Optimization of Low-Frequency Magnetic Metamaterials for Efficiency Improvement in Magnetically Coupled Resonant Wireless Power Transfer Systems

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ABSTRACT With the extensive applications of magnetically coupled resonant wireless power transfer (WPT), improving the power transfer efficiency (PTE) of WPT systems is critical. Magnetic metamaterials (MTMs) can effectively improve the PTE. However, currently the design of MTMs is mainly based on microwave or optical theory, and their applicability in low-frequency WPT systems needs to be further explored. In this paper, a fast and effective method for the design and optimization of low-frequency MTM is provided. Firstly, based on circuit theory and mutual inductance calculation, the PTE of WPT systems is concluded in an equation that concerns the position of a 2-D MTM slab and compensation capacitance of MTM unit coils. Two WPT systems with large differences in coil size of the transmitter (Tx) and receiver (Rx) and the same size are considered. In addition, a directly separating variables method is proposed to optimize the MTM slab position and the compensation capacitance of unit coils to maximize PTE when the MTM slab position changes axially between Tx and Rx. It is found that the optimal MTM slab positions for the two systems are remarkably different, and the PTE impressively presents a capacitance splitting phenomenon with the change of compensation capacitance. Finally, two MTM-enhanced WPT systems are established to verify the theoretical calculation. The experiments show that the maximum PTE is significantly increased by 144% and 31% respectively by adopting optimized compensation capacitance compared with that without MTM for the two systems with different or same transceiver coil size.

INDEX TERMS Wireless power transfer (WPT), metamaterials (MTMs), circuit theory, mutual inductance calculation, power transfer efficiency (PTE), optimization.

I. INTRODUCTION

Magnetically coupled resonant wireless power transfer (WPT) systems have great prospects in many fields due to convenient charging, including for electric vehicles [1], biological implants [2], consumer electronics [3], unmanned aerial vehicles [4], robots [5] and sensor networks [6], et al. However, the power transfer efficiency (PTE) of the WPT system decreases sharply due to divergence of magnetic field when transfer distance increases, especially in the under-coupled region. Researchers have proposed many measures to improve PTE, such as impedance matching [7], frequency tracking [8], adding relay resonators [9], using magnetic materials (such as ferrite) [1], or shaping the magnetic field to improve the coupling between transmitter (Tx) and receiver (Rx) [10]. Alternatively, some scholars have proposed to utilize magnetic metamaterials (MTMs) to control the electromagnetic field (EMF) of WPT systems to enhance the coupling between Tx and Rx. MTMs are artificial composite materials which are composed of periodic...
structural units and exhibit negative magnetic permeability, negative permittivity or negative refraction characteristics. Owing to the property of amplifying evanescent waves and reducing the leakage of EMF, MTMs have been increasingly studied and applied in the WPT systems in recent years.

Currently, the study on MTM-inspired WPT systems mainly relies on microwave or optical theory, such as negative refraction effect [11], perfect lensing property [12], magnetic dipole coupling model [13], magneto-inductive wave (MIW) [14], and transformation optics [15]. At the operating frequency of the magnetically coupled resonant WPT system, the magnetic and electrical fields are mostly uncoupled, so only the effective permeability of the MTM needs to be controlled. The real value of the relative permeability determines the angle of magnetic field direction [16, 17], and the ideal permeability for WPT enhancement is \( \mu_r' = -1 \) (\( \cdot \)' denotes the real part operator) according to the perfect lensing property [18]. But some authors have proposed using the properties different from the perfect lensing condition for MTMs designs used in WPT [19], [20]. For example, in the reference [20], the MTM composed of periodic arrangements of ferrite loaded solenoids is studied by the simulated and experimental mutual impedances for the PTE improvement of WPT system under the operating frequency of 5.574 MHz. And it is pointed out that the \( \mu_r' \) of the MTM corresponding to the maximum mutual inductance value is not \( \mu_r' = -1 \) but \( \mu_r' = -0.61 \). Also, through the magnetic dipole coupling model reference [13] concludes that the PTE of the WPT system is a function of MTM position relative to Tx/Rx and the maximum PTE exists. However, the model is obtained based on the assumptions that Tx and Rx are point dipoles and their axes are not on the same straight line but parallel and separated by a certain distance, which is not applicable for near-field analysis of common aligned WPT systems. And the theories mentioned above (except MIW and transformation optics) are afflicted by several approximations [21], one of which is that the MTM slab is considered infinite in extent (Actually, it is difficult to satisfy in the practical fabrication of MTM, especially in low-frequency applications such as WPT.). For MIW, as a form of propagation, it can only be found in specific types of MTMs, or only used for qualitative analysis [22]; for transformation optics theory, MTMs designed by the theory are generally anisotropic, which poses great challenges for fabrication.

Furthermore, when the WPT system operates at the frequencies of low MHz, the applicability of the microwave or optical theory to the near-field coupling of the WPT system also needs to be further explored and they are difficult to be applied to design and optimize the systems. Therefore, some scholars have applied circuit theory. A multiscale wireless power transfer system using MTM is proposed [23], and the unit coil of the MTM is equivalently processed as an RLC circuit, and the added capacitors are utilized to adjust the operating frequency of the WPT system at 6.78 MHz. In [24], a whole MTM slab composed of multi units is equivalent to an RLC circuit, and then the circuit model of the MTM-enhanced WPT system operating near 6.78 MHz is provided. The capacitance added to the units can affect the quality factor and resonant frequency of the unit coils, which in turn would affect the PTE. However, the authors point out that the resonant frequency is difficult to calculate due to the interaction between the units. In [25], the selective shielding feature of MTM is used to eliminate the fundamental (85 kHz) and harmonic (255 kHz) leakage magnetic fields around a WPT system, and the circuit equation of the WPT system with an MTM slab composed of 3 \times 3 units is offered. However, the equation is simplifies into a 5 \times 5 matrix from the original 11 \times 11 matrix where the coupling between symmetrical units is directly ignored, which would greatly affect the calculation results. The PTE of a four-coil system with MTM at 13.56 MHz based on circuit theory is analyzed in [26], and the EMF distribution with or without MTM is given. The experiments display that the PTE achieves the maximum when the MTM slab is located at middle position between Tx and Rx. However, both PTE calculation and EMF distribution are utilized to qualitatively analyze and in-depth quantitative study combined with circuit theory is lacking.

In this paper, two WPT systems with large differences in coil size of Tx and Rx and the same size are considered. The operating frequency of the WPT system is 1 MHz, and thus each MTM unit works in the realm of deep subwavelength. The positions of the Tx and Rx of the WPT system are fixed, and an MTM slab is moved axially between Tx and Rx. Different compensation capacitors are adopted for MTM unit coils (shown in Fig. 1) to achieve the maximum PTE. Firstly, the PTE of the system is concluded in an equation that concerns the MTM slab position and the compensation capacitance of the unit coils. Then, a directly separating variables method is proposed to optimize the MTM slab position and the compensation capacitance to maximum the PTE. And the PTE of the WPT system with an MTM slab using the optimized capacitors is compared with that of the system with an MTM slab using the capacitor corresponding to \( \mu_r' = -1 \) and that of the system without MTM. There are four significant findings: 1) For the two WPT systems with large differences in coil size of Tx and Rx and the same size, the optimal MTM slab positions are remarkably different regardless of calculations and experiments. 2) The change of optimal compensation capacitance of unit coils is
obtained when the MTM is moved axially between Tx and Rx. And the PTE presents an impressive capacitance splitting phenomenon to the compensation capacitance when MTM slab locating at any position. 3) For the two WPT systems, the effect of the MTM slab using the capacitors corresponding to $\mu_r' = -1$ on the PTE is obviously different, and the MTM slab can improve the PTE of the former system while reduce the PTE of the latter system. 4) The experimental maximum PTE of the WPT system with the optimized capacitors can achieve 144% and 31% increase respectively compared with that without MTM for the systems with different or same transceiver coil size.

The rest of the paper is organized as follows. The theoretical calculation method is provided by taking the MTM-enhanced WPT system with different transceiver coil sizes as an example: specifically, in Section II, the circuit model of the MTM-enhanced WPT system is introduced and the mutual inductance calculation between Tx/Rx and unit coils is offered; and a directly separating variables method is proposed to optimize the MTM slab position and the compensation capacitance of MTM unit coils to maximize PTE in Section III; Section IV reports the PTE comparison of WPT systems with an MTM slab using the optimal capacitors, with an MTM slab using the capacitors corresponding to $\mu_r' = -1$ and without MTM. Section V displays the experimental results for two MTM-inspired WPT systems with different or same transceiver coil size. And the last part is the conclusion.

II. DERIVATION OF PTE EXPRESSION

For the WPT system studied in this paper, the shape and position of Tx and Rx are fixed, and the MTM slab is moved axially between Tx and Rx. The added compensation capacitors are changed to achieve the maximum PTE, as presented in Fig. 1. Firstly, the model of the MTM-inspired WPT system is first provided based on circuit theory, and then the expression of the PTE can be obtained through matrix operations. The expression includes the mutual inductance parameters (i.e., the mutual inductance between Tx/Rx and unit coils) that changes with MTM slab positions in addition to the compensation capacitance. So the second step is to calculate the mutual inductance between Tx/Rx and the MTM unit coils.

A. CIRCUIT MODEL OF MTM-INSPIRED WPT SYSTEM

An MTM slab containing $3 \times 3$ matrix units is used as an example for description. The unit is composed of a disc coil and FR4 board (see Fig. 1). The number of turns of the disc coil is 3. An added compensation capacitor is used to adjust its resonant frequency. The unit coil can be equivalent to an RLC circuit. The model of the MTM-inspired WPT system can be obtained as shown in Fig. 2 combined with the circuit model of the WPT system. In the Fig. 2, $R_M$, $L_M$ and $C_M$ are the internal resistance, inductance and tuning capacitance of the unit coil, respectively. The capacitance includes parasitic inter-turn capacitance and additional compensation capacitance, and the parasitic inter-turn capacitance can be neglected as it is very small compared with compensation capacitance. $Z_S$, $Z_L$ are the source impedance and load impedance, respectively. $R_T$($R_R$), $L_T$($L_R$) and $C_T$($C_R$) are the internal resistance, inductance and tuning capacitance of Tx (Rx), respectively. L-shaped matching networks are added on the power supply side and the load side to reduce the power loss caused by the mismatch and improve the PTE of the two-coil WPT system. $L_{SM}$($C_{SM}$), $L_{LM}$($C_{LM}$) are the inductance (capacitance) of the matching network for the source and load side, respectively. $M_{TR}$ is the mutual inductance between Tx and Rx. $M_{T_1}$ and $M_{R_3}$ are the mutual inductance betweenTx and No. 5 unit coil (displayed in Fig. 3 (b)) and between Rx and No. 3 unit coil, respectively. $M_{1_2}$ is the mutual inductance between No. 1 and No. 2 unit coils.

The following equations can be obtained for the Tx circuit with a matching circuit according to Kirchhoff’s voltage law...
and Kirchhoff’s Current law,

\[
\begin{align*}
\dot{U}_{in} &= (R_T + joL_T + \frac{1}{joC_T} + \frac{1}{joC_{SM}})\dot{I}_T + joM_{T,R}\dot{I}_R \\
&+ \sum_{i=1}^{9} joM_{T,M_i}\dot{I}_{M_i} \\
\dot{U}_S &= \dot{U}_{in} + Z_S(\dot{I}_T + \frac{\dot{U}_{in}}{joL_{SM}})
\end{align*}
\]  

(1)

Combining the two equations yields

\[
\dot{U}_S = [(R_T + joL_T + \frac{1}{joC_T} + \frac{1}{joC_{SM}})(1 + \frac{Z_S}{joL_{SM}}) + Z_S]\dot{I}_T \\
+ joM_{T,R}(1 + \frac{Z_S}{joL_{SM}})\dot{I}_R + (1 + \frac{Z_S}{joL_{SM}})\sum_{i=1}^{9} joM_{T,M_i}\dot{I}_{M_i}
\]  

(2)

Define \(Z_T = (R_T + joL_T + \frac{1}{joC_T} + \frac{1}{joC_{SM}})(1 + \frac{Z_S}{joL_{SM}}) + Z_S\), and \(F_M = 1 + \frac{Z_S}{joL_{SM}}\), then formula (2) can be written as

\[
\dot{U}_S = Z_T\dot{I}_T + joM_{T,R}F_M\dot{I}_R + F_M\sum_{i=1}^{9} joM_{T,M_i}\dot{I}_{M_i}
\]  

(3)

In the same way, the following equation can be obtained for Rx,

\[
Z_R\dot{I}_R + joM_{T,R}\dot{I}_T + \sum_{i=1}^{9} joM_{R,M_i}\dot{I}_{M_i} = 0
\]  

(4)

where \(Z_R = R_R + joL_R + \frac{1}{joC_R} + \frac{1}{joL_{LM}} + joL_{L3}\). And for the unit coil (taking No. 5 coil as an example), the following equation can be obtained,

\[
Z_{M5}i_{M5} + joM_{T,M5}\dot{I}_T + joM_{R,M5}\dot{I}_R \\
+ \sum_{i=1}^{9} joM_{M5,M_i}\dot{I}_{M_i} = 0
\]  

(5)

where \(Z_{M5} = R_{M5} + joL_{M5} + \frac{1}{joC_{M5}}\). The internal resistance and inductance are equal for all the unit coils. The added compensation capacitors are assumed equal for all the unit coils, so \(Z_{M_i} = Z_M (i = 1, 2, 3, \ldots 9)\). The matrix equation (6), as shown at the bottom of the page, can be obtained by combining equations (3), (4) and (5), as shown at the bottom of Page 4. When Tx and Rx resonate, there are \(L_T = 1/\omega C_T\) and \(o_L = 1/\omega C_R\). And \(Z_S\) and \(Z_L\) are both taken as 50 \(\Omega\). The load current can be obtained by formula (7).

\[
\dot{I}_L = \frac{joL_{LM}}{Z_L + joL_{LM}}\dot{I}_R
\]  

(7)

Due to symmetry, the currents in unit coils No.1, No.3, No.7 and No.9 can be considered equal, and the currents in unit coils No.2, No.4, No.6, No.8 can also be considered equal. In addition, the coupling between unit No.1, No.3, No.7, No.9 and Tx (or Rx) is also considered equal, i.e. \(M_{T,M1} = M_{T,M3} = M_{T,M7} = M_{T,M9}\), and \(M_{R,M1} = M_{R,M3} = M_{R,M7} = M_{R,M9}\). Similarly, we can get \(M_{T,M2} = M_{T,M4} = M_{T,M6} = M_{T,M8}\), and \(M_{R,M2} = M_{R,M4} = M_{R,M6} = M_{R,M8}\). Furthermore, the mutual inductance between MTM unit coils can be divided into five types, as shown in Table 1. In this way, equation (6) can be further simplified, and the solution of equation (6) can be obtained by matrix operations as formula (8), as shown at the bottom of this page.
TABLE 1. Mutual inductance between MTM unit coils.

| Unit position | Type 1 | Type 2 | Type 3 | Type 4 | Type 5 |
|---------------|--------|--------|--------|--------|--------|
| Mutual inductance | $M_{type1}$ | $M_{type2}$ | $M_{type3}$ | $M_{type4}$ | $M_{type5}$ |

The PTE of the system can be expressed as

$$\eta = \frac{I^2 R_L}{I^2 R_T + (4 \times I^2_{M1} + 4 \times I^2_{M2} + I^2_{M5})R_M + I^2_R R_T + I^2_L R_L}$$

(9)

B. CALCULATION METHOD OF MUTUAL INDUCTANCE BETWEEN TX/RX AND UNIT COILS

To obtain the PTE of the system, all the mutual inductances among Tx, Rx and MTM unit coils are required in addition to the $R$, $L$, and $C$ parameters of the circuit (shown in Fig. 2). The $R$, $L$, $C$ parameters and the mutual inductance $M_{TM}$ and five types of mutual inductance between unit coils can be obtained only through one electromagnetic simulation (using software such as ANSYS MAXWELL/Q3D) or experiment. However, as the MTM slab is moved coaxially between the Tx and Rx, the mutual inductance of Tx/Rx with unit coils (i.e. $M_{T,M1}$, $M_{T,M2}$, $M_{T,M5}$ and $M_{R,M1}$, $M_{R,M2}$, $M_{R,M5}$) also changes. To obtain these mutual inductance values as a function of MTM slab positions, a sufficient number of simulations are required, which would be extremely time-consuming. Therefore, the analytical expressions of these mutual inductances with the change of MTM slab positions is to be derived through mutual inductance theory. The MTM slab position is denoted by the distance between Tx and MTM slab ($d_{TM}$). The distance between MTM slab and Rx can be expressed as $d_{M} = d - d_{TM}$, where $d$ is the distance between Tx and Rx. The calculation of $M_{T,M1}$, $M_{T,M2}$, $M_{T,M5}$ and $M_{R,M1}$, $M_{R,M2}$, $M_{R,M5}$ is divide into two categories, one is the aligned mutual inductance, such as $M_{T,M5}$ and $M_{R,M5}$, and the other is the non-aligned mutual inductance. It should be noted that there are two ferrite plates on the outside of the coils of Tx and Rx to improve the quality factor, as presented in Fig. 3 (For the case without ferrite plates, the calculation of mutual inductance would be simpler, and is not repeated in this paper.). For the two WPT systems with large differences in the coil size of the Tx and Rx and the same size, the calculation process of the former is only offered. In addition, during the two categories of mutual inductance calculation, the multi-turn coils of Tx (Rx, or MTM unit) are always regarded as multi-turn coaxial rings. The calculation of mutual inductance between two single-turn rings is as follows.

1) CALCULATION OF ALIGNED MUTUAL INDUCTANCE

For calculation of aligned mutual inductance $M_{T,M5}$ and $M_{R,M5}$ ($M_{T,M5}$ is taken as an example), the calculation expression is as follows [27], [28],

$$M_{TM,5} = \text{imag} \left\{ \frac{\mu_0 \pi}{h_T h_M \ln(r_{T,O}/r_{T,1}) \ln(r_{M,O}/r_{M,1})} \right. \times \int_0^\infty S(kr_{T,1}, kr_{T,O}) Q(k h_T, k h_M) e^{-k \phi} d\phi \right. \times \left[ \int_0^\pi r_{M,1} r_{M,0} - d_{offset} \frac{\cos \phi}{D} J_1(kD) d\phi \right] d\phi$$

(10)

where

$$Z_e' = \frac{\mu_0}{h_T h_M \ln(r_{T,O}/r_{T,1}) \ln(r_{M,O}/r_{M,1})} \times \int_0^\infty S(kr_{T,1}, kr_{T,O}) Q(k h_T, k h_M) \lambda(t) d\phi \times \left[ \int_0^\pi r_{M,1} r_{M,0} - d_{offset} \frac{\cos \phi}{D} J_1(kD) d\phi \right] d\phi$$

(11)

which characterizes the influence of ferrite plates. And $Q(k h_T, k h_M) = \frac{2}{k^2}(\cosh k h_T h_M - \cosh k h_T h_M), \lambda(t) = \frac{\phi(k) - \frac{1}{2} \cos \phi}{1 - \phi(k) - \frac{1}{2} \cos \phi}, \phi(k) = \frac{h_T h_M}{k d_{TM} + \frac{1}{k} - \frac{1}{2} \cos \phi}$, $\eta = \sqrt{k^2 + \lambda(0) \mu_\sigma F}$, and $t$ is the thickness of the ferrite plate. $\mu_\sigma F$ and $\sigma F$ are the relative permeability and electrical conductivity of the ferrite, respectively. $d_{offset}$ is the distance from the ferrite plate to the Tx (or Rx), here $d_{offset} = s/2$ as the ferrite plate fits tightly to the coil, where $s$ is the diameter of the coil wire. $h_T$, $h_M$ are the thicknesses of the Tx and MTM unit coil wiring, respectively. $h_M$ is considered equal to $s$ in this paper. $r_{T,1}$ ($r_{T,O}$), $r_{M,1}$ ($r_{M,O}$) are the inner (outer) radii of the Tx and unit coil, respectively. The main parameters of the WPT system is displayed in Table. 2.

2) CALCULATION OF UNALIGNED MUTUAL INDUCTANCE

For the calculation of unaligned mutual inductances $M_{T,M1}$, $M_{T,M2}$, $M_{R,M1}$ and $M_{R,M2}$, $M_{R,M1}$ is taken as an example, and the calculation expression is as follows [27], [28],

$$M_{TM,1} = \text{imag} \left\{ \frac{\mu_0}{h_T h_M \ln(r_{T,O}/r_{T,1}) \ln(r_{M,O}/r_{M,1})} \right. \times \int_0^\infty S(kr_{T,1}, kr_{T,O}) Q(k h_T, k h_M) e^{-k \phi} d\phi \right. \times \left[ \int_0^\pi r_{M,1} r_{M,0} - d_{offset} \frac{\cos \phi}{D} J_1(kD) d\phi \right] d\phi$$

(12)

where

$$Z_e' = \frac{j \mu_0}{h_T h_M \ln(r_{T,O}/r_{T,1}) \ln(r_{M,O}/r_{M,1})} \times \int_0^\infty S(kr_{T,1}, kr_{T,O}) Q(k h_T, k h_M) \lambda(t) d\phi \times \left[ \int_0^\pi r_{M,1} r_{M,0} - d_{offset} \frac{\cos \phi}{D} J_1(kD) d\phi \right] d\phi$$

(13)

where $D = \sqrt{r_{M,1}^2 + d_{offset}^2 - 2r_{M,0}d_{offset} \cos \phi}$, and $d_{offset}$ is the horizontal offset distance of the unit coil relative to Tx, which is equal to $\sqrt{2} l$, where $l$ is the length of the MTM unit.

Finally, after the mutual inductance between two single-turn rings is obtained, the total mutual inductance can be
calculated by equation (14) (taking the mutual inductance between Tx and No.5 unit coil as an example) [29].

\[ M_{T,Ms}^{total} = \sum_{i=1}^{N_T} \sum_{j=1}^{N_M} M_{T,Ms} \]  \hspace{1cm} (14)

where \( N_T \) and \( N_M \) are the turns of Tx and unit coil respectively.

III. OPTIMIZATION OF PTE CONCERNING COMPENSATION CAPACITANCE OF MTM UNIT COILS AND MTM SLAB POSITIONS

The function of the PTE concerning MTM slab positions (i.e. \( d_{TM} \)) can be obtained combining the formulas (7-9) and (10-14). To find the optimal solution of the PTE with respect to \( d_{TM} \), the derivative of PTE with respect to \( d_{TM} \) is required to be zero. Therefore the following equation can be obtained,

\[
\frac{\partial \eta}{\partial d_{TM}} = \frac{\partial M_{T,M1}}{\partial d_{TM}} \frac{\partial M_{T,M1}}{\partial d_{TM}} + \frac{\partial M_{T,M2}}{\partial d_{TM}} \frac{\partial M_{T,M2}}{\partial d_{TM}} + \frac{\partial M_{R,M5}}{\partial d_{TM}} \frac{\partial M_{R,M5}}{\partial d_{TM}} = 0
\]

(15)

On the other hand, the PTE is also a function of the compensation capacitance \( C_M \) of MTM unit coils. To find the optimal \( C_M \) corresponding to the maximum PTE, the following equation can be obtained,

\[
\frac{\partial \eta}{\partial C_M} = \frac{\partial M}{\partial Z_M} \frac{\partial Z_M}{\partial C_M} = \frac{\partial \eta}{\partial Z_M} \frac{\partial Z_M}{\partial C_M} = 0
\]

(16)

It can be seen from equation (15) and (16) that the analytical expression of PTE concerning \( d_{TM} \) and \( C_M \) is extremely complex, which makes it difficult to solve by analytical method. In addition, it is also arduous to solve it by general optimization methods. To solve the problem, a directly separating variables method is proposed. The basic idea of the method is: when the MTM slab is moved coaxially between Tx and Rx, its positions can only be changed limitedly in practice. The distance between Tx and Rx is separated by a finite number of points with equal distances. And at each point the maximum PTE with respect to the \( C_M \) is found, which is called the local optimal solution of PTE. By comparing all the local optimal solutions to find the largest value, the global optimal solution of PTE can be obtained. During finding local optimal solution of PTE, a limited number of separations are also performed on \( C_M \). Therefore, it is important to determine the variation range and separation step of \( d_{TM} \) and \( C_M \) for ultimate optimal PTE. In this paper, \( d = 250 \) mm (displayed in Table. 2), i.e., Tx is located at zero position and Rx is situated at \( d = 250 \) mm. The separation step of \( d_{TM} \) is set as 10 mm, which is considered to be small enough in practice. Meanwhile, considering the special point of the middle position between Tx and Rx, the variation range of \( d_{TM} \) is finally set as 10 mm-240 mm with separation step of 10 mm and the midpoint of 125 mm included. For determination of the \( C_M \) range, the self inductance of the unit coil is firstly obtained (\( \mu_M \)) by simulation since the MTM property of amplifying evanescent waves is generally affected by the resonance of unit coil. The corresponding tuning capacitance at the operating frequency (1MHz) of the WPT system is calculated (the value is 20.51 nF). Its variation range around the tuning capacitance is fully expanded and the final variation range of \( C_M \) is determined as \( 1 \times 10^{-6} \text{nF} \sim 300 \text{nF} \), and the separation step is small enough (taking it as 0.001 nF). The specific optimization process is displayed in Fig. 4.

In the specific optimization, the main parameters can be seen in Table. 2. Some circuit parameters such as resistance are measured experimentally (see Table. 3). Firstly, the values of the five types of mutual inductance between the unit coils are obtained by simulation, as shown in Table. 2. It can be seen that the values of \( M_{Type3}, M_{Type4}, M_{Type5} \) are obviously smaller than those of \( M_{Type1}, M_{Type2} \). But all five types of mutual inductance are considered in the PTE calculation to acquire accurate results. Then the mutual inductance between Tx (or Rx) and unit coils is obtained when the MTM slab is moved, as presented in Fig. 5.

It can be seen that when the MTM slab is moved coaxially from Tx side to Rx side, the mutual inductance between Tx and unit coil No. 5 gradually decreases, and the mutual inductance between Rx and No. 5 gradually increases; the mutual inductance between Tx and No. 2 first increases and then decreases, and the maximum value is at \( d_{TM} = 20 \text{ mm} \); the mutual inductance value between Tx and No. 1 (or between Rx and No. 1 or No. 2) appears negative, which
FIGURE 5. Change of mutual inductance between Tx (or Rx) and unit coils with change of MTM slab position.

TABLE 2. Main parameters of WPT system with large differences in coil size of Tx and Rx.

| Symbol | Physical meaning | Value |
|--------|------------------|-------|
| $r_{T,0}$, $r_{R,0}$ | Outer radius of certain ring for Tx or Rx | For outermost ring 100 mm, 50 mm |
| $r_{T,1}$, $r_{R,1}$ | Inner radius of outermost ring for Tx or Rx | For outermost ring 100 mm, 50 mm |
| $r_{M,0}$, $r_{e,1}$ | Outer and inner radius of certain unit coil for Rx | For outermost ring 48 mm, 20 mm |
| $h_M$ | Thickness of MTM unit coil | 0.035 mm |
| $w_M$ | Strip width of MTM unit coil | 3 mm |
| $s_s$ | Inter-turn spacing of Tx (Rx) and unit coils | 2.1 mm, 5 mm |
| $d$ | Distance between Tx and Rx | 250 mm |
| $d_{TM}$, $d_{RM}$ | Distance between Tx and unit coils, or between Rx and unit coils | Changed |
| $l$ | Length of MTM unit | 100 mm |
| $N_t$, $N_R$, $N_M$ | Turns of Tx, Rx and unit coil | 8, 6, 3 |
| $\mu_0$ | Permeability of free space | $4\pi \times 10^{-7}$ H/m |
| $\mu_{ef}$, $\sigma_T$ | Relative permeability and electrical conductivity of ferrite plates | 2500, 0.01 S/m |
| $l_f$, $h_f$ | Length and thickness of ferrite plates | 300 mm, 3 mm |
| $M_{T12}$, $M_{T21}$, $M_{R12}$, $M_{R21}$, $M_{M12}$, $M_{M21}$ | Five types of mutual inductance between MTM unit coils | $-4.11$ nH, $-10.7$ nH, $-4.13$ nH, $-3.71$ nH and $-2.90$ nH |

is due to the large horizontal offset between Tx and No. 1 (or between Rx and No. 1 or No. 2. Thus the magnetic flux emitted by Tx (or Rx) mainly passes through coil No. 1 or No. 2 in the opposite direction. It should be noted that the $M_{T,0}$ value when the MTM slab is very close to Tx is smaller than $M_{R,0}$ when the MTM slab is very close to Rx, which is owing to the large difference in size between Tx and unit coil. Part of the magnetic flux received by Tx from No. 5 cancels out when the MTM slab is very close to Tx, resulting in a decrease in mutual inductance.

Furthermore, the compensation capacitor of the unit coil is experimentally realized by a capacitor bank composed of multiple capacitors in parallel (see Fig. 9(b)). The internal resistance of the compensation capacitor varies due to different values and quantities of capacitors. It is found that the resistance of the compensation capacitor varies between 350-550 mΩ by experiment. So through the optimization method, the corresponding optimal capacitance is obtained, which is reduced from 19.35 nF to 19.02 nF. The capacitance variation is small, and the optimal positions are always at $d_{TM} = 130$ mm. So the resistance value of the compensation capacitor is determined as 450 mΩ for simplicity. The maximum PTE of the system by optimization is obtained at $d_{TM} = 130$ mm and $C_M = 19.16$ nF, and the maximum PTE reaches 8.6%, which is 161% higher than the PTE of 3.3% without MTM.

FIGURE 6. Variation of PTE with MTM slab position and compensation capacitance of MTM unit coils: (a) optimal capacitance at different MTM slab positions, (b) PTE as a function of compensation capacitance of MTM unit coils.

The optimal capacitance obtained by the optimization method at different MTM slab positions is also obtained, as shown in Fig. 6 (a). On the whole, the optimal capacitance varies between 0-23 nF. The optimal capacitance is large and is close to the tuning capacitance (20.51 nF) when the MTM slab is located near the middle position between Tx and Rx ($d_{TM} = 125$ mm). The optimal capacitance is far away from the resonant capacitance when MTM slab is close to Tx or Rx. This difference reflects two important roles of MTM in improving PTE of the WPT system: (1) near the middle position, MTM mainly amplifies the evanescent wave, so the optimal capacitance is close to the tuning capacitance;
(2) when it is close to Tx or Rx, the MTM should minimize its own resistive losses, so the optimal capacitance should be away from the tuning capacitance. In addition, the optimal capacitance increases abruptly at $d_{TM} = 160$ or $170$ mm, which can be explained by Fig. 6(b).

Fig. 6(b) shows the PTE as a function of the compensation capacitance of unit coils when the MTM slab is located at some positions. It can be seen that the PTE presents a capacitance splitting phenomenon to the compensation capacitance just like frequency splitting phenomenon, that is, there is a local minimum value of the PTE near the tuning capacitance of the unit coil, and two local maximum values appear on both sides of it. The apparent discontinuous points at $d_{TM} = 160$ and $170$ mm in Fig. 6(a) are precisely owing to the fact that the maximum PTE has shifted from the local maximum point on the left side of the curve to the local maximum point on the right side. It can also be found that the local minimum of PTE occurs near the tuning capacitance, which is mainly due to the large resistance loss caused by the large currents in the MTM unit coils.

IV. PTE COMPARISON OF WPT SYSTEMS WITH TWO KINDS OF MTM SLABS AND WITHOUT MTM

To further illustrate the PTE improvement by the optimal capacitance obtained through the optimization method in this paper, the PTE of the WPT system containing an MTM slab with the optimal capacitors on the unit coils is compared with that of the WPT system containing an MTM slab with negative relative permeability which is designed by microwave theory.

A. CONSTRUCTION OF MTM WITH NEGATIVE RELATIVE PERMEABILITY

The same MTM slab as the former but with different compensation capacitors on the unit coils is adopted. The compensation capacitor is determined by the S-parameter retrieval method to acquire the MTM with its real part of the equivalent relative permeability being negative (i.e., $\mu'_r < 0$), preferably equaling to $-1$. Firstly, a simulation model of the MTM unit is built in software ANSYS HFSS as shown in Fig. 7(a). The Perfect E Boundary, Perfect H Boundary and Wave Port are added onto the planes of the air domain parallel to the XOZ plane, XOY plane and YOZ plane, respectively. The $S_{11}$ and $S_{21}$ values can be obtained by simulation. The wave impedance $z$ and refractive index $n_{\text{eff}}$ of the MTM can be obtained by combing formulas (17), (18) with (19) [30],

$$z = \pm \sqrt{(1 + S_{11})^2 - S_{21}^2(1 - S_{11})^2 - S_{21}^2}$$  \hspace{1cm} (17)

$$n_{\text{eff}} = \frac{1}{k_0d_M} \left\{ \left[ \ln(e^{jkd_M}) \right]'' + 2m\pi \right\} - i\left[ \ln(e^{jkd_M}) \right]'$$ \hspace{1cm} (18)

$$e^{jkd_M} = \frac{S_{21}}{1 - S_{11}\frac{z - 1}{z + 1}}$$ \hspace{1cm} (19)

where $k_0$ denotes the wave number of the incident wave in free space, $d_M$ is the thickness of the homogeneous slab for the unit of the MTM, and $m$ is an integer related to the branch index of $n'$, and $(\cdot)'$ and $(\cdot)''$ denote the real part and imaginary part operators, respectively. $\mu'_r$ of the MTM can be obtained according to the formula (20) [31]. Based on the method above and through multiple HFSS simulations and calculations, it is found that when the compensation capacitance is set to 24.25 nF, Fig. 7(b) can be obtained and it can be observed $\mu'_r = -1.028$ at 1 MHz.

$$\mu'_r = n_{\text{eff}}z$$ \hspace{1cm} (20)

B. PTE COMPARISON

The PTE of the WPT system having an MTM slab with the optimal compensation capacitors is compared with that of the
system having an MTM slab with compensation capacitors corresponding to $\mu_r' = -1$ and with that without MTM, as displayed in Fig. 8. It can be seen that adding the optimal capacitance to the unit coils can always improve the PTE when the MTM slab is located at any position between Tx and Rx. Especially in the vicinity of the middle position, the PTE improvement is distinct. And it is found that the maximum PTE is not achieved at middle position (i.e., $d_{TM} = 125$ mm) but at a point near the middle position and biased towards to Rx position ($d_{TM} = 130$ mm), which is the most likely location of the weakly coupled region between Tx and Rx for the system with large Tx coil and small Rx coil. In addition, adding an MTM slab with $\mu_r' = -1$ in the WPT system, the PTE can be improved when the slab is near the middle position but biased towards Rx position (the maximum PTE is obtained at $d_{TM} = 150$ mm). Whereas the MTM slab reduces the PTE of the WPT system when the slab is close to Tx or Rx.

V. EXPERIMENTAL RESULTS FOR TWO MTM-INSPIRED WPT SYSTEMS

Two WPT systems with large differences in coil size of Tx and Rx and the same size are considered. Firstly, the software ADS (Advanced Design System) is utilized to acquire the parameters for L-shaped matching circuits on the source and load sides of the WPT system (see Fig. 2) to achieve impedance matching. The obtained parameters are shown in Table. 3. The experiment setup is shown in Fig. 9 and Fig. 12. The Keysight (Santa Rosa, CA, USA) E5061B Vector Network Analyzer is applied to measure S-parameters. The measured $S_{11}$ and $S_{22}$ values for the two WPT systems without MTM slab are both less than $-19$ dB at 1 MHz (Fig. 9 (d) displays the $S_{11}$ and $S_{22}$ of system with different transceiver coil sizes), indicating the systems have achieved impedance matching. Finally, the PTE of the WPT systems can be obtained by the formula (21) [32].

$$\eta = \frac{|S_{21}|^2}{1 - |S_{11}|^2} \times 100\% \quad (21)$$

During the experiment, the minimum separation for the MTM slab from Tx/Rx should be chosen without degrading the impedance matching. In this work, the separation of 70 mm is chosen, which is the closest distance to the Tx or Rx from MTM not much degrading the impedance matching [32]. Furthermore, it cannot be moved any closer to Tx than 90 mm due to the physical constraints imposed by the shape of the system (shown in Fig. 9 (a) and Fig. 12). So the position variation range of the MTM slab is set between 90 mm and 180 mm with separation step of 10 mm including the special point at midpoint of 125 mm. The total of 11 sets of measurements are obtained.

A. MTM-INSPIRED WPT SYSTEM WITH DIFFERENT TRANSCIEVER COIL SIZES

The experimental setup of the WPT systems with large differences in coil size of Tx and Rx is presented in Fig. 9. The corresponding parameters are displayed in Table. 3. The PTE changes of the WPT system with the compensation capacitance of the MTM coils when the MTM slab is located at $d_{TM} = 130$ mm, 140 mm, 160 mm, are obtained as shown in Fig. 10. The theoretical calculated values are also displayed for comparison.

Fig. 10 presents that the experimental trend of the system PTE with increasing capacitance is in good agreement with the theoretical calculations. In particular, the capacitance splitting can both be observed near the tuning capacitance. When the compensation capacitance increases from 0 to
330 nF, the PTE of the system first increases, then the capacitance splitting phenomenon occurs near the resonant capacitance. After that, it decreases rapidly. Finally it increases gradually and tends to a certain PTE value which is lower than that without MTM. This is mainly owing to the fact that at this time the evanescent wave amplification of MTM does not play an important role but resistive losses increase.

Further observing the capacitance splitting phenomenon shows that for $d_{TM} = 130$ mm and 140 mm the experimental maximum PTE is obtained at the local maximum point on the left side, which is the same as the theoretical calculation. The differences between calculations and experiments are mainly due to the following three reasons in addition to the manufacturing errors of the experimental setup and measurement errors: 1) in the theoretical calculation of mutual inductance, some assumptions such as coaxial rings are adopted; 2) the $L$ and $C$ parameters used in the experimental L-type matching networks are approximate values of those obtained by ADS simulation owing to the limited values of practical inductors and capacitors. And the experimental values of the compensation capacitor banks are also the approximation of the calculated values; 3) affected by the MTM slab, it is difficult for $S_{22}$ values to be small enough when the MTM slab is close to Rx.

Fig. 11 displays the theoretical calculations and experiments of the PTE change of the WPT system with different MTM slab positions. “$C_M$ Optimized” indicates that the corresponding optimized capacitance is adopted for $C_M$ at each MTM slab position (see Fig. 6(a)). “$C_M$ Single-Optimized” means that the single compensation capacitor value corresponding to the maximum PTE of the system is always adopted at any MTM slab position. “$C_M\mu_r = -1$” represents the MTM slab with $\mu_r = -1$. It can be seen that the PTE change with the MTM slab position by experiment and theoretical calculation is consistent though there are some differences in values. The optimal position corresponding to the maximum PTE by experiment is

| Parameter symbol | Physical meaning | Transceiver coils of different size | Transceiver coils of same size |
|------------------|------------------|------------------------------------|--------------------------------|
| $f_0$            | Operating frequency | 1 MHz                             |                                |
| $L_{T}/L_{R}$    | Inductance of Tx (Rx) | 39.620$\mu$F/9.22$\mu$F | 39.620$\mu$F/39.703$\mu$F |
| $C_{T}/C_{R}$    | Tuning capacitance of Tx (Rx) | 633pF/2.75nF | 633pF/630pF |
| $R_{T}/R_{R}$    | Internal resistance of Tx (Rx) | 7.05$\Omega$/1.4$\Omega$ | 7.05$\Omega$/7.52$\Omega$ |
| $L_{SM}, R_{M}$  | Inductance and internal resistance of MTM unit coil | 1.235$\mu$H/0.45$\mu$F |                                |
| $L_{SM}, C_{SM}$ | Inductance and capacitance of source-side matching network | 3.2$\mu$H/9.7nF | 3.2$\mu$H/9.7nF |
| $L_{SM}/C_{SM}$  | Inductance and capacitance of load-side matching network | 1.32$\mu$H/20nF | 3.3$\mu$H/9.1nF |
| $M_{TR}$         | Mutual inductance between Tx and Rx | 188.7nH | 955nH |
| $C_{SM}$         | Tuning capacitance of MTM unit coil | 20.51 nF |                                |
| $C_{M, \mu_r = -1}$ | Capacitance of MTM unit coil corresponding to $\mu_r = -1$ | 24.25 nF |                                |
| $C_{M, Optimized}$ | Optimal capacitance of MTM unit coil | 19.16 nF | 18.57 nF |

\[ d_{TM} = 140 \text{ mm}, \] which is slightly different from 130 mm by theoretical calculation. But obviously, the optimal position is not at middle position but at the point near the middle position and bias towards Rx position regardless of the theoretical calculations or experiments. In addition, the optimal PTE can be achieved by optimizing $C_M$ at each MTM slab position, but it is cumbersome to change the compensation capacitor value when the MTM slab position is changed. Yet approximately same PTE values as that by using optimized $C_M$ can be achieved by adopting single-optimized $C_M$ around middle positions.

Finally, it can be seen that the calculated and experimental PTEs of the system without MTM are low, mainly due to the long transmission distance, which is 5 times the radius of Rx, and the low quality factor of the coils of Tx.
Rx and MTM unit, which are measured to be 35, 41 and 17 respectively at 1 MHz. However, after adding the MTM slab with optimized compensation capacitors, the system PTE can be significantly increased by 144%, which reaches 6.1%. And the maximum system PTE with the MTM slab with $\mu'_r = -1$ by experiment is 4.7%, which is acquired at $d_{TM} = 150$ mm. Furthermore, in addition to the error mentioned above, the reason for the disparity between the experiments and theoretical calculations in Fig. 11 is due to the influence of the internal resistance of different compensation capacitor banks, resulting in a slight difference between the actual optimal capacitance and the theoretically calculated value (as illuminated in the third-to-last paragraph of Section III).

**B. MTM-INSPIRED WPT SYSTEM WITH SAME TRANSCIEVER COIL SIZE**

In the same way, the optimal MTM slab position and the optimal compensation capacitance of MTM unit coils for the MTM-inspired WPT system with the same size of the transceiver coils are obtained. The optimization process is not repeated here. The comparison of PTE of the WPT system with the change of the MTM slab position through theoretical calculations and experiments is presented in Fig. 13 (a). The experimental measurement setup is shown in Fig. 12, and the specific parameters are shown in Table 3.

As shown in Fig. 13 (a), the calculated optimal MTM slab position corresponding to the maximum PTE is $d_{TM} = 120$ mm which is approximately at the middle position between Tx and Rx. Furthermore, the MTM slab with $\mu'_r = -1$ reduces the PTE of the WPT system at any MTM position. Comparing the calculations and experiments shows that the change trend of the two is very consistent, though the optimal position by experiment is slightly different from that by theoretical calculation. The optimal position is obviously different from that in the MTM-inspired WPT system with different transceiver coil sizes, which is mainly due to the difference in the location of the weakly coupled regions between Tx and Rx. The optimal position of the MTM slab is located near the center position between Tx and Rx but biased towards Rx position for the system with different transceiver coil sizes, while the optimal position is near the middle position for the system with same transceiver coil size. Furthermore, the maximum PTE obtained by experiment is at $d_{TM} = 125$ mm and $C_M = 18.57$ nF, and the maximum PTE is 16.8%, which is increased by 31% compared with that without MTM. The maximum PTE of the WPT system containing the MTM slab with $\mu'_r = -1$ by experiment is 8.2%, which is acquired at $d_{TM} = 110$ mm and less than that without MTM (12.8%), as expected by theoretical calculations in the Fig. 13(a). And this is apparently shown in Fig. 13 (b), which displays the theoretically calculated PTEs as a function of compensation capacitance of MTM unit coils when the MTM slab is located at several positions near the middle position where the system PTE is most likely to be improved.

Table 4 lists the performance comparison of this work with previously reported studies. As PTE is closely related to the transfer distance and the size of Tx and Rx, a normalized distance is defined as the ratio of the transfer distance and square root of product of Tx radius and Rx radius for fair comparison. As shown in Table 4, the PTE of the WPT system is greatly improved when a single MTM slab with optimal compensation capacitors is located around the optimized position. In addition, the optimal capacitors and optimized positions of MTM slab can be expected before experiment by theoretical calculation, which evidently surpasses the previous studies where the optimal MTM positions can only be found through multiple experiments.
TABLE 4. Performance comparison of this work with other MTM-enhanced WPT systems.

| Ref. | Size of Tx/Rx | Operating frequency | Normalized distance | MTM configuration | MTM property | PTE without/with MTM | MTM position for maximum PTE | Whether under guidance of theoretical calculation |
|------|---------------|---------------------|---------------------|------------------|--------------|----------------------|-----------------------------|------------------------------------------|
| [16] | Same          | 3 MHz               | 3.2                 | 3×3 array, 2 slabs, single sided | μ′r = -1   | 49.7% /72.1%        | Near middle position but biased towards Rx dtx=560mm, d=800mm     | ×                                         |
| [17] | Same          | 6.78 MHz            | ~2.7                | 5×5 array, 1 slab, double sided  | μ′r = 0.1  | 10.7% /54.9%        | Not Given                    | ×                                         |
| [20] | Same          | 5.57 MHz            | 2                   | 3D structure, 1 slab           | μ′r = 1    | 20% /35%            | Near Tx                     | dtx=5mm, d=45mm                     | ×                                         |
| [23] | Same          | 6.78 MHz            | 5.6                 | 2×2 array, 2 slabs, single sided | μ′r = 1    | 5.4% /50.1%        | Near Tx                     | dtx=10mm, d=50mm                    | ×                                         |
| [24] | Same          | 6.78 MHz            | ~1.2                | 8×8 array, 1 slab, double sided | Negative μ′r | 8.2% /18.2%        | Near middle position but biased towards Tx dtx=40mm, d=100mm           | ×                                         |
| [26] | Same          | 13.56 MHz           | 5                   | 3×3 array, 1 slab, single sided | μ′r = 1.11 | 9.5% /25.2%        | Middle position              | ×                                         |
| This work | Large Tx, small Rx | 1 MHz | 3.5 | 3×3 array, 1 slab, single sided | With Cm optimized (With μ′r =-1) | 2.5% /6.1% (2.5% /4.7%) | Near middle position but biased towards Rx dtx=140mm, d=250mm (dtx=150mm, d=250mm) | ×                                         |
|      | Same          | 1 MHz               | 2.5                 | 3×3 array, 1 slab, single sided | With Cm optimized (With μ′r =-1) | 12.8% /16.8% (12.8% /8.2%) | Near middle position dtx=125mm, d=250mm (dtx=110mm, d=250mm) | ○                                         |

VI. CONCLUSION

The expression of PTE of the WPT system with respect to 2-D MTM slab position and unit coil compensation capacitance is derived combining circuit model and mutual inductance calculation. The optimal MTM slab position and the optimal compensation capacitance have been found to achieve maximum PTE by a directly separating variables method for two WPT systems with different or same coil size in Tx and Rx. The maximum PTE is increased by 144% and 31% respectively by adopting optimal compensation capacitance compared with that without MTM for the two systems. And it can be realized that the PTE of the WPT system is always greater than that without MTM when an MTM slab with optimized compensation capacitance is inserted at any position between Tx and Rx.

Furthermore, it is found that the MTM slab with μ′r = −1 designed by microwave theory (specifically S-parameter retrieval method) has different effects on the PTE of the two systems working at 1 MHz. For the WPT system with large Tx coil and small Rx coil, the MTM slab can improve the PTE when locating near the middle position between Tx and Rx, whereas for the WPT system with the transceiver coils of the same size the slab reduces the PTE when locating at any position. This is because the retrieval method is mainly based on some assumptions, such as the infinite flat slab composed of periodic units, the vertical incidence of electromagnetic waves, etc. And the WPT system with a small Rx coil in this paper is more inclined to satisfy the assumption of infinity for the MTM slab relative to the system with the large Rx coil. This also exposes the limitations of the retrieval method for the design of MTM with finite units working in the near field of the low-frequency WPT system.

The theoretical computation and optimization method proposed in this paper, however, can be widely applicable to the WPT systems containing MTM slabs with different dimensions or different numbers of units, flat or curved structures for not relying on the assumptions mentioned above, and thus the method provide an important approach for the design of low-frequency MTM. Especially, the research on the MTM-enhanced WPT system with large Tx size and small Rx size presents a guidance for the design of the MTM applied to the medical implantable WPT system where Rx coil size is generally small.

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