Fourier Analysis and Loss Modeling for Inductive Wireless Electric Vehicle Charging with Reduced Stray Field

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Abstract—With the growth of electric vehicle (EV) popularity, different charging options to increase user convenience and reduce charging times are being considered and researched. Among these, inductive wireless power transfer (WPT) systems for EVs are being designed to meet specifications such as stray field, power level, efficiency, misalignment tolerance, and ground clearance, which are all heavily influenced by the coil geometry. The proposed Fourier Analysis Method (FAM) is an analytical method to directly design coil geometries to meet stray field and power level requirements through an optimization of Fourier basis function coefficients. The outputs of the optimization are complex, planar coil geometries that meet the power level and stray field constraints with minimized loss factors. Contours of these potentials determine the coil conductor paths and loss models predict the system efficiency and performance over misalignment. The Fourier representation of the geometry is used to conveniently calculate the coupling over misalignment, external proximity effect loss, and ferrite loss. A 6.6 kWh prototype WPT system with low stray field and high efficiency is built from the optimization results to validate the models and showcase the usefulness of the FAM design approach.

Index Terms—wireless power transfer, inductive power transmission, coil design, electric vehicles, Fourier analysis

I. INTRODUCTION

INDUCTIVE wireless power transfer (WPT) is useful in a variety of applications including the automotive and transportation sectors. WPT is a safe, convenient, flexible, and efficient charging solution that can easily be automated. The design of WPT systems to meet specifications such as power level, coupling, airgap, misalignment tolerance, stray field, and efficiency requires the computation of the fields, coupling, and inductances of various coil geometries. In particular, meeting the 15 µT pacemaker limit or the 27 µT ICNIRP 2010 public exposure magnetic field limit at the nominal WPT frequency of 85 kHz [3] and radiated EMI field limits in CISPR 11 [4] depends heavily on the coil geometry. More complex coil geometries such as bipolar coils and coils with shielding turns have been demonstrated to achieve higher power levels under stray field limits [5], [6]. Integrating shielding turns into the coil geometry design has several advantages over adding conductive or magnetic materials around the coil area to reduce stray field. In [7], [8], many Litz wire shielding turn designs were more efficient than copper ring shields or shielding plates of aluminum. In [9], eddy-current shielding with aluminum sheets was shown to actually increase stray field for bipolar coil geometries. Likewise, adding ferrite teeth [10], aluminum sheets, or ferrite outside the coil geometry [9] decreases the mechanical airgap of the system or increases the effective coil area and weight.

Attempting to consider a wider range of coil geometries including shielding turns, as well as other design parameters such as ferrite thicknesses, number of turns, and conductor types, results in a large design space. With the rising number of iterations needed, optimization based on brute-force iteration in FEA with full or partial 3D-modeling similar to [5] become increasingly computationally expensive. Likewise, many analytical methods like [6], [7] are pertinent only to circular or rectangular coils and are not general enough to model a large variety of possible geometries and aspect ratios. This limitation is also present in ferrite [11] and external proximity effect loss models [12].
To address these issues, this work applies Fourier analysis to WPT coil design and loss modeling. Magnetic component modeling and optimization using Fourier basis functions is already well-known in the design of magnetic resonance imaging (MRI) gradient coils [13], [14], fusion devices [15], and electric machines [16], [17]. In these works, basis functions, objectives, and constraints are added to regularize each coil design problem as an optimization to converge to a unique, optimal solution. For example, [14] receives a desired MRI gradient field distribution at a distance and iterates over basis function spaces to optimize for objectives such as minimum power dissipation, stored energy, and maximum current density. In these works, Fourier basis functions were shown to be well suited for symmetrical coil geometries and take advantage of symmetry to reduce matrix sizes, number of iterations, and overall computational times compared to other types of basis functions.

In the field of inductive wireless power transfer, Fourier analysis has been used to analytically calculate fields and mutual inductance of coils, but not to optimize coil designs including loss calculations. In [18], it is used to analyze circular and rectangular filament coils backed by magnetic or conductive media and predict mutual inductance. Similar modeling is used to model closely spaced rectangular coils in [19]. These works showcase the applicability of Fourier functions and analysis to model WPT coil geometries. Furthermore, many of the common types of WPT coil geometries are symmetric such as circular, rectangular, and bipolar geometries.

This work instead applies Fourier basis functions and analysis to optimize coil geometries to meet objectives of minimized current or loss factor under constraints of power level, stray field, and coil dimensions. The optimized geometry is implemented in a discretized coil design by selecting the ferrite and litz wire with the loss models. This work extends the Fourier Analysis Method (FAM) coil design method first published in [1] by adding loss and misalignment models. The methodology is experimentally validated with a 6.6 kW prototype including efficiency and field measurements over different airgaps and alignments.

The remainder of this work is organized as follows: Section II details the Fourier Analysis Method (FAM) for coil design optimization. Section III describes a misalignment model for translational and rotational misalignment and the application of loss models to the FAM including external proximity effect, ferrite losses, and others. To validate these models, the experimental results of a 6.6 kW prototype are given in Section IV. Finally, comparisons of this prototype with the literature are summarized in Table VII and conclusions are given in Section V.

II. THE FOURIER ANALYSIS METHOD

The Fourier Analysis Method (FAM) is an analytical method to directly design coil geometries through an optimization of Fourier basis function coefficients to meet specifications such as power level, coil dimension, and stray field. The use of Fourier basis functions allows rapid computation of the currents, field, and coupling of a wide range of symmetric coil shapes. The outputs of the optimization are the magnetic scalar potentials of coil geometries that meet the power, stray field, and dimensional constraints with minimum current magnitude. This loss factor corresponds to the total amount of loss in the geometry. This first optimization step of the FAM design process in Fig. 1(a) is an inverse design step that directly optimizes the coil geometry from the input parameters by iterating the weights of the Fourier basis functions to minimize the loss factor objective function within the constraints. The optimization outputs are then used to flexibly determine coil conductor paths and adjust the impedance of the coils for varying numbers of turns without changing the overall coil shape. Afterwards, the number of turns is added to the Litz wire type, ferrite thickness and losses are calculated over misalignment to compare options and finally select the design.

In comparison, Fig. 1(b) summarizes the conventional coil design approaches that sweep over geometric parameters and evaluate the stray field and loss of each combination. Compared to three-dimensional FEA-based design methods like [5], the FAM is a two-dimensional approach, much less computationally complex, and faster. Compared to most analytical design methods like [6], [7], the FAM is more general and can analyze much more complex symmetric geometries than circular or rectangular coils. With these benefits, the FAM enables a more generalized coil design methodology that rapidly evaluates complex symmetric coil shapes, broadening the scope of design optimization for WPT systems.
The overall WPT system layout used in FAM is illustrated in Fig. 2 where the outer dimensions of the primary and secondary coils are \( x_{ext} \) and \( y_{ext} \), and the system airgap \( z_{gap} \), strain field limits outside a region with dimensions of \( x_{meas} \) and \( y_{meas} \), and magnetic scalar potential \( \Psi \) used in the Fourier Analysis Method (FAM). (b) FAM axes layout and key equations.

which can be rapidly computed by conventional Fast Fourier Transform algorithms.

For ferrite-backed coils, the potential of the coil surface is defined by the surface current boundary condition

\[
K = \nabla \times \hat{k} \Psi = \frac{\partial \Psi_z}{\partial y} i - \frac{\partial \Psi_y}{\partial x} j
\]

where \( \hat{i}, \hat{j}, \) and \( \hat{k} \) are unit vectors in the \( x, y, \) and \( z \) directions, respectively, and \( K \) is the surface current vector. This assumes the change in the magnetic potential and field in the ferrite are close to zero due to the low reluctance of the ferrite compared to the airgap. The surface currents in the \( x \) and \( y \)-direction, \( K_x \) and \( K_y \), are then

\[
K_x(x, y, 0) = \sum_{m,n=-N+1}^{N-1} j k_x \psi(m,n) e^{j(k_x x+k_y y)/4}
\]

\[
K_y(x, y, 0) = \sum_{m,n=-N+1}^{N-1} -j k_y \psi(m,n) e^{j(k_x x+k_y y)/4}
\]

This surface current representation allows for the coil geometry to be analyzed as a surface current density independent from the number of turns. To determine the coil conductor paths and current, the surface currents are grouped by dividing the net change of potential into a number of turns \( N_T \). The number of turns can be adapted to flexibly change the impedance of the coils to meet various driving voltages and loads. This results in the RMS current in each turn: \( I_1 \) for the primary or transmitter coil, and \( I_2 \) for the secondary or receiver coil.

\[
I_1 = (\max \Psi(x,y,0) - \min \Psi(x,y,0))/N_T.
\]

The conductor paths are the contours of the potential at values

\[
C = \min \Psi(x,y,0) + \left( 0 : (N_T - 1) + \frac{1}{2} \right) I_1.
\]

As in [1], the number of turns are limited by the outer diameter of the gauge of wire, \( d_{out} \), and the maximum current density, \( K_{max} \), by

\[
\max \Psi(x,y,0) - \min \Psi(x,y,0) = \frac{I_1}{d_{out}} < K_{max}
\]

so that the geometries fit in a single winding layer.

In the Fourier domain, the potential \( \Psi \) is differentiated to obtain algebraic relationships between the potential and the field \( B \).

\[
B = \mu_0 H = -\mu_0 \nabla \psi.
\]

Neglecting displacement current in quasi-magnetostatic conditions, the wavenumber in the \( z \)-direction, \( k_z \), is derived by observing that \( \nabla \times B = 0 \) in the absence of airgap currents. Here, it is assumed there are no conductors or volume currents between the primary and secondary coils with only air in the airgap. Combined with \( \nabla \cdot B = 0 \), the field and potential satisfy

\[
\nabla^2 \psi = \nabla^2 B = 0.
\]
Fig. 3. Diagrams of the 4 basis functions sets considered in the Fourier Analysis Method. (a) The symmetry conditions for each basis function set comprised of real and complex conjugate relationships. With these relationships, each basis function can be represented by one value in the first quadrant, limiting the number of variables and constraints needed in the optimization function. (b) Example of a \( \cos x \cos y \) basis function. (c) Example of a \( \sin x \cos y \) basis function. (d) Example of a \( \cos x \sin y \) basis function. (e) Example of a \( \sin x \sin y \) basis function.

Therefore, when real, non-zero wavenumbers exist in the \( x \) and \( y \)-directions, \( k_z \) is imaginary and is

\[
k_z = \pm \sqrt{-k_x^2 - k_y^2} = \pm j \gamma,
\]

where \( \gamma = \sqrt{k_x^2 + k_y^2} \), and the magnetic potential in the airgap must satisfy

\[
\frac{\partial^2 \Psi}{\partial z^2} - k^2 \Psi = 0
\]

which has solution

\[
\Psi(z) = c_1 e^{-\gamma z} + c_2 e^{\gamma z}.
\]

The constants \( c_1 \) and \( c_2 \) are found using the boundary conditions at \( \Psi(0) \) and \( \Psi(z_{gap}) \), yielding the relationship [16],

\[
\Psi(z) = \frac{\sinh \gamma z}{\sinh \gamma z_{gap}} \Psi(z_{gap}) = \frac{\sinh \gamma(z - z_{gap})}{\sinh \gamma z_{gap}} \Psi(0).
\]

By (9)–(13), coils with potentials with higher \( k_x \) and \( k_y \), i.e. with shorter wavelengths, will consist of components that have higher \( k_z \) or \( \gamma \) and decay faster in the \( z \)-direction than those with more lower \( k_x \) and \( k_y \) potentials. This scattering relationship describes how coils with larger diameters have fields that decay slower away from the coil surface than those of smaller coils. An example of this for a rectangular primary coil is shown in Fig. 4. At the surface of the coil, the DFT of the potential and discretized potential are shown in Fig. 4(a)-4(b). The \( z \)-field at the surface of the coil by (16) is given in Fig. 4(c)-(d). The \( z \)-field from the primary at the surface of the secondary at \( z_{gap} \) is shown in Fig. 4(e)-(f), where only the low-frequency components of the field are largely remaining. By combining (8) with (13), the fields at \( z \) are a function of \( z_{gap} \) and \( \gamma \) for ferrite-backed coils with single-sided flux generation.
A. Optimization of Stray Field and Current

Using the FAM framework, an optimization is formulated and solved to design coil geometries to minimize the loss factor \( \Gamma_k^{\text{Loss}} \) for a fixed power level when bounded by coil extents at \( x_{\text{ext}} \) and \( y_{\text{ext}} \) and field limits at the measurement extents \( x_{\text{meas}} \) and \( y_{\text{meas}} \). For this optimization, the dimensions of the design space are \( D_x = D_y = 1.4 \) m with a resolution of 2 cm.

\[ \Gamma_k^{\text{Loss}}(\psi) \] corresponds with the total current magnitude in the coil geometry defined by

\[ \Gamma_k^{\text{Loss}}(\psi) = \int_{\Omega} K(x, y, 0)^2 d\Omega = \left( ||K_x(\psi)||_2^2 + ||K_y(\psi)||_2^2 \right)/16. \]
This is quickly calculated in the Fourier domain from the basis function coefficients by noting that the Fourier transform is a unitary function. This avoids the computation of $K(x, y, 0)$ in each objective function evaluation step. The integral of the current density squared multiplied by the equivalent resistance of the coil surface area is the coil conduction loss. The squared current integral is used as the loss factor because it is only a function of the geometry and is not dependent on the selection of the number of turns and conductors chosen later in the design process.

The objective function is formulated as the minimization of the loss factor $\Gamma_{K, Loss}^2$ added to the 1-norm of the magnitude of the basis function coefficients to eliminate small values of zero-valued basis functions such as $\sin 0 \cos y$ yielding the objective function

$$\frac{\Gamma_{K, Loss}^2(\psi)}{P} + \alpha \left\| \psi \right\|_1$$

where $\alpha = 0.1$ to minimize the weight of the zero-valued functions compared to the normalized loss factor and $P$ is the desired coil-coil power level.

Constraints are imposed on the optimization to achieve a specified power level, a stray field level, and coil dimensions or extents. The first constraint is the coil-coil power transfer,

$$\left( P - 2\pi f E_m(\psi) \right) / P \leq 0. \quad (24)$$

The next constraint uses the maximum average stray field magnitude $B_{str,max}(\psi)$,

$$(B_{str,max}(\psi, x_{meas}, y_{meas}) - B_{str,lim}) / B_{str,lim} \leq 0 \quad (25)$$

where

$$B_{str,max}(\psi, x_{meas}, y_{meas}) = \left\| B_{str,avg}(x, y) \right\|_{50}. \quad (26)$$

The inclusion of stray field as a constraint incorporates the need to add shielding for compliance with safety standards for public magnetic field exposure and EMI standards. $B_{str,max}(\psi, x_{meas}, y_{meas})$ is computed as the 50-norm of the spatial stray-field matrix, $B_{str,avg}$, which approximates the infinity norm or the maximum value of the matrix. $B_{str,avg}$ is the average field magnitude over the z-direction of the airgap $B_{avg}$ as in (20) outside the measurement extents $x_{meas}$ and $y_{meas}$. The use of the average or 1-norm of the field across the airgap in the z-direction accounts for the stray fields in the entire airgap such as near the surface of the primary coil, the middle of the airgap, and at the surface of the other coil across the airgap. A higher-order norm could also be used to approximate the maximum across the airgap in the z-direction, but the 1-norm results in a smoothed field matrix similar to the field at the middle of the airgap.

The third constraint limits the current density to the desired coil extents such that the surface integral of the stray current squared $\kappa_{str}^2(\psi, x_{ext}, y_{ext}) = \int_{x_{ext}} \int_{y_{ext}} K_{str}(x, y, 0)^2 d\Omega$ outside the coil extents $x_{ext}$ and $y_{ext}$, is a small percentage, $\beta = 10^{-4}$ of the surface integral of the total current squared $\Gamma_{K, Loss}^2(\psi)$.

$$\frac{\kappa_{str}^2(\psi, x_{ext}, y_{ext})}{\Gamma_{K, Loss}^2(\psi)} - \beta \leq 0 \quad (27)$$

This constraint is included to restrict the optimized coil geometry to the specified coil dimensions, $x_{ext}$ and $y_{ext}$.

In summary, the objective function and constraints form the optimization

$$\min \left( \frac{\Gamma_{K, Loss}^2(\psi)}{P} + \alpha \left\| \psi \right\|_1 \right)$$

s.t.

$$(P - 2\pi f E_m(\psi)) / P \leq 0,$$

$$(B_{str,max}(\psi, x_{meas}, y_{meas}) - B_{str,lim}) / B_{str,lim} \leq 0,$$

$$\frac{\kappa_{str}^2(\psi, x_{ext}, y_{ext})}{\Gamma_{K, Loss}^2(\psi)} - \beta \leq 0. \quad (28)$$

This optimization was performed with the gradient-based fincon optimizer in MATLAB. The results of this optimization are plotted in Fig. 5 for $B_{str,lim}$ of 5 $\mu$T to 1 mT with $x_{ext} = 0.7$ m and $y_{ext} = 0.5$ m, and with $x_{meas} = 0.7$ m and $y_{meas} = 0.5$ m [1]. This approximately 4:3 aspect ratio for the coils and measurement distance was chosen to show the flexibility of the method to handle a large number of coil shapes and sizes as most analytical methods focus on 1:1 square coils or circular coils. It also allows for a difference between the bipolar $\cos x \sin y$ and $\sin x \cos y$ basis function results. The misalignment range in the SAE J2954 standard [3], ± 10 cm and ± 7.5 cm, also supports this type of aspect ratio, as the misalignment performance of the x-direction must be better than that of the y-direction to support parking. Vehicle alignment is easier to adjust in the front and back direction relative to the side to side direction. The coil dimensions are set by the airgap and the measurement distances were chosen to be the coil extents to make the stray fields in the optimization higher magnitude.

Contours of the outputs at $B_{str,lim} = 20 \mu$T and 1 mT are shown in Fig. 6. As shown, when $B_{str,lim} = 1$ mT in the geometries are simple rectangular, bipolar or double-D, and quadrupole shapes limited by the extents of the coil in Fig. 6(a)-6(d). When constrained by $B_{str,lim} = 20 \mu$T, the geometries develop smaller poles and shielding structures to provide flux cancellation to reduce the field outside the stray field boundary in Fig. 6(e)-6(h). The smaller poles reduce the coupling of the coils for the fixed airgap. Likewise, shielding structures slightly detract from the coupling and require additional current. These two factors increase the amount of current needed for a given amount of power for Fig. 6(e)-6(h) relative to the simpler structures of Fig. 6(a)-6(d). The coupling of the geometries decreases and the current required to achieve the 6.6 kW power level increases for the fixed coil extents and airgap.

Assuming matched currents in both coils, these results can be scaled to different power levels by multiplying the current density and fields of the solutions by the square root of the ratio of the desired power level and 6.6 kW. The optimization could also be run again with a different choice of power level $P$. To analyze the currents and fields of the geometries with unmatched current, such as when the system is operated with non-ideal loading or at different output powers with a fixed output voltage, the currents of each coil can be
Fig. 5. Optimization Outputs: The maximum average stray field magnitude vs. the square root of the integral of the current magnitude squared at 6.6 kW and coupling coefficient at $z_{gap} = 200$ mm at alignment for $x_{ext} = 0.7$ m and $y_{ext} = 0.5$ m for $B_{str,lim}$ of 5 $\mu$T to 1 mT with $x_{meas} = 0.7$ m and $y_{meas} = 0.5$ m [1].

Fig. 6. Plots of coil contours from each basis function for $z$ and coupling coefficient at $1.0$ m and $0.6$ m and $y_{meas} = 0.7$ m. As in Fig. 9, both the $\cos x\cos y$ and $\sin x\sin y$ outputs extend in the $x$-direction for $x_{meas} = 1.0$ m and $y_{meas} = 0.7$ m. Here, the $\sin x\sin y$ set does not converge below $B_{str,lim} = 20$ $\mu$T.

The best-performing coil designs depend on the values and aspect ratio of both the coil and field extents. Bipolar geometries perform better with their long-axis, or the direction over which the two poles lie, along larger coil extents and when more stray field is allowed on their long-axis. For example, the optimization outputs in Figs. 7-9 are with square coil extents $x_{ext} = 0.6$ m and $y_{ext} = 0.6$ m with a $B_{str,lim}$ of 3 $\mu$T to 200 $\mu$T at measurement extents of $x_{meas} = 0.7$ m and $y_{meas} = 0.7$ m and also for $x_{meas} = 1.0$ m and $y_{meas} = 0.7$ m. The stray field is lower than the previous optimization outputs in Fig. 5 as the measurement distances are slightly further away. For this optimization, $D_x = D_y = 1.4$ m with a resolution of 2 cm.

As in Figs. 7 and 8, the $\sin x\cos y$ set performs well when the measurement extents are $x_{meas} = 1.0$ m and $y_{meas} = 0.7$ m and the stray field can be larger in the $x$-direction. Meanwhile, the $\sin x\cos y$ and $\cos x\sin y$ outputs are identical when the measurement extents are the same in $x$ and $y$. As in Fig. 9, both the $\cos x\cos y$ and $\sin x\sin y$ outputs extend in the $x$-direction for $x_{meas} = 1.0$ m and $y_{meas} = 0.7$ m. Here, the $\sin x\sin y$ set does not converge below $B_{str,lim} = 20$ $\mu$T.
B. Comparison to FEA Sweep of Rectangular Coils

In Fig. 10, the \( \cos \theta \cos y \) FAM outputs in Figs. 5-6 are compared to a sweep of rectangular coil geometries in finite-element analysis (FEA) software. The FAM outputs are exported from FAM as discretized surface current densities and interpolated in FEA so that they could be compared in the same FEA setup. This sweep is similar to the approach of [5]. In Table I, the parameters of the simulations are shown. As seen, most of the FAM optimization outputs are on the Pareto front of the FEA rectangular geometry solutions. The FAM outputs are constrained by the selection of \( N \), \( D_x \), and \( D_y \) as they define the lowest and highest wavenumbers of basis functions used in the optimization. For this comparison, the stray field at the middle of the airgap was used as the comparison metric for ease of calculation in FEA instead of the average stray field for both the rectangular coil geometries and \( \cos \theta \cos y \) outputs. On the same computer, the 3D FEA simulation took 2.4 minutes to simulate each geometry during the sweep of rectangular coils, while each FAM function evaluation, which evaluates a candidate geometry in the MATLAB fmincon optimizer for the objective function and constraints, averaged 3.7 ms.

III. MISALIGNMENT AND LOSS MODELING

After the coil geometry is optimized, the misalignment performance and system efficiency can be evaluated to realize a physical design. The representation of the coil geometry in the Fourier domain can assist with several aspects of these calculations such as with calculating the coupling over misalignment and calculating the ferrite and wire loss which are dependent on the fields of the geometry. The loss models given here assume a series-series compensated system as realized in the experimental prototype.

A. Misalignment Model

The FAM can predict system performance in both translational and rotational misalignments by calculating the fields and mutual inductance of the system for misaligned conditions. For example, the efficiency over misalignment of the geometries of Fig. 16(a)-16(b) are given in Fig. 16(g)-16(h). In the Fourier domain, translational misalignment is modeled by adding phase-shift to the Fourier components, where \( x_{sft} \) and \( y_{sft} \) are the translational misalignment of the coil in the \( x \)-direction and \( y \)-direction respectively.

\[
\psi_{sft}(m,n) = \psi(m,n)e^{-(k_x x_{sft} - k_y y_{sft})}.
\]

An example of this operation for the selected geometry potential in Fig. 11(a) is shown in Fig. 11(c).

In a similar manner, rotation in the spatial domain produces rotation in the Fourier domain,

\[
[k_x' \ k_y'] = \begin{bmatrix}
\cos \theta_{sft} & -\sin \theta_{sft} \\
\sin \theta_{sft} & \cos \theta_{sft}
\end{bmatrix} [k_x \ k_y]
\]

where \( \theta_{sft} \) is the rotational misalignment and \( k_x' \) and \( k_y' \) are the rotated wavenumbers. An example of this is shown in Fig. 11(b). Due to the limited number of basis functions used

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**Fig. 7.** Optimization Outputs: The maximum average stray field magnitude vs. the square root of the integral of the current magnitude squared at 6.6 kW and coupling coefficient at \( z_{gap} = 200 \text{ mm} \) at alignment for \( x_{ext} = 0.6 \text{ m} \) and \( y_{ext} = 0.6 \text{ m} \) for \( B_{str,lim} \) of 3 \( \mu \text{T} \) to 200 \( \mu \text{T} \) with \( x_{meas} = 0.7 \text{ m} \) and \( y_{meas} = 0.7 \text{ m} \).

**TABLE I**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Min. (m)</th>
<th>Step (m)</th>
<th>Max. (m)</th>
<th>Points</th>
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<tr>
<td>Ferrite Dimensions</td>
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<td></td>
<td>1.4 m x 1.4 m</td>
<td></td>
</tr>
</tbody>
</table>
in the optimization, this operation is performed by taking the discrete Fourier transform of the rotated spatial domain potential, which has a finer discretization, to interpolate and determine the rotated Fourier components.

**B. Coil Conduction Losses**

As alternating current (AC) current flows through coupled transformer windings, the losses in the wire are

\[ P_w = \frac{1}{2} I_1^T I_2 \begin{bmatrix} R_{11} & 0 \\ 0 & R_{22} \end{bmatrix} I_2^T \]  

(31)

where \( I_1 \) and \( I_2 \) are the current phasors of the primary and secondary coils respectively [20]. In inductive power transfer, the relative phase shift between the current in the windings will be near 90° such that the mutual resistance terms are neglected. This leaves only the self-resistance of the coils, \( R_{11} \) and \( R_{22} \).

The increase in self-resistance of a conductor carrying AC is comprised of two effects: the skin effect and the proximity effect. The skin effect is caused by the change in the magnetic field within the conductor due to the change in current within that wire as in Ampere’s Law and is characterized by the skin depth

\[ \delta_s = \sqrt{\frac{\rho}{\pi \mu f}} = \sqrt{\frac{2\rho}{\omega \mu}} \]  

(32)

which is defined as the depth in which the field and current density in the conductor falls to \( e^{-1} \) of its initial value at the surface of a conductor with resistivity \( \rho \) and permeability \( \mu \) when conducting AC at frequency \( \omega = 2\pi f \).

Given the crowding of the current and the associated increase in current density, the skin depth is used in the calculation of the total AC resistance of WPT coil conductors due to the skin effect. For an individual circular conductor with a ratio of strand diameter \( d_{str} \) and skin depth \( \delta_s \) the ratio \( F_R(\zeta) = r_s/r_{DC} \) is

\[ F_R = \frac{\zeta}{2\sqrt{\zeta}} \left( \frac{\beta_0}{\beta_1} - \frac{\beta_0}{\beta_1} \right) \frac{\beta_1}{\beta_1^2 + \beta_1^2} \]  

(33)
Due to the skin effect, the resistance of solid conductors increases rapidly with frequency. Therefore, in this work, Litz wire is used to reduce the AC resistance relative to solid wire. Litz wire is comprised of fine, insulated strands bundled together. When these strands are around or less than a skin depth in diameter, the impact of skin effect can be reduced at cost of decreased packing factor and increased cost. The comprehensive comparison and selection of Litz wire stranding as was done in [23] integrated into FAM is left for future work and Litz wire with 40 AWG strands are used in this work based on availability. Within a circular Litz wire cable, there are \( n \) conductors of diameter \( d_{str} \). Once \( F_R \) is computed, the total resistance of the Litz wire including skin effect and DC-resistance, \( R_s \), is

\[
R_s = \left( \frac{r_{DC} \cdot F_R(\zeta)}{n} \right) L_T \cdot 1.015^{N_b} \cdot 1.025^{N_c} \tag{36}
\]

where \( r_{DC} \) is the resistance of one strand in the Litz wire per unit length and \( n \) is the total number of parallel strands. Here \( L_T \) is the total length of wire in the coil multiplied by additional factors to account for the additional length of each conductor due to the number of bundling operations \( N_b \) and number of cabling operations \( N_c \) as given in the manufacturer datasheet [24].

Losses also occur due to the effect of external fields from conductors near each other, known as the proximity effect. For Litz wire, this is subdivided into two primary categories: internal proximity effect losses and external proximity effect losses. The increase of resistance from these two effects are governed by the factor \( G_R \) as defined in (37) which is a function of \( \zeta \) derived in a similar fashion to \( F_R \) [21], [22].

\[
G_R = -\frac{\pi^2 d_{str}^2}{2} \left( \text{ber}_2(\zeta)\text{ber}_1(\zeta) + \text{ber}_2(\zeta)\text{bei}_1(\zeta) \right) + \frac{\text{bei}_2(\zeta)\text{bei}_1(\zeta) - \text{bei}_2(\zeta)\text{bei}_1(\zeta)}{\text{ber}_0(\zeta)^2 + \text{bei}_0(\zeta)^2} \tag{37}
\]

Internal proximity effect losses occur within Litz wire cables because the magnetic field generated from the total current within the cross section of the cable varies the distribution of current within individual strands, increasing radially from center of the cable. The strands are bundled and cabled to vary the position of each strand over the length of Litz wire. The external proximity effect is due to the effect of the total field of the coil on each section of a conductor. For ferrite-backed coils with a non-zero gap between the ferrite and windings, the external field \( \vec{H}_e \) is the field on each section of the wire. The cross-product of this and the wire direction yields the field orthogonal to the winding section direction unit vector \( \vec{d}r \). This includes all \( H_e \) components, as the coils are assumed to be in the \( x-y \) plane, and \( H_x \) and \( H_y \) components depending on the direction of the section of wire [12]. The fields are calculated by FAM by (14)-(16). The result of this line integration of the field along the coil contours of the wire is taken and normalized by the RMS current in the wire, \( I_1 \) or \( I_2 \), to yield the external proximity effect resistance. The summation of these two terms yields the total increase in resistance due to the proximity effect for Litz wire with outer diameter \( d_{str} \),

\[
R_{prox} = n \cdot r_{DC} \cdot G_R(\zeta) \cdot 1.015^{N_b} \cdot 1.025^{N_c} \cdot \left( \int |\vec{H}_e \times \vec{d}r|/I_1^2 + \frac{1}{2\pi^2 d_{str}^2} L_T \right). \tag{38}
\]

An example of the coil contours and fields used to compute (38) for \( I_1 = 17.3 \) A is given in Fig. 12. The external fields and direction of each section of wire are calculated with an interpolation of the field values calculated by the FAM to determine orthogonal field components. Once the resistance increases due to the skin effect and proximity effect

Fig. 10. Comparison of the \( \cos x \cos y \) FAM outputs from the 70 cm x 50 cm optimization with a sweep of rectangular coils: The stray field maximum at the middle of the airgap taken at \( x_{meas} = 0.8 \) m and \( y_{meas} = 0.6 \) m vs. (a) the current norm at 6.6 kW and (b) coupling coefficient at \( z_{gap} = 200 \) mm at alignment.

where

\[
\zeta = \frac{d_{str}}{\sqrt{2}\delta}
\]

as derived in [21], [22], where \( r_s \) is the AC resistance of a circular strand of wire including skin effect and \( r_{DC} \) is the DC resistance. Equation (33) is comprised of Kelvin functions that separate the real and imaginary parts of the value of Bessel functions of the first kind \( J_v(\cdot) \) of order \( v \) with complex argument \( j^{3/2}x \) as in

\[
J_v(j^{3/2}x) = \text{ber}_v(x) + j\text{bei}_v(x). \tag{35}
\]
are calculated, the total AC resistance of each coil at a given frequency, $R_{11}$ and $R_{22}$, are found by the addition of $R_s$ and $R_{prox}$.

This methodology is used to determine the AC resistance of Litz wire coils with complex geometries. In each case, the overall length of Litz wire and the total external proximity effect is conveniently determined by a line integral of the conductor contours. The orthogonal external fields along the contours are interpolated from the field calculations in (14)-(16) without the need for 3D FEA simulation.

C. Ferrite Losses

Losses in soft-magnetic materials are primarily broken down into hysteresis loss and conduction losses. However, the resistivity of ferrite materials is high, on the order of 5 Ωm, such that the eddy currents in the material are neglected. The Steinmetz equation

$$P_f = C_f f^\alpha B_p^\beta (T_{fer} - T_{fert} C_{10} + C_{11} + C_{12})$$  \hspace{1cm} (39)$$

is used as a curve-fit of hysteresis loss plots within certain ranges where $C_m$, $\alpha$, and $\beta$ are curve-fit coefficients, $P_f$ is the specific hysteresis loss of the material, and $B_p$ is the peak flux density in the material. The Steinmetz equation is used as the coil currents and flux densities are highly sinusoidal and the flux densities in the ferrite are well under the saturation flux density of the material. The losses are also a function of ferrite temperature, $T_{fer}$ with curve fit parameters $C_{10}$, $C_{11}$, and $C_{12}$.

The Steinmetz equation coefficients for the Ferroxcube 3C95 ferrite are summarized in Table II [25]. Many ferrite materials have lower losses when operating at temperatures well above room temperature. For Ferroxcube 3C95, an operating temperature of 25°C increases the loss by 16% relative to the nominal level given by the Steinmetz parameters at 85°C.

To evaluate (39), the spatial flux density in the ferrites must be calculated. In the Fourier Analysis Method, this is conveniently done by taking the integral of each field component to the distance of $\mu_{fer} t_{fer}$ in the $z$-direction to yield the average flux density in the ferrite of a thickness of $t_{fer}$ and relative permeability $\mu_{fer}$. For a coil in the $x$-$y$ plane, the average flux density in each direction are

$$B_{x,fer}(x, y) = \frac{N-1}{m, n = -N + 1} \frac{\mu_0 j k_x \psi(m, n) e^{j(k_x x + k_y y)}}{\gamma t_{fer} (1 - e^{-\gamma t_{fer} \mu r})}$$  \hspace{1cm} (40)$$
The switching losses of the inverter are minimized by operating slightly above resonance so that the inverter is soft-switching. Therefore, the switching losses of the inverter and dynamic effects on $R_{DS}$ are not included.

Likewise, on the secondary side, the secondary RMS current $I_2$ will flow through the forward voltages of the diodes $V_f$ and diode resistances $R_f$ of the rectifier. The reverse-recovery losses of the diodes are negligible as Schottky diodes are used.

$$P_{R_{DS}} = 2I_f^2R_{DS}. \quad (44)$$

The average diode forward current is calculated by

$$I_{2,avg} = \frac{2\sqrt{2}}{\pi}I_2. \quad (46)$$

E. Compensation Component Losses

Power capacitors such as the ones used in the WPT system are designed to have low series resistance and dielectric loss. The ratio of real power, or loss in this case, to reactive power is
Dissipation Factor

10
10
10
10
-4
-3
-2
-1
Capacitor (3S-50nF) Dissipation Factor

Fig. 14. The frequency-dependent loss tangent of polypropylene (PP) capacitors as in [26] compared to the measured DF of the HC1 capacitor bank with 3, 50 nF capacitors in series.

expressed as the tangent of the angle $\phi$ of the vector sum of the real power loss $P_c$ and reactive power $Q_c$ in the capacitor, or the dissipation factor (DF). The real power loss of resonant capacitors is comprised of both dielectric and conduction losses. For polypropylene-based capacitors, the dielectric component of the dissipation factor remains constant with frequency at around $10^{-4}$ to $2 \times 10^{-3}$. The conduction loss component of the loss tangent, however, scales with frequency such that the total loss tangent increases with frequency as illustrated in Fig. 14. Here, the measured dissipation factor of the capacitors is about half of the nominal curve given by the manufacturer at 85 kHz. The equivalent series resistance (ESR) of the resonant capacitors $R_C$ and the total power loss $P_c$ are

$$R_C = \frac{\tan(\phi(f))}{\omega C}$$

(47)

$$P_c = I_1^2 R_{C1} + I_2^2 R_{C2}$$

(48)

**F. Series-Series System Circuit Model**

With an equivalent AC load resistance on the secondary side $R_L$, the fundamental frequency model of a pair of series-tuned WPT coils is

$$\begin{bmatrix} V_1 \\ 0 \end{bmatrix} = \begin{bmatrix} Z_1 & -j\omega M \\ -j\omega M & Z_2 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}.$$ 

(49)

where the primary and secondary lumped series impedances are, respectively

$$Z_1 = 2R_{DS} + R_{C1} + R_{11} + j\omega L_1 + 1/(j\omega C_1)$$

(50)

$$Z_2 = 2R_f + R_{C2} + R_{22} + R_L + j\omega L_2 + 1/(j\omega C_2).$$

(51)

The AC input voltage $V_1$ and the equivalent AC load resistance $R_L$ are found by the first-harmonic approximation of a square wave as a function of the DC input voltage $V_{1,DC}$ and the DC output load resistance $R_{L,DC}$.

$$R_L = \frac{8}{\pi^2} R_{L,DC}$$

(52)

Likewise, the DC output voltage is

$$V_{2,DC} = \frac{\pi}{2\sqrt{2}} I_f R_L - 2V_f.$$  

(54)

In this linear circuit model, the resistances due to the ferrite losses are neglected. In the final loss calculation, the ferrite losses are calculated from the currents found by this linear model. By inverting this matrix, the input impedance of the system seen by the inverter is

$$Z_{in} = \frac{Z_1 Z_2 + (\omega M)^2}{Z_2}.$$  

(55)

To ensure zero-voltage switching of the inverter switches, the frequency and load of the WPT system are chosen so that the input impedance of the primary coil is inductive by choosing an operating frequency slightly greater than the resonant frequency of the tanks.

Using the output geometries derived from the optimization for a given $B_{str,lim}$ constraint, the loss and misalignment models detailed in this work are used to analyze efficiency for various conductor types, ferrite thicknesses, and number of turns. As an example, the two geometries in Fig. 16(a)-16(b) from the 70 cm x 50 cm optimization are selected for further analysis. These are the optimization outputs where $B_{str,lim} = 100 \mu T$ for the $\cos x \cos y$ and the $\sin x \cos y$ basis functions. These two solutions were selected because they have the two lowest currents of the four basis function outputs. By varying the number of turns, the impedance of the coils is tuned to meet input voltage requirements with series-series compensation in Fig. 16(c)-16(d). Using the loss models, the efficiencies at alignment are given in Fig. 16(e)-16(f). Here, more turns with higher voltages and lower currents are more efficient than having fewer turns with higher currents mostly due to the reduction of switching device conduction losses as a function of current. Here, it is assumed that the inverter and rectifier have constant device on-state resistances and forward voltages. Solutions with thicker gauges of wire and higher turns are limited by (7) so that the conductors fit in a single layer.

**IV. EXPERIMENTAL RESULTS**

A 6.6 kW prototype was built to evaluate one of the candidate coil geometries in Fig. 12(a) and validate the loss and field models with experimental measurements. The $\cos x \cos y$ geometry was chosen because it had the lowest loss metric of the four basis function sets at the moderate to large stray field constraints. The parameters used to model the system performance are given in Table II. The overall system consists of a set of two matched planar coils, two compensation capacitor banks, and power electronics consisting of a full-bridge inverter with four MOSFETs and gate driver boards, a control board with an FPGA, and a full-bridge diode rectifier as in Fig. 15(a).

One of the two identical prototype coils is shown in Fig. 15(b). Each coil is a sandwich structure comprised, from top to bottom, of a polycarbonate coil former with litz wire,
ferrite tiles of two thicknesses, hardboard spacers, and an aluminum sheet held together with nylon bolts and reinforced tape. The total length of the wire in each coil was measured to be 29.8 m, with 24.8 m in the coils themselves. The calculated length of the contours was 24.8 m. The additional lead length needed to connect to the inverter, capacitors, and rectifier is accounted for in the loss model. The airgap holder in Fig. 15(c) was made out of wood and nylon bolts to suspend one of the coils at different airgaps. The four threaded nylon rods allow for the quick adaptation of the airgap and support the upper coil from the bottom. Litz wire was used to symmetrically wind the coils.

The compensation capacitors are constructed of high-density resonant capacitors from Illinois Capacitor. Due to the need to reach high voltage levels, three capacitors were placed in series as seen in Fig. 15(d). An identical bank was constructed and connected to the secondary side to provide series-series tuning for the system. The area between the nuts on the underside of the capacitors required additional voltage insulation where the corners of nuts were facing each other. To resolve this issue, FR4 fins placed on top of Mylar tape were made to provide additional voltage insulation between the capacitor nuts. These are held in place by long pieces of FR4 aligning the busbars and nylon bolts. The assembly is enclosed in a PVC and plexiglass box.

A. Impedance Measurements

In Table III, the measured and calculated inductances of the system are compared. The measurements were obtained with an Agilent Technologies E4990A impedance analyzer. As measured, the self-inductance of the coils change as a function of airgap due to the presence of ferrite across the airgap as also seen in [27]. The FAM model also captures this effect. This change in self-inductance slightly changes the resonant frequency of the tank. To account for this effect, the operating frequency is chosen such that the input impedance is inductive for the largest airgap, when the coil self-inductances are lowest and the tank resonant frequency is the highest. The airgap used in this document is defined as the magnetic airgap of the system, i.e. the distance between the ferrites of each coil.

In Fig. 17, the measured series impedances of the primary and secondary tanks are shown when tuned to the 86.5 kHz operating points with three 50 nF capacitors in series. The parasitic parallel resonances of the tanks were seen to produce a high-frequency resonant current, as seen in the later test results. This high frequency current produces a negligible additional loss not captured in the fundamental frequency loss models.

B. Finite-Element Analysis Simulation of the Prototype

The physical prototype was implemented with litz conductors, a finite ferrite sheet close to the size of the coil extents, and an aluminum backing sheet. The aluminum sheet is used in the prototype for structural support and backside shielding and extends 1 cm on all sides beyond the outer ferrite extents. These non-ideal elements were included in an FEA simulation of the prototype to determine if they have any effect on the estimated impedance or fields relative to the FAM-derived values. The contours of the coil geometry were exported to finite-element analysis (FEA) software to accomplish this.

In Fig. 18, the FEA outputs of prototype fields are plotted with and without the aluminum sheet. As seen, the aluminum sheet provides some additional shielding for the fields far away from the coils at 80 cm, while the larger fields near the coils in the airgap are not significantly altered. The fields on the backside of the ferrite are also shielded by the aluminum sheet. This is similar to the result of [28] for unipolar coil shielding by aluminum sheets, though not as significantly as
the aluminum and ferrite are similarly sized. The FEA-derived inductance values are also compared with the FAM and measured values in Table III. Regarding mutual inductance, the FAM calculation method tends to slightly overestimate mutual inductance relative to FEA and the measured values at larger airgaps such as 210 mm and 250 mm. This is likely due to the homogenous boundary conditions used in FAM modeling, which approximates a large, continuous ferrite sheet. This error may become worse at larger airgaps as the coupled fields spread out. In contrast, the FEA simulations include finite ferrite dimensions and tend to underestimate mutual inductance at lower airgaps. As shown, the FAM and FEA derived inductance values both match the measured values with a maximum absolute error of 8.5% and 5.9%, respectively. FEA airgap fields also match the FAM-derived and measured values in Fig. 21.

### C. Efficiency Measurements and Waveforms

Tests were run to validate the efficiency and loss models over misalignment. Two power supplies were used in the tests: a Keysight N8935A for the high-power tests at 125 mm at 3.4 kW and 210 mm at 6.7 kW at alignment and a BK Precision PVS60085MR for the other low-power tests. A BK Precision 8612 electronic load was used alone or in parallel with wirewound resistors for these low-power tests. For the high-power tests, the wirewound resistor bank alone was used. Because of the voltage and current limitations of the electronic load and power supply for the low-power tests, the power levels are limited to 1.1 kW and below at 250 mm and at 125 mm and 210 mm in misaligned conditions. The system waveforms were obtained with a Tektronix MSO4104B-L. The DC current and voltage measurements used to derive DC-DC efficiency were obtained by multimeter and power supply current measurements. For these tests, the system was run in open-loop with constant load resistance.

The waveforms of the system at 210 mm and 6.7 kW are shown in Fig. 19(a). The frequency components of the waveforms are also given in Fig. 19(b)-19(c), showing the square-wave harmonics of the inverter and rectifier voltages and coil currents. The effect of the parasitic resonances of the tanks in Fig. 17 is seen in high-frequency components of the current waveforms. At the 6.7 kW operating point, the temperatures of the system elements were captured by thermal camera as shown in Fig. 19(d)-19(e). These temperatures were used to determine the temperatures used in the loss models in Table II. As seen, due to the 6.6 kW power level and high efficiency of the system, the wire temperature only rose 19°C above ambient with no active cooling. This temperature was used in the model to estimate the wire loss. The modeled loss breakdown of the system and efficiency measurements over varying misalignments are shown in Fig. 20. In these figures, the loss models matched the experimental measurements well over varying misalignments.

A summary of the measured efficiency values is given in Table IV. As shown, the efficiency of this work is similar to the previous works summarized in Table VII when compared by the airgap divided by the geometric mean length (GML) of the coil dimensions similar to the metric proposed in [29].

### D. Field Measurements

A sensor cubic similar to [5] was made to measure the fields of the system during operation for comparison with model values. The voltages induced in the sensor windings were measured with a Tektronix MDO3104 oscilloscope. Due to the high-frequency self-resonance of the sensor windings, an RC filter was used to damp the high-frequency voltages and is included in the calculation of the stray field. The parameters of the cubic and RC-filter are given in Table V. Field measurements at the center of a 210 mm airgap are shown in Fig. 21. Here, the FAM field models match the experiment with some accuracy, especially for the \( B_z \) values.

<table>
<thead>
<tr>
<th>Airgap</th>
<th>( L_{1, L_2} = M )</th>
<th>FEA</th>
<th>FAM</th>
<th>Measured</th>
<th>FEA Error (%)</th>
<th>FAM Error (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>125 mm</td>
<td>75.2 µH</td>
<td>224.9 µH</td>
<td>218.9 µH</td>
<td>217.2 µH</td>
<td>-1.3%, -0.5%</td>
<td>-5.9%</td>
</tr>
<tr>
<td>210 mm</td>
<td>29.8 µH</td>
<td>204.5 µH</td>
<td>205.4 µH</td>
<td>203.8 µH</td>
<td>-1.1%, -0.3%</td>
<td>-1.3%</td>
</tr>
<tr>
<td>250 mm</td>
<td>20.3 µH</td>
<td>201.7 µH</td>
<td>201.7 µH</td>
<td>200.3 µH</td>
<td>0.7%, 1.0%</td>
<td>0.7%, 1.0%</td>
</tr>
</tbody>
</table>

### Table IV

<table>
<thead>
<tr>
<th>Airgap</th>
<th>Misalignment (X, Y)</th>
<th>DC/DC Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 cm, 0 cm</td>
<td>97.6%</td>
<td></td>
</tr>
<tr>
<td>-10 cm, 0 cm</td>
<td>96.6%</td>
<td></td>
</tr>
<tr>
<td>0 cm, -7.5 cm</td>
<td>96.3%</td>
<td></td>
</tr>
<tr>
<td>45°</td>
<td>97.0%</td>
<td></td>
</tr>
</tbody>
</table>

### Table V

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnet Wire AWG</td>
<td>30 AWG</td>
</tr>
<tr>
<td>Number of Turns</td>
<td>( N_T = 45 ) turns</td>
</tr>
<tr>
<td>Turn Area</td>
<td>( A = 23.04 ) cm²</td>
</tr>
<tr>
<td>Field Sensitivity (dV/dB)</td>
<td>( 86.5 ) kHz, ( 56.4 ) mV/µT</td>
</tr>
<tr>
<td>RC Filter Values</td>
<td>( R = 1.6 ) kΩ, ( C = 300 ) µF</td>
</tr>
<tr>
<td>RC Corner Frequency</td>
<td>( 300 ) kHz</td>
</tr>
</tbody>
</table>
Fig. 16. Points from the 70 cm x 50 cm optimization shown for with $B_{str,max} = 100 \mu T$, $I = 85$ kHz, $z_{gap} = 210$ mm, and $P = 6.6$ kW and modeled loss using the parameters of Table II. (a) Potential of the $\cos x \cos y$ coil geometry. (b) Potential of the $\sin x \cos y$ coil geometry. Current and voltage for the (c) $\cos x \cos y$ coil geometry and the (d) $\sin x \cos y$ coil geometry. Efficiency at alignment of the (e) $\cos x \cos y$ coil geometry and the (f) $\sin x \cos y$ coil geometry from the loss models in this paper where $I_1 = I_2$. Misalignment efficiency of the (g) $\cos x \cos y$ coil geometry with $N_T = 25$ and the (h) $\sin x \cos y$ coil geometry with $N_T = 40$ with $I_1 = I_2$.

The $B_{x}$ and $B_{y}$ values were affected by the accuracy of positioning the sensor cubic height between the coils. These measurements were taken at a DC output power of 474 W to limit the voltages induced in the sensor windings from the large fields within the coil extents. The measurements are scaled by the square root of the ratio of 6.6 kW to 474 W.

The measured stray fields at 80 cm are summarized in Table VI. The field measurements at misalignment are on the side closest to the misaligned secondary. Compared to the literature on a normalized basis of stray field divided by the square root of output power, the measured stray fields at 80 cm were less than the works summarized in Table VII and well below the ICNIRP 27 $\mu T$ and 15 $\mu T$ pacemaker limits. In Table VII, the stray field metric $\mu T/kW^{0.5}$ comes from the direct relationship between the coil currents and field magnitude and the coil to coil power equation, $P = \omega M I_1 I_2$, where the power is proportional to the product of the coil currents. This relationship allows the stray fields of systems operating at different power levels to be compared and for the scaling of stray field measurements to higher or lower power levels as done in this work.

V. CONCLUSION

In this work, the Fourier Analysis Method (FAM) was used to predict the system efficiency, inductances, fields, and performance of complex, symmetric planar coil geometries derived from an optimization of Fourier basis function coefficients. The optimization outputs are geometries that meet the stray field and power level constraints with minimized loss factors. The FAM is then used to estimate the efficiency over various number of turns, conductor sizes, and ferrite thicknesses over varying airgaps and misalignments. As shown, the Fourier representation of the coil geometry also enables the convenient
calculation of coupling over misalignment, the external field on the conductors in proximity effect loss calculation, and the ferrite flux density for the calculation of ferrite loss for complex coil geometries.

A 6.6 kW WPT prototype with a shielded rectangular coil geometry was built from an optimization output to validate the method. The prototype was tested over a range of misalignments and airgaps to test the inductance, field, and loss models. As shown in Table VII, the prototype achieved similar efficiency with lower stray field when compared to the literature by a metric of the airgap over the geometric mean length (GML) of the coils and a metric of stray field over the square root of output power. Improving the efficiency and stray field is essential to achieve inductive charging for EVs that comply with stray field and EMI limits. In future work, the loss models will be imported into a time-domain simulation platform to assess thermal effects and cooling to design and validate a higher-power WPT system.

TABLE VI
SUMMARY OF SCALED RMS FIELD MEASUREMENTS (X, Y) AT 0.8 M, 86.5 kHz AND 6.6 kW

<table>
<thead>
<tr>
<th>Airgap</th>
<th>Misalignment (X,Y)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0 cm, 0 cm</td>
</tr>
<tr>
<td>125 mm</td>
<td>1.8 µT, 1.4 µT</td>
</tr>
<tr>
<td>210 mm</td>
<td>4.4 µT, 3.6 µT</td>
</tr>
<tr>
<td>250 mm</td>
<td>6.5 µT, 4.5 µT</td>
</tr>
</tbody>
</table>

ACKNOWLEDGMENT
The authors would like to thank Bob Martin and Caden Webb for their assistance with the material procurement, manufacture, and setup of the experimental system.

REFERENCES
### TABLE VII

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Power Level</th>
<th>Coil Dimension</th>
<th>Airgap (mm)</th>
<th>Airgap/ GML</th>
<th>Stray Field 80 cm - X, Y</th>
<th>Stray Field Metric kW/µT</th>
<th>DC/DC Efficiency</th>
<th>Freq. (kHz)</th>
<th>Coil Shape</th>
<th>Power Density</th>
</tr>
</thead>
<tbody>
<tr>
<td>[5]</td>
<td>50 kW</td>
<td>76 cm x 41 cm</td>
<td>160 (0.29)</td>
<td>N/A</td>
<td>N/A, 22.5 µT (N/A)</td>
<td>95.8%</td>
<td>85</td>
<td>Rect.</td>
<td>2.0 kW/kg</td>
<td></td>
</tr>
<tr>
<td>[5]</td>
<td>50 kW</td>
<td>76 cm x 41 cm</td>
<td>160 (0.29)</td>
<td>N/A</td>
<td>N/A, 12.5 µT (N/A)</td>
<td>95.3%</td>
<td>85</td>
<td>DD</td>
<td>2.0 kW/kg</td>
<td></td>
</tr>
<tr>
<td>[9], [30]</td>
<td>120 kW</td>
<td>88 cm x 67 cm</td>
<td>125 (0.16)</td>
<td>21.9 µT, 12.3 µT (11 kW)</td>
<td>97.1%</td>
<td>25</td>
<td>DD</td>
<td>2.28 kW/kg</td>
<td></td>
<td></td>
</tr>
<tr>
<td>[31]</td>
<td>50 kW</td>
<td>54 cm x 47 cm</td>
<td>150 (0.30)</td>
<td>N/A</td>
<td>N/A, 34.7 µT (N/A)</td>
<td>95.1%</td>
<td>85</td>
<td>30-DD</td>
<td>3.65 kW/kg</td>
<td></td>
</tr>
<tr>
<td>[32]</td>
<td>3 kW</td>
<td>60 cm O.D.</td>
<td>100 (0.17)</td>
<td>N/A</td>
<td>N/A</td>
<td>95.8%</td>
<td>94.2%</td>
<td>35</td>
<td>Circular</td>
<td>0.79 kW/kg</td>
</tr>
<tr>
<td>This Work</td>
<td>6.6 kW</td>
<td>71 cm x 54 cm</td>
<td>125 (0.20)</td>
<td>1.8 µT, 1.4 µT</td>
<td>(1.43, 1.84)</td>
<td>97.6%</td>
<td>95.5%</td>
<td>Shielded</td>
<td>86.5 Rect.</td>
<td>0.81 kW/kg</td>
</tr>
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**Fig. 21.** Scaled field measurements, FEA values, and FAM model outputs at 86 kHz, 210 mm, and 6.6 kW. (a) Fields on the X-axis and (b) the Y-axis.

**Fig. 20.** (a) Modeled loss breakdown of the system at alignment over varying output resistances. (b) Modeled efficiency of system over load resistance at alignment. (c) Model efficiency vs. measured efficiency at alignment, (d) at -10 cm in the x-direction, (e) at -7.5 cm in the y-direction, and (f) at 45° rotation.


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