, 2026; date of current version July 13, 2026. This work was supported in part by the National Natural Science Foundation of Fujian Province under the grant 2025J011294 and Natural Science Foundation of Xiamen, China under the grant 3502Z202372043. (Corresponding author: Kaining Fu.)

Both authors are with School of Electrical Engineering and Automation, Xiamen University of Technology, Xiamen 361024, China (e-mail: 2664818025@qq.com; fkn@xmut.edu.cn).

Digital Object Identifier 10.24295/CPSSPEA.2026.00014

as complex modeling and long simulation time [13]. In terms of near-field coupling suppression, [14]–[16] focus on the cancellation of mutual parasitic parameters of EMI filters, and eliminate the influence of leakage magnetic fields on filter components by designing the placement of components and PCB wiring, so as to improve the high-frequency characteristics of filter components. [17]–[19] propose methods to optimize the leakage magnetic field properties of CM inductors via magnetic core and winding designs. These approaches can reduce the magnitude of leakage flux from the magnetic emission source, thereby mitigating the impact of near-field coupling on other components.

This paper studies the near-field coupling modeling method of magnetic components under different coupling degrees, which considers the frequency-dependent characteristics of the permeability of magnetic cores. Using the lumped circuit modeling method, it comprehensively considers the coupling behavior of leakage flux and electric field caused by the influence of self-parameters of magnetic components on mutual parameters. The proposed model is verified by scattering parameter measurement, which can accurately characterize the near-field coupling characteristics in the range of 100 kHz–60 MHz.

The main contribution of this paper is to propose a modeling method for magnetic components that can fully characterize near-field electromagnetic coupling within the frequency range of 100 kHz to 60 MHz. The key innovation lies in the integration of frequency-dependent inductance into the proposed near-field coupling model, which enables accurate representation of the magnetic core's impact on near-field coupling behavior.

The remainder of this paper is structured as follows. Section II presents the extraction methods for electric field coupling and magnetic field coupling. Section III analyzes the influence of self-parameters on mutual parameters, followed by the proposal of a wideband frequency modeling method for near-field coupling. The effectiveness of the proposed modeling method is validated through scattering parameter (*S*-parameter) measurements. Finally, Section V summarizes the conclusions of this study.

II. NEAR-FIELD COUPLING MODEL AND TEST EVALUATION METHOD OF MAGNETIC COMPONENTS

As shown in Fig. 1, the main flux of the magnetic component forms a closed loop through the magnetic core, while the leakage flux forms a closed loop through the air. For one aspect, leakage flux couples with adjacent components to create near-field coupling. For another aspect, since magnetic component winding terminals are usually connected to switching devices, high-frequency potential fluctuations in the windings generate leakage electric fields that couple to surrounding filter components. Furthermore, since ferrite core permeability is frequency-dependent (nonlinear), the ratio of main flux to leakage flux in ferrite-cored magnetic components also varies nonlinearly with frequency, leading to nonlinear winding potential distribution. Therefore, electric and magnetic

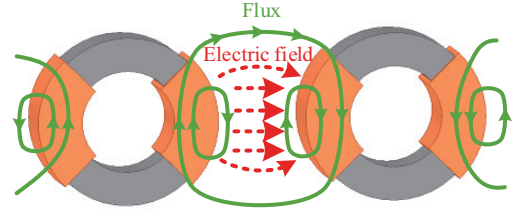


Fig. 1. Mechanism of magnetic field coupling and electric field coupling between magnetic components.

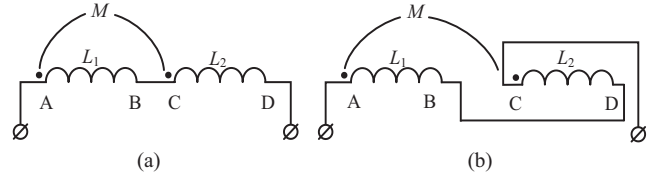


Fig. 2. Mutual inductance measurement method. (a) Series aiding. (b) Series opposing.

field coupling dominate different frequency ranges in near-field coupling. Accordingly, the established near-field coupling model must comprehensively account for both effects on near-field transmission across various frequency ranges.

A. Magnetic Field Coupling Extraction Method

The degree of magnetic field coupling between two magnetic components is characterized by the mutual inductance coefficient k , which is defined as:

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (1)$$

where M is the mutual inductance between them, and L_1 and L_2 are the self-inductances of the two inductors respectively.

To determine the value of the coefficient k , it is necessary to measure the mutual inductance M and the self-inductances L_1 and L_2 . The self-inductance L can be measured by an LCR meter. For measuring the mutual inductance M , two methods can be adopted which depends on the value of the mutual coefficient k :

1. When $k > 0.1$ (indicating strong magnetic coupling between the two inductors), M can be derived by measuring the total inductance of the inductors in both series-aiding and series-opposing configurations using an LCR meter.

2. When $k < 0.1$ (indicating weak magnetic coupling between the two inductors), M can be measured by using the insertion loss method.

1) Series-Aiding and Series-Opposing Method

Fig. 2 shows the measurement of mutual inductance conducted by series-aiding and series-opposing method. Series-aiding configuration connects dots of opposite polarity, and series-opposing configuration connects dots of the same polarity. The two coupled inductors are configured in both modes, with their inductances measured via an LCR meter.

According to the measurement shown in Fig. 2(a) and (b),

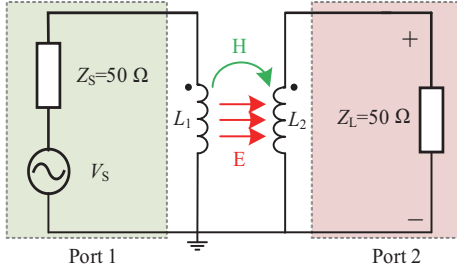


Fig. 3. Schematic diagram of mutual inductance measurement by insertion loss method.

the corresponding inductances L_P and L_N are respectively:

$$\begin{cases} L_P = L_1 + L_2 + 2M \\ L_N = L_1 + L_2 - 2M \end{cases} \quad (2)$$

From (2), the mutual inductance M between the two magnetic components is:

$$M = \frac{L_P - L_N}{4} \quad (3)$$

This method is only applicable for a large mutual inductance coefficient k , where the mutual inductance M and self-inductances L_1 and L_2 are of the same order of magnitude.

2) Insertion Loss Method

For small mutual inductance between two inductors, direct measurement via the series-aiding/series-opposing method introduces significant measurement errors. To accurately evaluate mutual inductance under weak magnetic coupling conditions, the insertion loss method is applicable. It is suitable for scenarios where two magnetic components are relatively far apart and the coupling mode is complex (considering the combined transmission effects of magnetic and electric field coupling), with its measurement principle illustrated in Fig. 3. Port 1 and Port 2 of the network analyzer connect to the two test inductors: Port1 applies a high-frequency voltage to L_1 , and the resulting leakage magnetic/electric fields couple spatially to L_2 for detection by Port 2's RF input. This method accurately characterizes such inter-inductor spatial leakage electromagnetic coupling.

According to the definition of insertion loss, it can be expressed as:

$$IL = 10 \log \frac{P_1}{P_2} = 20 \log \left| \frac{U_1}{U_2} \right| \quad (4)$$

where U_1 represents the voltage drop across load resistor Z_L when Port 1 and Port 2 of the network analyzer are directly connected, while U_2 represents the corresponding voltage drop across Z_L after inserting the two inductors under test.

Based on the above, U_1 can be expressed as:

$$U_1 = \frac{1}{2} V_s \quad (5)$$

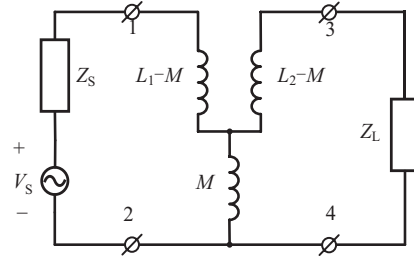


Fig. 4. Circuit decoupling equivalent circuit.

For only magnetic coupling between the two test inductors, circuit decoupling is needed to compute U_2 , with the equivalent decoupled circuit shown in Fig. 4.

At this time, the voltage drop U_2 across the load resistor Z_L can be expressed as:

$$U_2 = \left| \frac{50V_s \frac{j\omega M}{50 + j\omega(L_2 - M) + 2j\omega M}}{50 + j\omega(L_1 - M) + \frac{1}{\frac{1}{j\omega M} + \frac{1}{50 + j\omega(L_2 - M)}}} \right| \quad (6)$$

When the coupling coefficient $k \ll 0.1$, $M \ll \sqrt{L_1 L_2}$. Therefore, the expression of voltage U_2 in (6) can be further simplified as:

$$U_2 = \left| \frac{50 j\omega M V_s}{(50 + j\omega L_1)(50 + j\omega L_2)} \right| \quad (7)$$

By combining (4)–(7), the insertion loss for weak coupling is obtained as:

$$IL(\text{dB}) = 20 \log \left| \frac{100 j\omega M}{(50 + j\omega L_1)(50 + j\omega L_2)} \right| \quad (8)$$

Then, the mutual inductance is obtained as:

$$|M| = \left| \frac{(50 + j\omega L_1)(50 + j\omega L_2)}{100 j\omega} \right| \cdot 10^{\frac{IL(\text{dB})}{20}} \quad (9)$$

B. Electric Field Coupling Extraction Method

Parasitic capacitance between two magnetic components arises from high-frequency potential variations across each winding turn and their opposing area; high-frequency displacement current conducts through this capacitance, leading to electric field coupling between the components. The LCR meter test and insertion loss method are common for extracting parasitic capacitance between two magnetic components.

1. LCR meter test method: The two magnetic components are regarded as the metal plates of a parallel-plate capacitor, through which displacement current flows. The parasitic capacitance obtained via this method is the structural capacitance, determined

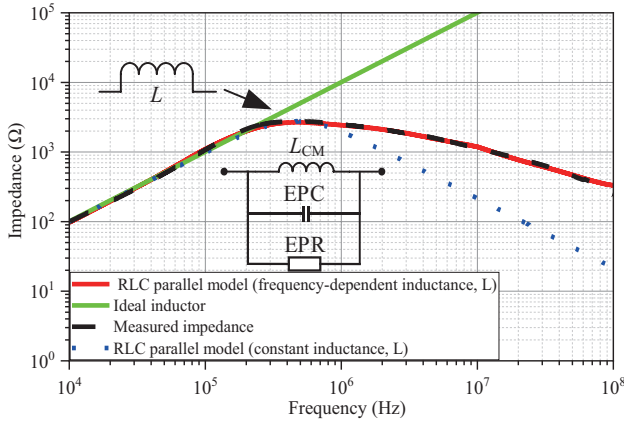


Fig. 5. Common-mode impedance characteristics.

exclusively by their physical parameters (e.g., opposing area and distance).

2. Insertion loss method: The measurement configuration in Fig. 3 is employed, with the key difference that the inter-inductor coupling is modeled as capacitive coupling. For the parasitic capacitance equivalent to lumped parameter C_Q , the voltage drop U_2 across load resistor Z_L is given by:

$$U_2 = \left| \frac{50 j\omega C_Q V_s}{1 + 100 j\omega C_Q} \right| \quad (10)$$

Combining with (5) and (10), the insertion loss can be expressed as:

$$\text{IL(dB)} = 20 \log \left| \frac{1 + 100 j\omega C_Q}{100 j\omega C_Q} \right| \quad (11)$$

Based on the above, the coupling capacitance between the two inductors is given by:

$$C_Q = \left| \frac{1}{100 j\omega (10^{\frac{\text{IL(dB)}}{20}} - 1)} \right| \quad (12)$$

III. WIDEBAND NEAR-FIELD COUPLING MODELING METHOD

A. Self-Parameter Model of CM Inductors

As shown in Fig. 5, the CM impedance of the CM inductor is affected by the equivalent parallel capacitance (EPC) and exhibits linear characteristics only within the 10 kHz–10 MHz range. Within this frequency band, its self-impedance can be characterized by a pure inductor L_{CM} . For higher frequencies, LC resonance occurs between the CM inductor component L_{CM} and the EPC. Additionally, the CM inductor's magnetic core typically uses high-permeability Mn-Zn ferrite or nanocrystalline materials, whose permeability exhibits fre-



Fig. 6. Series equivalent circuit.

quency dependence. The self-parameter modules adopt frequency-dependent inductors $L(f)$ derived from the complex permeability curve, which dynamically adapt to the magnetic core's permeability variations.

The permeability of Mn-Zn ferrite material can be expressed as complex permeability, as shown in (13).

$$\mu = \mu' - j\mu'' \quad (13)$$

where μ' denotes the real part of the complex permeability, which characterizes the magnetic properties of the magnetic core, and μ'' denotes the imaginary part, which accounts for the core losses.

To extract the permeability of the magnetic material, a single-turn inductor is employed to measure the corresponding self-impedance. Fig. 6 depicts the corresponding series equivalent circuit for this measurement.

According to Fig. 6, (14) defines the series equivalent impedance of the inductor. The complex permeability defined in (13) is substituted in (14).

$$Z = R_S + j\omega L_S = j\omega \frac{\mu N_2 A_e}{l_e} \quad (14)$$

where Z denotes the impedance of the series equivalent circuit, and the magnetic core's geometric parameters (effective magnetic path length l_e , effective cross-sectional area A_e) along with the inductor's structural parameter (number of turns N) are also defined.

From the relationship between the complex permeability and the impedance of the series equivalent circuit model, the real and imaginary parts of the complex permeability, i.e., μ' and μ'' , can be derived:

$$\mu' = \frac{l_e L_S}{\mu_0 A_e N^2} \quad (15)$$

$$\mu'' = \frac{l_e R_S}{\mu_0 A_e N^2 \omega} \quad (16)$$

where $\mu_0 = 4\pi \times 10^{-7}$ H/m (vacuum permeability), and $\omega = 2\pi f$ (angular frequency).

Fig. 7 presents the frequency dependence of the complex permeability of TDK's PC95 ferrite. As observed, the real part μ' of the complex permeability decreases sharply above 1 MHz with increasing frequency. This induces frequency-dependent proportions of the main flux and leakage flux in the magnetic core, significantly impacting the leakage magnetic field around the CM inductor. The blue dashed line in Fig. 5 represents the RLC parallel model ($R=2.7542$ kΩ, $C=0.1995$ μF, $L=1.543$ mH), which only characterizes the self-impedance within 10 kHz–1 MHz. As frequency increases, significant errors arise between

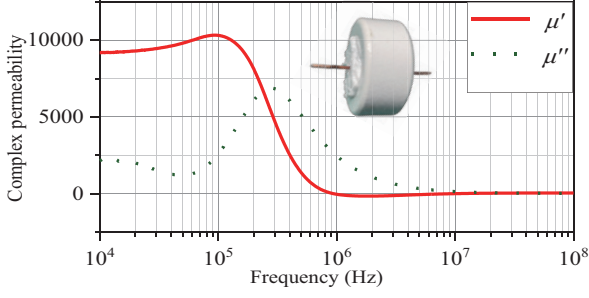


Fig. 7. PC95 core permeability.

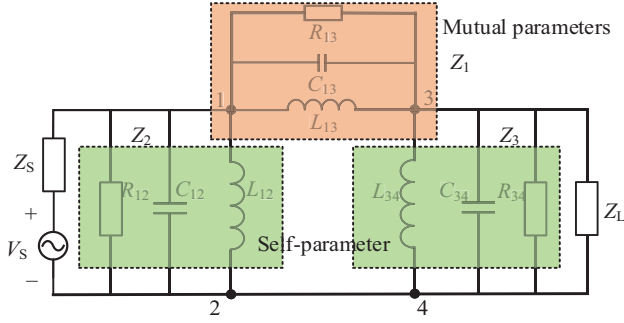


Fig. 8. Near-field coupling wideband model.

the model and actual performance, attributed to the use of a fixed inductance that neglects the frequency dependence of the magnetic core permeability. To improve model accuracy, the fixed inductor in the RLC parallel model is replaced with a frequency-dependent inductor $L(f)$ accounting for permeability variations (red curve in Fig. 5). As observed, the modified model achieves full consistency with measurements across 10 kHz–100 MHz.

Based on the above analysis, the self-impedance characteristics of magnetic components indirectly influence their near-field coupling behavior. Thus, the effect of self-impedance must be fully considered when establishing the near-field coupling model for such components.

B. Proposed Wideband Near-Field Coupling Model

To evaluate the near-field coupling model of magnetic components, the insertion loss measurement method in Fig. 3 is adopted. Accordingly, Fig. 8 presents the proposed wideband near-field coupling model for such components, consisting of self-parameters and mutual parameters.

For the self-parameter modules, the self-parameter model of inductor 1# is composed of the parallel connection of inductance L_{12} , equivalent parallel capacitance C_{12} and equivalent resistance R_{12} for core loss. Similarly, the self-parameters of inductor 2# are derived by analogy. For the mutual parameter module, L_{13} characterizes the magnetic field coupling between the two inductors, C_{13} characterizes the electric field coupling between the two inductors, and R_{13} characterizes the loss caused by the electric and magnetic coupling between the two inductors.

Notably, since both the self-parameter and mutual parameter modules employ the RLC parallel model, both sets of parameters

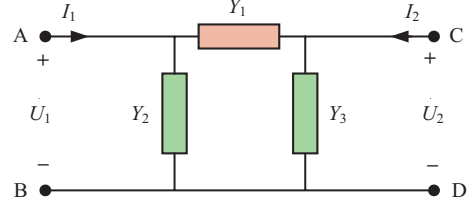


Fig. 9. Π-type circuit schematic diagram.

reflect the electric-magnetic field interaction.

To extract each parameter in the model, the mutual parameter module and the two self-parameter modules are characterized in the form of Y -parameter, which are defined as Y_1 , Y_2 , and Y_3 respectively, as shown in the Π-type circuit in Fig. 9.

Combined with the circuit connection of each component in Fig. 8, the parameters Y_1 , Y_2 , and Y_3 can be expressed as:

$$\begin{cases} Y_1 = \frac{1}{j\omega L_{13}} + \frac{1}{R_{13}} + j\omega C_{13} \\ Y_2 = \frac{1}{j\omega L_{12}} + \frac{1}{R_{12}} + j\omega C_{12} \\ Y_3 = \frac{1}{j\omega L_{34}} + \frac{1}{R_{34}} + j\omega C_{34} \end{cases} \quad (17)$$

The corresponding Y -parameter relationships in the Π-type circuit of Fig. 9 are expressed as:

$$\begin{cases} Y_1 = -Y_{12} \\ Y_2 = Y_{11} + Y_{12} \\ Y_3 = Y_{22} + Y_{12} \end{cases} \quad (18)$$

To extract the Y -parameters, the S -parameters are first measured via a network analyzer and then converted to Y -parameters. The conversion relationship between Y -parameters and S -parameters is expressed as:

$$\begin{cases} Y_{11} = \frac{1}{Z_0} \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \\ Y_{12} = \frac{1}{Z_0} \frac{-2S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \\ Y_{21} = \frac{1}{Z_0} \frac{-2S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \\ Y_{22} = \frac{1}{Z_0} \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \end{cases} \quad (19)$$

As Y -parameters and Z -parameters are reciprocals, Z -parameters can be derived by inverting the Y -parameters.

IV. EXPERIMENTAL VERIFICATION

To verify the accuracy of the proposed wideband nearfield coupling model for magnetic components, the near-field coupling transmission characteristics of CM inductors are

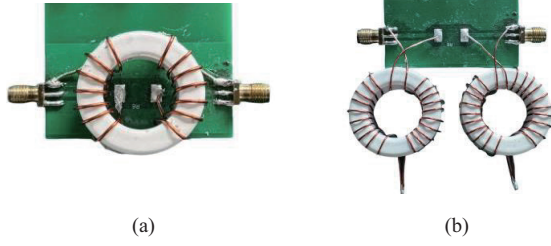


Fig. 10. Near-field coupling test site layout. (a) Strong coupling. (b) Weak coupling.

TABLE I
CM INDUCTOR SPECIFICATION PARAMETERS

Size	TX32/19/13	
Manufacturer	Ferroxcube	
Material	Mn-Zn ferrite	
Number of turns	Strong coupling	7
	Weak coupling	11
Opening angle	Strong coupling	60°
	Weak coupling	30°

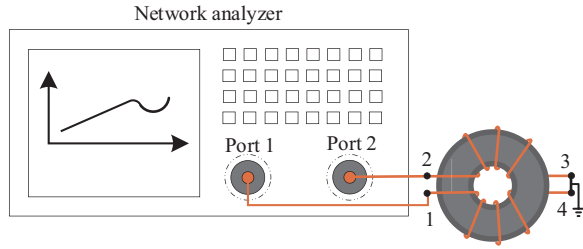


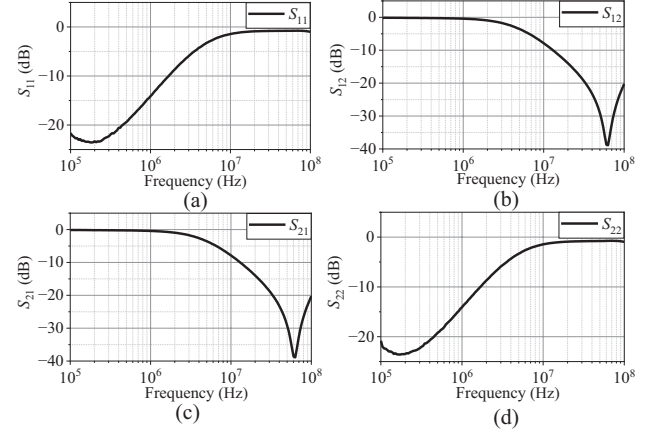
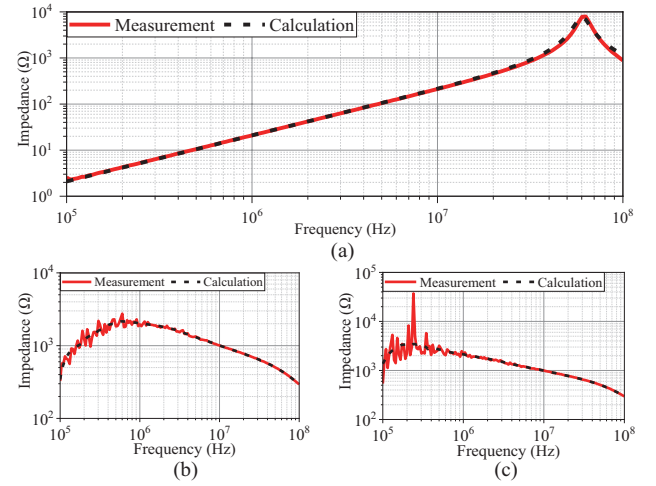
Fig. 11. Network analyzer measurement of scattering parameters schematic diagram.

investigated under strong and weak coupling conditions. The wound CM inductors and corresponding test setup are illustrated in Fig. 10, with their key specifications provided in Table I.

A. Near-Field Transmission Characteristics Under Strong Coupling

To investigate the near-field transmission characteristics of two inductors under strong coupling, the CM inductor in Fig. 9(a) is taken as an example. Owing to the shared magnetic core of its dual-side windings, strong magnetic coupling exists between the two inductors. To comprehensively evaluate the near-field coupling characteristics of the inductor under strong coupling, the measurement scheme in Fig. 11 is adopted. A high-frequency voltage excitation is applied between terminals 1 and 4 of Port 1, inducing a voltage across terminals 2 and 3. The displacement current generated by the electric-magnetic coupling between the inductors flows into Port 2 and is detected by the network analyzer.

The network analyzer employed in this study is the Keysight 5080A. Figs. 12(a)–(d) present the measured S -parameters (S_{11} , S_{12} , S_{21} , S_{22}). Based on the near-field coupling model in Fig. 8—comprising three Z -parameter modules (Z_1 , Z_2 , Z_3)—the measured S -parameters are first converted to Y -parameters via

Fig. 12. S -parameters under strong coupling. (a) S_{11} . (b) S_{12} . (c) S_{21} . (d) S_{22} .Fig. 13. Z -parameters curve under strong coupling. (a) Z_1 . (b) Z_2 . (c) Z_3 .

(19), then inverted to obtain the Z -parameters. The conversion results are illustrated in Fig. 13. As observed in Fig. 13(a), the high-frequency RLC parallel model ($R_{13}=7.9537\text{ k}\Omega$, $C_{13}=1.909\text{ pF}$, $L_{13}=3.333\text{ }\mu\text{H}$) accurately characterizes the impedance of Z_1 across 100 kHz–100 MHz. Both the measured Z_1 and the RLC model exhibit a linear impedance amplitude increase of 20 dB/dec within 100 kHz–30 MHz, where Z_1 behaves inductively. Beyond the resonance point, electric field coupling dominates, rendering Z_1 capacitive.

As shown in Fig. 13(b), the self-inductances L_{12} and L_{34} of the two inductors exhibit significant frequency dependence due to the magnetic core's complex permeability, leading to obvious oscillations in the impedance curves of (Z_2 and Z_3 within 100 kHz–1 MHz). Thus, fixed-value models are no longer applicable for the self-inductances in Z_2 and Z_3 ; instead, frequency-dependent models accounting for permeability variations must be adopted. The self-inductance values are calculated via the complex permeability curve combined with the inductor core's dimensional parameters, while the equivalent parallel capacitance C_{12} and C_{34} , and core loss resistance R_{12} and R_{34} are derived from the resonance point.

According to the comparison of the impedance curves Z_2

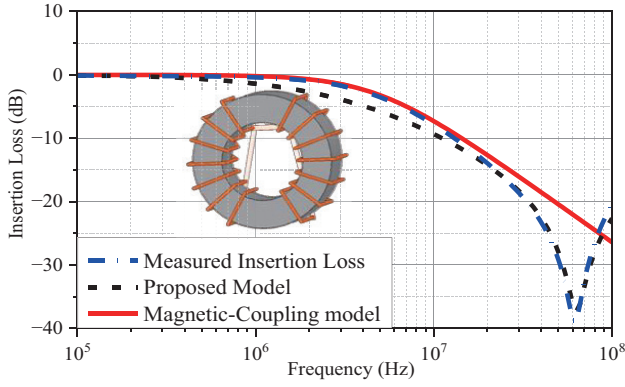


Fig. 14. Comparison of measured and theoretically calculated insertion loss curves.

and Z_3 shown in Fig. 13(b) and (c), it can be found that in the frequency range of 100 kHz–100 MHz, the measured curves of Z_2 and Z_3 are consistent with the curves output by the RLC parallel model. Therefore, the inductor model parameters with frequency-dependent characteristics can well characterize the impedance characteristics of the equivalent circuit.

Based on the obtained model parameters in Z_1 , Z_2 , and Z_3 , Fig. 14 shows the comparison results of the insertion loss curve measured by the network analyzer, the traditional model (only considering magnetic field coupling), and the proposed model (considering both electromagnetic coupling and the interaction between self-parameters and mutual parameters).

In the low-frequency band (100 kHz–1 MHz), the magnetic core's high permeability confines the main flux internally, with mutual inductance dominating. Thus, magnetic field coupling prevails in this band, and the three curves show good consistency. As frequency increases (1 MHz–10 MHz), mutual inductance decreases, while parasitic capacitances (between windings and between windings and the core) start to contribute, inducing combined electric and magnetic field coupling. In the high-frequency band (10 MHz–100 MHz), both the measured results and the proposed near-field coupling model exhibit obvious resonance points, which almost completely coincide. Therefore, the proposed model effectively characterizes the near-field coupling behavior of magnetic components across 100 kHz–100 MHz. The proposed model is also capable of accurately characterizing near-field coupling over the 1 kHz–100 kHz frequency range. The proposed model does not provide the prediction accuracy in the 1 kHz–100 kHz frequency range, because this range lies beyond the core bands of concern for EMI. For reference, the concerned frequency range for conducted EMI is 150 kHz to 30 MHz, and for radiated EMI, it is 30 MHz to 1 GHz.

B. Near-Field Transmission Characteristics Under Weak Coupling

When the distance between two CM inductors is large (mutual inductance coefficient $k < 0.1$), weak coupling exists between them. The insertion loss method in Fig. 15 is employed to evaluate their near-field coupling characteristics, with the measurement results presented in Fig. 16.

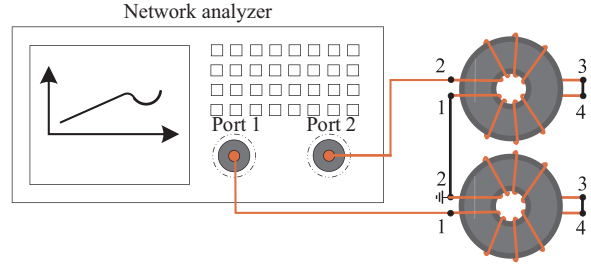


Fig. 15. Network analyzer measurement of insertion loss schematic diagram.

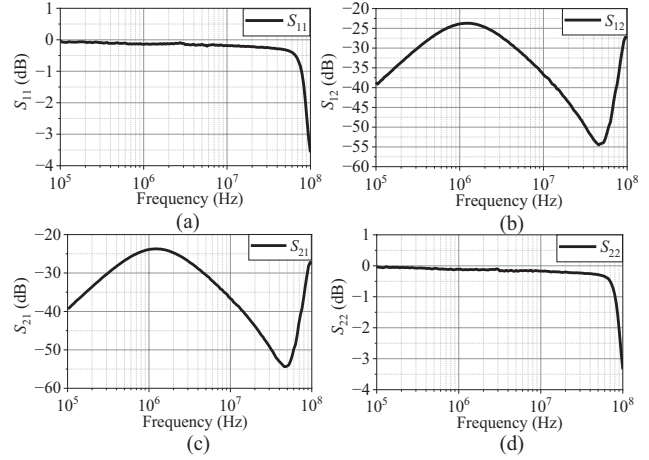


Fig. 16. S -parameters under weak coupling. (a) S_{11} . (b) S_{12} . (c) S_{21} . (d) S_{22} .

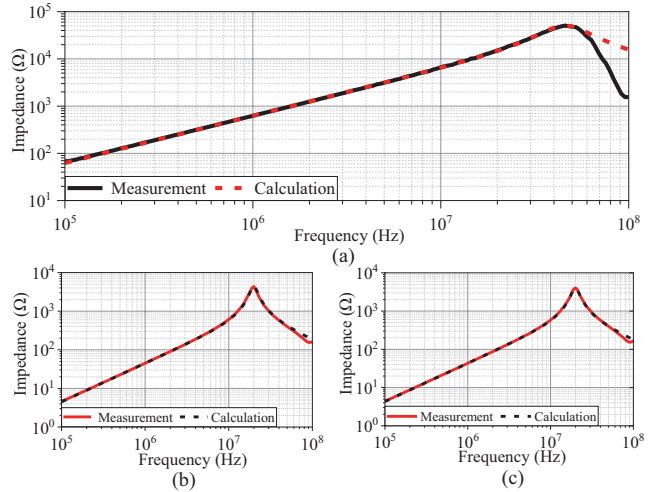


Fig. 17. Z -parameters curve under weak coupling. (a) Z_1 . (b) Z_2 . (c) Z_3 .

Similarly, the S -parameter to Z -parameter conversion results are presented in Fig. 17. As observed, the impedance curves of the three modules Z_1 , Z_2 , and Z_3 exhibit distinct LC resonance characteristics, enabling the RLC parallel model to characterize their Z -parameter curves. The relevant component parameter values are listed in Table II.

However, as illustrated in Fig. 17, Z_1 —one component of the proposed model—only exhibits good agreement with the measured results below 60 MHz. In contrast, the other components (Z_2 and Z_3) show consistency with experimental

TABLE II
PARAMETER COMPONENTS IN Z MODULES

Parameter	Near-field coupling equivalent parameter	
Z_1	$L_{13}/\mu\text{H}$	99.21
	$R_{13}/\text{k}\Omega$	50.703
	C_{13}/pF	0.1222
Z_2	$L_{12}/\mu\text{H}$	6.877
	$R_{12}/\text{k}\Omega$	4.0075
	C_{12}/pF	9.252
Z_3	$L_{34}/\mu\text{H}$	7.15
	$R_{34}/\text{k}\Omega$	4.3721
	C_{34}/pF	8.899

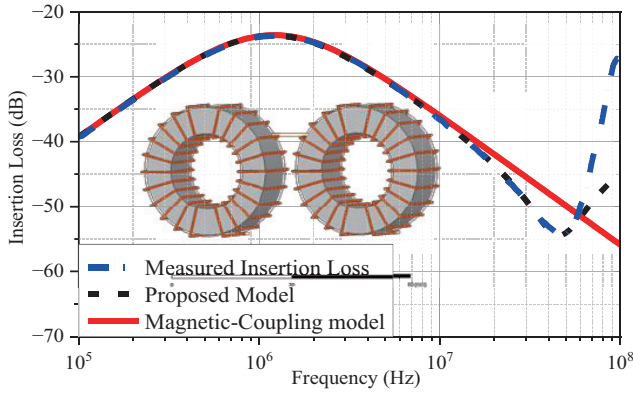


Fig. 18. Comparison of measured and theoretically calculated insertion loss curves.

measurements across the full 100 kHz–100 MHz range. Thus, comparing the measured and modeled results of Z_1 , Z_2 , and Z_3 enables identification of the model components that limit the predictive accuracy of the proposed model.

Baед on the above results, Fig. 18 compares the proposed wideband near-field coupling model, the magnetic field coupling model, and the measured insertion loss. The magnetic field coupling model can only characterize the near-field transmission characteristics within 100 kHz–10 MHz. In contrast, the proposed wideband model extends the effective frequency range to 60 MHz.

L , C , and R synergistically cover magnetic field coupling, electric field coupling, and loss mechanisms, respectively, enabling the model to comprehensively characterize near-field behavior under strong/weak coupling and wideband conditions. Their integration is the key to achieving the model's high accuracy (100 kHz–60 MHz) as verified by experiments.

Lumped parameter models for near-field coupling assume that magnetic and electric field coupling are independent and can be separated into lumped inductance L and lumped capacitance C . At GHz frequencies, the near-field region ($\leq \lambda/2\pi$) is drastically reduced in size, and the magnetic and electric field coupling becomes strongly intertwined due to electromagnetic wave propagation effects—rendering it impossible to decouple these two field components into discrete individual lumped parameters. For GHz applications, distributed parameter models (e.g., transmission line models) or full-wave

electromagnetic simulation (e.g., FDTD, FEM) are required to accurately characterize the component behavior.

V. CONCLUSIONS

This paper proposes a wideband near-field coupling modeling method for magnetic components considering electric-magnetic coupling and the frequency-dependent characteristics of magnetic core permeability. The correctness of the proposed modeling method is verified by comparing with measured results and traditional magnetic coupling models, which provides a good reference value for the prediction and optimization of near-field characteristics of magnetic components. The main conclusions are as follows:

1) Near-field coupling characteristics of magnetic components can be extracted via insertion loss measurement, and this method effectively captures electromagnetic coupling in near-field coupling.

2) The frequency-dependent characteristics of magnetic core permeability result in the self-parameters of magnetic components influencing the mutual parameters, with both sets of parameters affecting near-field coupling. Replacing fixed inductance with frequency-dependent $L(f)$ in the RLC parallel model ensures the self-parameter modules accurately characterize impedance characteristics across 100 kHz–100 MHz, resolving the high-frequency error issue of conventional models.

3) Whether magnetic components operate under strong or weak coupling, near-field coupling modeling must simultaneously account for both magnetic and electric field coupling, with the two fields dominating in distinct effective frequency ranges.

REFERENCES

- [1] P. S. Niklaus, D. Bortis, and J. W. Kolar, “Design and experimental analysis of a three-phase active CM/DM conducted EMI noise separator,” in *CPSS Transactions on Power Electronics and Applications*, vol. 5, no. 3, pp. 273–288, Sept. 2020.
- [2] S. Wang, F. C. Lee, J. D. van Wyk, and J. D. van Wyk, “A study of integration of parasitic cancellation techniques for EMI filter design with discrete components,” in *IEEE Transactions on Power Electronics*, vol. 23, no. 6, pp. 3094–3102, Nov. 2008.
- [3] S. Wang, F. C. Lee, and W. G. Odendaal, “Improving the performance of boost PFC EMI filters,” in *Eighteenth Annual IEEE Applied Power Electronics Conference and Exposition, 2003. APEC '03.*, Miami Beach, FL, USA, 2003, pp. 368–374.
- [4] S. Wang, F. C. Lee, and J. D. van Wyk, “Inductor winding capacitance cancellation using mutual capacitance concept for noise reduction application,” in *IEEE Transactions on Electromagnetic Compatibility*, vol. 48, no. 2, pp. 311–318, May 2006.
- [5] H. Zhao, Y. Li, Q. Lin, and S. Wang, “The parasitic capacitance of magnetic Components with Ferrite cores due to time-varying electromagnetic (EM) field,” in *2018 IEEE Energy Conversion Congress and Exposition (ECCE)*, Portland, OR, USA, 2018, pp. 3534–3541.
- [6] S. Wang, F. C. Lee, and J. D. van Wyk, “Integration of parasitic cancellation techniques for EMI filter design,” in *2008 Twenty-Third Annual IEEE Applied Power Electronics Conference and Exposition*, Austin, TX, USA, 2008, pp. 736–742.
- [7] J. He, J. Jiang and W. Chen, “Identification and model of near field magnetic coupling in a PFC converter,” in *2005 IEEE 36th Power Electronics Specialists Conference*, Recife, 2005, pp. 323–327.
- [8] K. Fu, J. Tu, W. Chen, and L. Huang, “Developed wide-band frequency transformer model for conducted common mode characteristics prediction,” in *CPSS Transactions on Power Electronics and Applications*, vol.

- 10, no. 1, pp. 110–119, Mar. 2025.
- [9] M. Antivachis, P. S. Niklaus, D. Bortis, and J. W. Kolar, “Input/output EMI filter design for three-phase ultra-high speed motor drive gan inverter stage,” in *CPSS Transactions on Power Electronics and Applications*, vol. 6, no. 1, pp. 74–92, Mar. 2021.
- [10] Q. Chen, F. Ye, and W. Chen, “The prediction of conducted EMI on switched mode power supply based on full-circuit simulation,” in *2015 IEEE 2nd International Future Energy Electronics Conference (IFEEEC)*, Taipei, Taiwan, China, 2015, pp. 1–5.
- [11] W. Chen, L. Feng, H. Chen, and Z. Qian, “Near field coupling effects on conducted EMI in power converter,” in *2006 37th IEEE Power Electronics Specialists Conference*, Jeju, Korea (South), 2006, pp. 1–6.
- [12] N. Wang, Z. Yan, J. Tang, Z. Ning, B. Xiao, and H. Wang, “Study on prediction models of EMI filter with near-field coupling effect,” in *2018 12th International Symposium on Antennas, Propagation and EM Theory (ISAPE)*, Hangzhou, China, 2018, pp. 1–5.
- [13] G. Liang and R. Zhu, “Predictive analysis for radiated electromagnetic disturbance in MMC-HVDC valve hall,” in *CPSS Transactions on Power Electronics and Applications*, vol. 5, no. 2, pp. 126–134, Jun. 2020.
- [14] S. Wang, F. C. Lee, W. G. Odendaal, and J. D. van Wyk, “Improvement of EMI filter performance with parasitic coupling cancellation,” in *2005 IEEE 36th Power Electronics Specialists Conference*, Dresden, Germany, 2005, pp. 1780–1786.
- [15] S. Wang, F. C. Lee, D. Y. Chen, and W. G. Odendaal, “Effects of parasitic parameters on EMI filter performance,” in *IEEE Transactions on Power Electronics*, vol. 19, no. 3, pp. 869–877, May 2004.
- [16] Y. Li, S. Wang, H. Sheng, and S. Lakshmikanthan, “Reduction and cancellation techniques for the near field capacitive coupling and parasitic capacitance of inductors,” in *2018 IEEE Symposium on Electromagnetic Compatibility, Signal Integrity and Power Integrity (EMC, SI & PI)*, Long Beach, CA, USA, 2018, pp. 432–437.
- [17] Rengang Chen, J. D. van Wyk, S. Wang, and W. G. Odendaal, “Improving the Characteristics of integrated EMI filters by embedded conductive Layers,” in *IEEE Transactions on Power Electronics*, vol. 20, no. 3, pp. 611–619, May 2005.
- [18] S. Wang and F. C. Lee, “Common-mode noise reduction for power factor correction circuit with parasitic capacitance cancellation,” in *IEEE Transactions on Electromagnetic Compatibility*, vol. 49, no. 3, pp. 537–542, Aug. 2007.
- [19] S. Wang, F. C. Lee, and J. D. Van Wyk, “Design of inductor winding capacitance cancellation for EMI suppression,” in *2006 37th IEEE Power Electronics Specialists Conference*, Jeju, Korea (South), 2006, pp. 1–7.



Xuanran Huang was born in Anhui, China, in 2001. She is presently working towards her M.S. degree in the College of Electrical Engineering and Automation, Xiamen University of Technology, Xiamen, China. Her research direction is electromagnetic compatibility (EMC).



Kaining Fu was born in Xiamen, China, in 1993. He received his M.S. and Ph.D. degrees from Fuzhou University, Fuzhou, China, in 2018 and 2021, respectively. He is currently a lecturer at the College of Electrical Engineering and Automation, Xiamen University of Technology, Xiamen, China. He is a member of the Magnetic Technology Committee of the China Power Supply Society (CPSS). His current research interests include power conversion, high-frequency magnetics, EMI debugging and solutions, and electromagnetic field analysis and applications.