Design and Optimization of Resonance-Based Efficient Wireless Power Delivery Systems for Biomedical Implants

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Abstract—Resonance-based wireless power delivery is an efficient technique to transfer power over a relatively long distance. This technique typically uses four coils as opposed to two coils used in conventional inductive links. In the four-coil system, the adverse effects of a low coupling coefficient between primary and secondary coils are compensated by using high-quality (Q) factor coils, and the efficiency of the system is improved. Unlike its two-coil counterpart, the efficiency profile of the power transfer is not a monotonically decreasing function of the operating distance and is less sensitive to changes in the distance between the primary and secondary coils. A four-coil energy transfer system can be optimized to provide maximum efficiency at a given operating distance. We have analyzed the four-coil energy transfer systems and outlined the effect of design parameters on power-transfer efficiency. Design steps to obtain the efficient power-transfer system are presented and a design example is provided. A proof-of-concept prototype system is implemented and confirms the validity of the proposed analysis and design techniques. In the prototype system, for a power-link frequency of 700 kHz and a coil distance range of 10 to 20 mm, using a 22-mm diameter implantable coil resonance-based system shows a power-transfer efficiency of more than 80% with an enhanced operating range compared to ~40% efficiency achieved by a conventional two-coil system.

Index Terms—Biomedical implants, coupling coefficient, inductive wireless power links, power transmission efficiency, resonance-based power delivery, telemetry, wireless power transfer.

I. INTRODUCTION

Implantable devices are becoming more and more popular in health and medical applications due to their ability to locally stimulate internal organs and/or monitor and communicate the internal vital signs to the outer world. The power requirement of biomedical implants depends on their specific application and typically ranges from a few microwatts [1]–[4] to a few tens of milliwatts [5]–[8]. Providing required power to implanted devices in a reliable manner is of paramount importance. Some implants use (rechargeable) batteries; however, their applications are limited due to the size and/or longevity of the batteries. Wireless power-transfer schemes are often used in implantable devices not only to avoid transcutaneous wiring, but also to either recharge or replace the device battery. Wireless power transfer is also used in other application domains where remote powering is required, for example, contactless battery charging [9] and radio-frequency identification (RFID) tags [10].

A popular technique for wireless power transfer, particularly in biomedical implants, is inductive coupling, which was first used to power an artificial heart [11], [12] and since then has commonly been used in implantable devices [1]–[8], [13], [14]. An inductively coupled power-transfer system consists of two coils that are generally referred to as primary and secondary coils. In these systems, power-transfer efficiency is a strong function of the quality factor (Q) of the coils as well as the coupling between the two coils. Hence, the efficiency depends on the size, structure, physical spacing, relative location, and the properties of the environment surrounding the coils. The coupling between the coils decreases sharply as the distance between the coils increases and causes the overall power-transfer efficiency to decrease monotonically. Inductive power transfer in a (co-centric) 2-coil system is extensively analyzed in the literature [5], [8], [10], [15].

The power-transfer efficiency, $\eta$, versus normalized distance $R$ (i.e., the ratio of the separation between the coils) ($d$) and the geometric mean of the primary and the secondary coil radii ($r_m = \sqrt{r_p r_s}$) is a commonly used performance metric for comparing different designs. Due to the low $Q$-factor (due to the source and load resistances) and low coupling of the coils in the two-coil system, two-coil-based power-transfer systems suffer from relatively low-power-transfer efficiency, typically, $\eta \leq 40\%$ for $d \geq r_m$ [5], [6], [16], and generally $\eta$ drops exponentially with distance ($\propto 1/d^3$) for $d \gg r_m$. In implantable devices, the size of the implanted coil is constrained by the implant site. Typically, an external coil can be made big enough to improve the power-transfer range. As coupling between coils depends on amount of magnetic flux linkage between primary and secondary coils, for a given operating range and small size of the implanted (secondary) coil, coupling reduces as difference between external (primary) coil...
radii and secondary coil radii increases. Increasing external coil dimension, however, increases the inductance of the coil and, hence, improves its $Q$-factor. Thus, an optimum dimension of an external coil exists for which the effect of coupling and $Q$-factor ($k \times Q_p$) is maximum.

For implants with a high-power requirement, a more efficient power-transfer mechanism (e.g., with $\eta$ of 60 to 90% for a distance of 20 mm or more) is desired. To provide high power at low efficiency, a strong alternating magnetic field is required which could result in excessive heat in the tissues and, in turn, violates the safety requirements of the federal regulations. For example, sonodynamic therapy (SDT), a drug delivery approach that uses ultrasonic cavitation to enhance the cytotoxicity of the chemotherapeutic drugs requires comparatively large power for stimulation and, hence, high-link efficiency is desired. Sonodynamic enhancement of doxorubicin cytotoxicity was investigated by using micro ultrasonic transducers (MUTs) in combination with a cancer drug in [17]. Using 60 s of toned burst ultrasound at 40 W/cm$^2$, the cytotoxicity of doxorubicin treatment increases from 27% to 91%. Using an ultrasound transducer of area 1 mm$^2$ (dimension 1 mm x 1 mm x 0.5 mm), the system requires 400 mW of power to stimulate the transducer. It should be noted that most of reported biomedical implants consume less than 100 mW of power [5]–[7].

Resonant-based power delivery is an alternative wireless power-transfer technique that typically uses four coils, namely, driver, primary, secondary, and load coils which will be discussed later. Coupled-mode theory [18] has been used to explain this phenomenon in [19]–[21]. Initially, this method was focused on high-power transfer and, hence, requires big coils. In [22], this technique is used for implantable and wearable devices, though a system with a large transmitter coil (radius of 176 mm) around the waist and several receivers is advocated. Similar independent work was performed in [23] using very big external coils (radius of 150 mm) and small load coils (radius 6.5 mm). This system uses inductive coupling between the driver and primary coil as well as between the secondary and load coil. We have also analyzed resonant-based power delivery for implantable devices and provided a simple electrical model for it [24]. In this paper, we present a more comprehensive circuit-based model for the system and analyze the effect of each design parameter. Furthermore, given the system requirements, we propose and analyze a step-by-step design procedure to optimize the system. A sample application of the technique for biomedical implants, in particular, implants that require relatively large power, such as [17] is provided, and design constraints are applied to find the optimum design to achieve maximum efficiency. Our focus is on the efficiency of the power-transfer link itself, and peripheral circuits, such as the power amplifier in the transmitter and rectifier and/or dc-dc converter in the receiver, are outside the scope of this paper.

This paper is organized as follows. Section II formulates the power-transfer efficiency of resonant-based systems. Section III describes the design steps. The resonance-based power-transfer system is described in Section IV. Section V presents the experimental setup. Results and analysis are provided in Section VI. Comparison with previous work is done in Section VII and concluding remarks are provided in Section VIII.

II. POWER EFFICIENCY IN RESONANCE-BASED SYSTEMS

This section presents the individual models for inductance, capacitance, and resistance of coils. Analytical models of each component are presented and are followed by a detailed analysis of the resonance power-transfer system. The models are based on a multilayer helical coil that uses Litz wire. However, the presented design steps are general. In case other types of coil are used, the respective inductance, capacitance, and resistance model of coil can be adjusted accordingly, and the remaining design steps and guidelines will be the same.

A. Inductor Model

Self inductance is a measure of magnetic flux through the area (cross section) enclosed by a current carrying coil. The self inductance of a coil with loop radius $a$ and wire radius $R$ (assuming $R/a \ll 1$) can be approximated as [25] and [26]

$$L(a, R) = \mu_0 a \left[ \ln \left( \frac{8a}{R} \right) - 2 \right].$$  \hspace{1cm} (1)

Mutual inductance is a measure of the extent of magnetic linkage between current-carrying coils. The mutual inductance of two parallel single-turn coils with a loop radius of $a$ and $b$ can be approximated by using (2) where $d$ and $\rho$ are the relative distance and lateral misalignment, respectively, between the two coils [25], [26]. The mutual inductance is a strong function of coil geometries and separation between them

$$M(a, b, \rho, d) = \pi \mu_0 \sqrt{ab} \left[ \int_0^\infty J_1 \left( \frac{x}{\sqrt{ab}} \right) J_1 \left( \frac{x}{\sqrt{ab}} \right) x \cdot J_0 \left( \frac{x}{\sqrt{ab}} \right) \exp \left( -\frac{d}{\sqrt{ab}} \right) dx \right].$$  \hspace{1cm} (2)

where $J_0$ and $J_1$ are the zeroth- and first-order Bessel functions.

For perfectly aligned loops ($\rho = 0$), the mutual inductance between the coils can be calculated as

$$M(a, b, \rho = 0, d) = \mu_0 \sqrt{ab} \left[ \left( \frac{2}{k} - k \right) K(k) - \frac{2}{k} E(k) \right].$$  \hspace{1cm} (3)

where

$$k = \left( \frac{4ab}{(a + b)^2 + d^2} \right)^{1/2}.$$  \hspace{1cm} (4)

and $K(k)$ and $E(k)$ are the complete elliptic integrals of the first and second kind, respectively [25]–[27].

Coils with different geometries have been used and modeled in the literature. The planar spiral coil is modeled in [13], [25], [26]. For a spiral coil with $N_s$ co-centric circular loops with different radii $a_i$ $(i = 1, 2, \ldots, N_s)$ and wire radius $R$, the self-inductance can be calculated as

$$L_1 = \sum_{i=1}^{N_s} L(a_i, R) + \sum_{i=1}^{N_s} \sum_{j=1}^{N_s} M(a_i, a_j, \rho = 0, d = 0)(1 - \delta_{ij}).$$  \hspace{1cm} (5)

where $\delta_{ij} = 1$ for $i = j$ and $\delta_{ij} = 0$, otherwise.

Printed spiral coils are implemented and optimized in [28]. It provides low self-inductance and constrains the maximum achievable $Q$-factor. To achieve large self-inductance, multi-layer helical coils can be used. For a helical coil with $N_s$ turns
per layer and $N_a$ coaxial layers, the total self-inductance can be modeled as

$$L_{\text{self}} = N_t \sum_{i=1}^{N_a} L(a_i, R) + \sum_{i=1}^{N_a} \sum_{k=1}^{N_t} M(a_{ik}, a_{ij}, \rho = 0, d = d[k - l]) \times (1 - \delta_{ij})(1 - \delta_{kl})$$

where $\delta_{ij}(or \delta_{kl}) = 1$ for $i = j(or k = l)$ and $\delta_{ij}(or \delta_{kl}) = 0$, otherwise. $d_l$ minimum distance between two consecutive turns.

**B. Parasitic Capacitance**

In general, inductors suffer from stray capacitance between turns. Stray capacitance causes self-resonance and limits the operating frequency of the inductor. Stray capacitance of a single-layer aird core inductors is modeled analytically in [29] and [30], and using numerical methods in [31]. For a multilayer solenoid with $N_a$ layers and $N_t$ turns per layer, stray capacitance is approximated as [32]

$$C_{\text{self}} = \frac{1}{N_a^2} \left[ C_b(N_t - 1)N_a + C_m \sum_{i=1}^{N_t} (2i - 1)^2(N_a - 1) \right]$$

where $N$ is the total turns, $C_b$ is parasitic capacitance between two nearby turns in the same layer, and $C_m$ is parasitic capacitance between different layers.

For a tightly wound coil, parasitic capacitance between two nearby turns is

$$C_b = \varepsilon_0 \varepsilon_r \int_0^{\pi/4} \frac{\pi D_k r_o}{\zeta + \varepsilon_r r_o (1 - \cos\theta)} d\theta$$

$$C_m = \varepsilon_0 \varepsilon_r \int_0^{\pi/4} \frac{\pi D_k r_o}{\zeta + \varepsilon_r r_o (1 - \cos\theta) + 0.5\varepsilon_r h} d\theta$$

where $D_k, r_o, \zeta, \varepsilon_r, h$ are the average diameter of coil, wire radius, thickness, and relative permittivity of strand insulation and separation between two layers, respectively [32].

**C. AC Resistance**

To achieve high quality factors, inductors with low effective series resistance (ESR) are required. At high frequencies, skin and proximity effect increases the ESR. To reduce the ac resistance, multistrand Litz wires are commonly used [1, 32]. Finite-difference time-domain (FDTD) techniques are used to model ac resistance numerically [33]. Analytical models of winding losses in the Litz wires are presented in [34] and [35]. Semiempirical formulation using finite-element analysis (FEA) is presented in [36]. The ac resistance of coils made of multistrand Litz wires, including skin and proximity effect, can be approximated as [32]

$$R_{\text{ac}} = R_c\left(1 + \frac{f_o^2}{f_h^2}\right)$$

where $f_h$ is the frequency at which power dissipation is twice the dc power dissipation and is given by

$$f_h = \frac{2\sqrt{2}}{\pi \varepsilon_r h_D \sqrt{\eta}}$$

where $R_{\text{dc}}, R_{\text{ss}}, N_a, \mu_0, \beta$ are the dc resistance of the coil, radius of each single strand, number of strands per bunch, permeability of free space, and the area efficiency of the bunch, respectively. $\eta$ is area efficiency of coil with width $b$ and thickness $t$ and can be calculated using [32, Fig. 1].

The dc resistance of the coil with $N_a$ coaxial layers and diameter $D_k$ can be calculated using

$$R_{\text{dc}} = \sum_{i=1}^{N_a} \frac{N_t}{A} \sum_{i=1}^{N_t} \pi D_k r_o$$

$$R_{\text{dc}} = \sum_{i=1}^{N_a} \rho_{\text{Litz}}(1.015)^{N_a} \frac{(1.025)^{N_c}}{AN_s}$$

**D. Coil Model**

Considering the effect of the stray capacitance and the ac resistance of an inductor, the total impedance of a coil can be written as [39]

$$Z_c = (j \omega L_{\text{self}} + R_{\text{ac}})\frac{1}{j \omega C_{\text{self}}}.$$  (14)

The coil can be modeled as an inductor with a self-inductance $L_{\text{eff}}$ and effective series resistance given by

$$R_{\text{esr}} = \frac{R_{\text{ac}}}{(1 - \omega^2 L_{\text{self}}/C_{\text{self}})^2}$$

$$L_{\text{eff}} = \frac{L_{\text{self}}}{(1 - \omega^2 L_{\text{self}}/C_{\text{self}})^2}.$$  (16)
As the operating frequency of coil approaches self-resonance frequency \( f_{\text{self}} \), ESR increases drastically. From (16), for a frequency higher than the \( f_{\text{self}} \), the coil behaves as a capacitor and, hence, it cannot be used as an inductor after its resonance frequency.

The \( Q \)-factor of an unloaded inductor can be written as

\[
Q_{\text{unloaded}} = \frac{\omega L_{\text{self}}}{\text{ESR}} = \frac{2\pi f L_{\text{self}}}{1 - \frac{f^2}{f_{\text{self}}^2}} \frac{1}{R_{\text{dc}} \left(1 + \frac{f^2}{f_{\text{self}}^2}\right)}, \tag{17}
\]

E. Power-Transfer Model

1) Two-Coil System: Conventionally, two coils are used in inductively coupled power-transfer systems and power is transferred from one coil to another coil. Power-transfer efficiency is a strong function of the \( Q \)-factor of primary coil \( (Q_p) \) and secondary coil \( (Q_s) \). Mutual coupling \( (k) \) between the coils is a function of alignment and distance between the coils. Efficiency of a two-coil-based power-transfer system is given by [6], [10], and [24]:

\[
\eta = \frac{k^2 Q_p Q_s}{1 + k^2 Q_p Q_s}. \tag{18}
\]

2) Four-Coil System: Couple-mode theory [18] has been originally used to describe resonance-based coupling [19], [20]. A simple circuit-based model for these systems is presented in [24]. The effect of the low \( Q \)-factor and the low coupling between the source and load coils can be compensated by using intermediate high-\( Q \)-factor coils. To realize efficient power transfer, the system consists of four coils referred to as driver, primary, secondary, and load coil (also denoted as coils 1 to 4). Fig. 3 shows the simplified schematic and electrical model of the four-coil system.

By applying circuit theory to this system, the relationship between current through each coil and the voltage applied to the driver coil can be captured in the following matrix form:

\[
\begin{bmatrix}
I_1 \\
I_2 \\
I_3 \\
I_4
\end{bmatrix} =
\begin{bmatrix}
Z_{11} & Z_{12} & Z_{13} & Z_{14} \\
Z_{21} & Z_{22} & Z_{23} & Z_{24} \\
Z_{31} & Z_{32} & Z_{33} & Z_{34} \\
Z_{41} & Z_{42} & Z_{43} & Z_{44}
\end{bmatrix}^{-1}
\begin{bmatrix}
E \\
0 \\
0 \\
0
\end{bmatrix} \tag{19}
\]

where

\[
Z_{mm} = R_n + j\omega L_n + \frac{1}{j\omega C_n}, \quad \text{for } m = n
\]

\[
= j\omega M_{nn}, \quad \text{for } m \neq n
\]

\( E \) is the amplitude of voltage source applied to the driver coil, and \( R_n, L_n, \) and \( C_n \) are the effective resistance, inductance, and capacitance of the coil \( n \). \( M_{mm} \) is the mutual inductance between coil \( m \) and \( n \).

\[
M_{mn} = k_{mn} \sqrt{L_m L_n},
\]

where \( k_{mn} \) is the coupling factor between coil \( m \) and coil \( n \).

Tuning all coils to the same resonance frequency and operating it at their resonance frequency, \( Z_{nm} = R_n \) (for \( m = n \) and \( n \in \{1, 2, 3, 4\} \)). For small driver and load coil inductance and relatively large distances between coils 1 and 4, coils 1 and 3, and coils 2 and 4, coupling coefficients \( k_{14}, k_{32} \) and \( k_{24} \) would be neglected. From (19), at resonance, the current in load coil can be calculated as

\[
I_4 = \frac{k_{12} k_{23} k_{34} Q_1 Q_2 Q_3 Q_4}{\sqrt{R_1 R_4 (1 + k_{12}^2 Q_1 Q_2)(1 + k_{23}^2 Q_3 Q_4)+k_{24}^2 Q_2 Q_3}} E \tag{20}
\]

where \( Q_n \) is the loaded quality factor of coil \( n \) at the frequency of operation. The power-transfer efficiency can be computed as

\[
\eta = \frac{(k_{12}^2 Q_1 Q_2)(k_{23}^2 Q_2 Q_3)(k_{34}^2 Q_3 Q_4)}{(1+k_{12}^2 Q_1 Q_2)(1+k_{23}^2 Q_2 Q_3)(1+k_{24}^2 Q_2 Q_3)+k_{24}^2 Q_2 Q_3}. \tag{21}
\]

F. Analysis of Four-Coil Power-Transfer System

To optimize the design to achieve high efficiency, the effects of different parameters on the power-transfer efficiency will be analyzed here.

1) High-\( Q \)-Requirement: From (21), low coupling between coils 2 and 3 can be compensated by a high-\( Q \)-factor of these coils. Efficiency is computed and plotted in Fig. 4 with a varying \( Q \)-factor of coil 2 (primary coil), \( Q_2 \), and the distance between coils 2 and 3. Note that respective \( k_{23} \), the coupling coefficient between primary and secondary coils corresponding to distance, decays exponentially with increasing distance between them.

Fig. 4 shows the effect of \( Q \)-factor on power-transfer efficiency ((21)). It can be deduced that for given coupling between
primary and secondary coils, as the \( Q \)-factor of the coils increases, power-transfer efficiency increases. To achieve high-power-transfer efficiency (e.g., 80% or beyond) for a high operating range, high-\( Q \)-factor coils are required.

Using moderate coupling between driver and primary coils \((k_{12})\) and secondary and load coils \((k_{34})\) along with high \( Q \)-factor primary \((Q_2)\) and secondary \((Q_3)\) coils, the following approximation can be derived:

\[
(1 + k_{12}^2 Q_1 Q_2)(1 + k_{34}^2 Q_3 Q_4) \approx (k_{12}^2 Q_1 Q_2)(k_{34}^2 Q_3 Q_4) \tag{22}
\]

\[
(k_{12}^2 Q_1 Q_2)(k_{34}^2 Q_3 Q_4) \gg k_{23}^2 Q_2 Q_3 \Rightarrow

[(1 + k_{12}^2 Q_1 Q_2)(1 + k_{34}^2 Q_3 Q_4) + k_{23}^2 Q_2 Q_3]

\approx (k_{12}^2 Q_1 Q_2)(k_{34}^2 Q_3 Q_4) \tag{23}
\]

\[
(1 + k_{23}^2 Q_2 Q_3) \gg k_{34}^2 Q_3 Q_4 \Rightarrow

(1 + k_{34}^2 Q_3 Q_4 + k_{23}^2 Q_2 Q_3) \approx (1 + k_{23}^2 Q_2 Q_3). \tag{24}
\]

Applying the aforementioned assumptions, the efficiency expression \((21)\) can be simplified to

\[
\eta \approx \frac{k_{23}^2 Q_2 Q_3}{1 + k_{23}^2 Q_2 Q_3}. \tag{25}
\]

This approximate model is similar to the model for two-coil systems. Since in the four-coil system, \(Q_2\) and \(Q_3\) are independent of the source and load resistances, high quality factors for primary and secondary coils can be achieved. Power-transfer efficiency increases monotonically as \(k_{23}^2 Q_2 Q_3\) increases.

Note that for \((25)\) to be a valid approximation, \(Q_4\) has to be within a certain range as will be explained.

For approximation in \((22)\) to be reasonable

\[
(1 + k_{34}^2 Q_3 Q_4) \gg 1 \Rightarrow (k_{34}^2 Q_3 Q_4) \gg 10 \Rightarrow Q_4 \geq \frac{10}{Q_3 k_{34}^2}. \tag{26}
\]

For \((24)\) to be a reasonable approximation, we should have

\[
(k_{23}^2 Q_2 Q_3) \gg 10 k_{34}^2 Q_3 Q_4 \Rightarrow Q_4 \leq \frac{k_{23}^2 Q_2}{10 k_{34}^2}. \tag{27}
\]

From \((26)\) and \((27)\), the range of \(Q_4\) can be shown as

\[
10 \frac{k_{23}^2 Q_2}{Q_3 k_{34}^2} \leq Q_4 \leq \frac{k_{23}^2 Q_2}{10 k_{34}^2}. \tag{28}
\]

2) Effect of \(Q_1\) and \(Q_4\) on Efficiency: The driver coil’s \( Q \)-factor is limited by the source series resistance, and the load coil’s \( Q \)-factor is limited by the load resistance as well as implant-size limitation. Due to high load resistance \((\sim 100 \Omega)\) and small size of the inner coil, \(Q_4\) is typically limited to a small value. However, the moderate \( Q \)-factor of 5 to 20 can be achieved for the driver coil. Fig. 5 plots the efficiency of a four-coil system as a function of \(Q_1\) and \(Q_4\).

Note that for the four-coil-based power-transfer system, efficiency does not vary much with respect to the driver coil’s \( Q \) factor and it has a maxima for a low load coil’s \( Q \)-factor (refer to Fig. 5). As mentioned before, it is common for the load coil to have a low \( Q \)-factor.

G. Design of High-\( Q \) Coils

To achieve a high \( Q \) factor for primary and secondary coils, the Litz wire, which provides low ac resistance, can be used. Based on operating frequency, the gauge of the single strand in the Litz wire is chosen. The number of strands in one bunch is used as a design parameter.

1) Wire Property: Litz wires are commonly used to reduce the ac resistance of wire and, hence, improve \(Q\)-factor of coils. To define the link operating frequency, the following considerations are taken into account. First, for the frequency range of the 100-kHz to 4-MHz band, no biological effects have been reported, in contrast to the extreme-low-frequency band and the microwave band [40]. Second, tissues have lower absorption for low-frequency RF signals compared to high-frequency signals. Third, due to the small size of the implanted coil, it has small inductance and small parasitic capacitance. For lower frequency of operation, the coil needs to be tuned by using a high value external resonating capacitor. Furthermore, by using large tuning capacitance, parasitic capacitances due to wire winding
and variation in capacitance due to tissues would be negligible. Hence, the resonating frequency of implanted coil will not be affected by proximity of tissue. Fourth, if the operating frequency is close to the self-resonant frequency of coil, the ESR of the coil increases drastically [(15)]; thus, for a moderate self-resonant-frequency coil, by operating at a lower frequency, high quality factor can be achieved [(17)]. Based on the aforementioned points, Litz wire with single-strand wire gauge of American Wire Gauge (AWG) AWG44 is chosen. AWG44 provides \( R_{ac}/R_{dc} = 1 \) for a frequency range of 350–850 kHz [38]. For applications where the operating frequency is fixed, respective wire gauge can be chosen to keep \( R_{ac} \) close to \( R_{dc} \). \( R_{dc} \) reduces as \( N_s \) increases. Due to the proximity effect, \( f_h \) reduces as \( N_s \) increases (\( f_h \propto 1/\sqrt{N_s} \)) and causes high ac resistance [(11)]. The diameter of the Litz wire increases as the number of enclosed strands is increased. For a given thickness of coils, the optimum number of strands that improves \( Q_2 \) and \( Q_3 \) can be calculated. Fig. 6 shows the product of \( Q_2 \) and \( Q_3 \) for a given dimension of primary and secondary coils with a varying number of strands. Fig. 7 shows the operating frequency for which this product is maximum. Based on Figs. 6 and 7, 40 strands Litz wire of strand gauge AWG44 is chosen for this work. A similar calculation can be performed based on design constraints to calculate the number of strands and single-strand gauge.

2) Number of Turns: To analyze the effect of the number of turns per layer on the \( Q \)-factor of the coils, using (6), (7), (10), and (17), one can derive an expression for the \( Q \)-factor and it can be maximized with respect to the number of turns per layer \( (N_t) \) for a given number of layers (for \( N_a \gg 1 \)) as follows:

\[
\frac{dQ}{dN_t} = 0 \Rightarrow 1 - 4 \frac{\omega^2}{\omega_{self}^2} - 3 \frac{\omega^2}{\omega_{self}^2} \frac{\omega^2}{\omega_h^2} = 0
\]  

(29)

where \( \omega_{self} = 1/\sqrt{L_{self}C_{self}}, \omega = 2\pi f, \) and \( \omega_h = 2\pi f_h. L_{self}, C_{self}, \) and \( \omega_h \) are a function of \( N_t \) (6), (7), (11).

Typically, the operating frequency is kept limited to \( 2 \times f_h \) as ac resistance increases exponentially with operating frequency

\[\begin{align*}
\text{Approximate operating Frequency (x10^4) Hz} & \\
0.5 & \\
1 & \\
2 & \\
3 & \\
4 & \\
5 & \\
6 & \\
7 & \\
8 & \\
9 & \\
10 & \\
11 & \\
12 & \\
13 & \\
14 & \\
15 & \\
16 & \\
17 & \\
18 & \\
19 & \\
20 & \\
\end{align*}\]

Fig. 7. Optimum frequency of operation.

As number of turns increases, \( I_{self} \) increases and the self-resonating frequency \( (f_{self}) \) decreases. With \( N_t \) for which \( \omega \leq \omega_{self} \), the \( Q \)-factor increases monotonically with an increment of \( N_t \) (\( dQ/dN_t \geq 0 \), (30)). \( N_{t(\text{opti})} \) can be defined as \( N_t \) for which \( \omega = (1/4)\omega_{self} \).

As an example, Fig. 8 shows the graph of the \( Q \)-factor for a coil with a fixed number of layers and fixed width. \( h \) is the distance between two layers and OD is the diameter of the wire. As can be seen from the figure, the \( Q \)-factor of the coil increases monotonically when the number of turns is less than 10.

3) Optimum Operating Distance and Effect of \( Q_1 \): In a typical case, the relative position and dimension of the driver coil and primary coil (and of secondary and load coil) is fixed and, hence, in normal operating mode \( Q_2, Q_3, Q_4, k_{12}, \) and \( k_{34} \) are fixed. For a given operating distance (respectively, \( k_{23} \)), only \( Q_1 \)
can be varied by using changing source resistance and, hence, the effect of $Q_1$ on power-transfer efficiency is shown here. The power-transfer efficiency is a strong function of coupling between coils 2 and 3 ($k_{23}$). By maximizing efficiency ($\eta$) with respect to the coupling coefficient, the optimum distance of operation can be achieved. From the efficiency equation [refer to (21)], we have

$$\frac{\partial \eta}{\partial k_{23}} = 0 \Rightarrow k_{23,(\text{opt})} = \left(\frac{\sqrt{\frac{k_{12}^2 Q_1 Q_2}{Q_2 Q_3}}}{1 + \frac{k_{34}^2 Q_3}{Q_2}}\right)^{1/2} \tag{31}$$

$$\frac{\partial^2 \eta}{\partial k_{23}^2} < 0, \quad \text{for} \quad k_{23} = k_{23,(\text{opt})}. \tag{32}$$

An expression for efficiency at $k_{23,(\text{opt})}$ is given by

$$\eta = \frac{(k_{34})^2}{(k_{23})^2} \frac{(k_{12} \sqrt{Q_1 Q_2})^3}{(1 + k_{12} \sqrt{Q_1 Q_2})^2 Q_2}. \tag{33}$$

To achieve maximum efficiency at any given distance (which is equivalent to a given $k_{23}$), $Q_1$ can be varied by controlling the source resistance.

Fig. 9 shows that for given system parameters for each value of $Q_1$, a corresponding distance exists for which efficiency is maximum. Equation (31) shows the dependence of the optimum value of $k_{23}$ on design parameter $Q_1$. So by changing the $Q$-factor of the driver coil ($Q_1$), the optimum operating distance changes.

4) Sensitivity of Efficiency ($\eta$) to Source Series Resistance:

To compare the effect of source resistance in two-coil-based systems and four-coil-based systems, we derive an expression for the slope of efficiency with respect to source series resistance ($R_1$)

$$\frac{\partial \eta}{\partial R_1} = \frac{\partial \eta}{\partial Q_1} \frac{\partial Q_1}{\partial R_1}. \tag{34}$$

For a two-coil system, the rate of change of efficiency is from (18) and (34), and $Q_1 = \omega L_1 / R_1$

$$\frac{\partial \eta}{\partial R_1} = -\frac{\omega L_1}{R_1^2} \frac{k^2 Q_2}{(1 + k^2 Q_1 Q_2)^2} \quad \text{for} \quad Q_1 \gg 1, \tag{35}$$

For a four-coil system, (36) gives an (approximate) analytical expression for the change in efficiency with respect to $R_1$

$$\frac{\partial \eta}{\partial R_1} \approx -\frac{k_{23}^2}{\omega L_1 k_{12}^2 k_{34} Q_4}. \tag{36}$$

For fixed design parameters, (35) and (36) show that efficiency decreases linearly as source resistance increases. To validate the accuracy of (35), efficiency is calculated by using (18) for different values of $R_1$ and plotted in Fig. 10. A linear regression model is used to find the slope of changes in efficiency with respect to $R_1$. From the linear regression model of efficiency for varying source resistance, slope $= -0.00768$ and from (35) slope $= -0.01069$.

Similarly to validate the accuracy of (36), with the same system parameters, efficiency is calculated using (21) with varying $R_1$ and plotted in Fig. 10. From the linear regression model of efficiency for varying source resistance, slope $= -0.00130$ and from (36), slope $= -0.001431$ (approximated model).

This example provides the validity of (35) and (36). For a given example, by comparing the slope of efficiency for a two-coil-based and four-coil-based system, the efficiency of the latter decreases $\approx 10$ times slower compared to that of the former. The slope for a two-coil-based system is inversely proportional to $k_{23}^2$ (equivalent to $k_{23}$) and, hence, have a high value compared to a four-coil-based system in which the slope is proportional to $k_{23}^2$ (note that $k_{23} \ll 1$). In the typical case, an increase in value of $R_1$ has a more severe effect in two-coil-based systems compared to four-coil-based systems.
5) Sensitivity of $Q$ to Frequency: The $Q$ factor of an inductor is a function of frequency

$$Q(f) = \frac{2\pi f L}{R_{dc}} \frac{1 - \left(\frac{f}{f_{res}}\right)^2}{\left(1 + \left(\frac{f}{f_h}\right)^2\right)},$$

(37)

At low frequencies, $Q(f)$ increases with frequency and for $f \geq f_h$, due to the dominance of proximity effect on ac resistance [32], the $Q$-factor decreases. To utilize the coil for a different operating frequency, it is desirable to have a $Q$-factor that is not too sensitive to frequency. The bandwidth of $Q$ is mainly defined by $f_h$. To reduce sensitivity to the operating frequency, $f_h$ should be kept sufficiently high so that the ac resistance stays small (10).

Fig. 11 shows the $Q$-factor variation with respect to operating frequency for coils with a different number of layers. As the number of layer increases, $f_h$ decreases (11) and the self-resonating frequency ($f_{self}$) decreases due to the increase in stray capacitance and inductance.

H. Effect of Operating Frequency Variation

A four-coil-based system provides better immunity to operating frequency variation compared to its corresponding two-coil-based system. For a four-coil system, by itself, the driver coil has a low $Q$-factor due to low inductance and, hence, has a high bandwidth of operation. The driver coil and high-$Q$ primary coil are closely coupled and mutual inductance seen by the driver coil due to the primary coil is high. This increases the influence of frequency variation on the driver coil. When the distance between the secondary coil and the primary coil decreases, the current in the secondary coil opposes the current in the primary coil which, in turn, reduces the effect of the primary coil on the driver coil (reduced $I_2$). Hence, the closer the secondary coil is to the primary coil, the lower the effective $Q$ factor of the driver coil. For the two-coil-based system, with comparable primary coil size and same source resistance as of the four-coil system, the $Q$-factor of the primary coil is higher than that of the driver coil in the four-coil system. Therefore, the two-coil system is a narrower band and is thus more sensitive to operating frequency.

Fig. 12(a) and (b) shows that for a four-coil-based power-transfer system as the coil separation between primary and secondary coils decreases, the frequency range over which the four-coil system has a higher efficiency is wider compared to the corresponding two-coil-based system. Note that for comparison in these two figures, practically the same-size external and implantable coils are used for two-coil- and four-coil-based systems.

I. Series Versus Parallel Connection of Load Resistance

To improve the $Q$-factor of the load coil, the load resistance can be either attached in series or parallel to a resonating capacitor ($C_R$, refer to Fig. 13). For parallel connection of the load resistance as shown in Fig. 13(a), we have

$$R_{eff} = \frac{R_L}{\left(\frac{\omega}{L}\right)^2}$$

(38)
J. Tissue Effects

Since the implantable coils are surrounded by tissue, in this subsection, an overview of the effects of tissue on the performance of the four-coil-based power-transfer system is provided. Let us first consider the effects of tissue on the implanted coil parameters (i.e., self inductance \( L_{\text{self}} \), parasitic capacitance \( C_{\text{par}} \), ac resistance \( R_{\text{ac}} \), self-resonance frequency \( f_{\text{self}} \), and the \( Q \)-factor. Since the tissue is typically free of magnetic materials, the permeability of tissue is close to permeability of free space [41]. The inductance of any coil depends on coil structure, dimensions, and the permeability of its surroundings and, hence, in the proximity of the tissue, the implantable coil inductance will be very close to that of the coil in the free space. However, the tissue has a high dielectric constant [1], [42] and, hence, the parasitic capacitance of the implantable coil increases compared to the case when the coil is not implanted. The ac resistance of the coil depends on the surrounding permeability and is independent of the dielectric property of tissue (10) and, therefore, when surrounded by the tissue, the ac resistance of the implanted coil will remain practically the same. Due to the increase in the parasitic capacitance of the implanted coil (because of tissue effects), the self-resonance frequency \( f_{\text{self}} \) of the implantable coil decreases. Thus, due to tissue effects, the \( Q \)-factor of the implanted coil will also decrease [refer to (17)]. However, by keeping the operating frequency of the system sufficiently lower than \( f_{\text{self}} \) of the implanted coil, the adverse effects of tissue on the implanted coil can be kept low (17). Also, considering that the inductance of the implanted coil is typically small (due to its small size), when the operating frequency is properly chosen (i.e., it is sufficiently low) (lower than \( \sim 4 \) MHz), the tuning capacitance \( C_R \) in Fig. 13, used in conjunction with the implanted coil, is much larger than the parasitic capacitance of the implanted coil and, hence, the change in the parasitic capacitance (due to tissue) has minimal effects on the operating frequency of the system. Note that as presented in [42], the effects of tissue on the behavior of the implanted coils, in particular, on their parasitic capacitance, can be included in the design cycle.

It should also be noted that in the proximity of tissue, part of the magnetic field is absorbed by tissue due to the generation of eddy currents in the tissue. However, it is known that the tissue has a lower absorption for low-frequency RF signals compared to high-frequency signals [40] and, thus, another advantage of using a low operating frequency (below \( \sim 4 \) MHz) for the power link is that the degradation in link efficiency due to tissue effects can be kept low. In this paper, the operating frequency of the power-transfer link is chosen to be 700 kHz.

III. DESIGN STEPS

In this section, design steps for resonance-based (4-coil) power delivery systems are presented. These steps are presented in the context of a design example that requires relatively high power to be transferred to the implantable device.

A. Design Constraints

The first step is to identify the design constraints. The specific application requirements constrain the design parameters (particularly in terms of size and source and load resistances) of
the implantable device. For example, Table I shows the design constraint dictated by the specific application of [17].

B. Initial Values and Range of Parameters

The initial design parameters are chosen as follows.

1) External Coil Radius: In a single-turn circular coil with radius $r$, the magnetic-field strength $H$ at distance $x$ along the axis can be written as [28]

$$H(x, r) = \frac{I r^2}{2 \sqrt{(r^2 + x^2)^3}}$$

(43)

For $r = x\sqrt{2}$, $H$ will be maximized and, therefore, a good choice of diameter for an external coil is $D_{\text{out}T} = d_2\sqrt{2}$. For $d = 20\text{ mm}$, $D_{\text{out}T}$ can be chosen as 60 mm.

2) Wire Property: Single strand copper wire has high ac resistance for high-frequency operation. Using multistrand Litz wire and using them in their operating frequency range, ac resistance can be kept very close to their dc resistance. For implantable coils, Litz wire is commonly used [11, 32]. Based on analysis done in Section II-G for properties of the Litz wire with a different number of strands (Fig. 6), to improve the $Q$-factor, one can choose a specific Litz wire. In our sample application, a 40-strand Litz wire with gauge 44 is chosen. Table II shows the properties of this particular Litz wire [38].

3) $Q$-Factor: As for the $Q$-factor of the four coils that are used in the system, we have:

$Q_1$: As shown in Fig. 5, for high values of $Q_2$ and $Q_3$, the effect of $Q_1$ on efficiency is not considerable. For constrained system dimensions, driver coil is made of a single layer ($N_1 = 1$) to keep enough room for primary coil winding. It is kept in the outermost layer to obtain considerable inductance compared to the case when driver coil is wrapped in the innermost layer of the primary coil. The number of turns are maximum permissible given the design constraints so that the overall size of the external coils in the four-coil system (i.e., a combination of the driver coil and the primary coil) is almost the same as the size of the external coil in the conventional two-coil system (i.e., primary coil). Since the power amplifier has an output impedance on the order of 5 to 6 $\Omega$, a source resistance of 5.6 $\Omega$ is chosen as sense resistance to mimic the source impedance.

$Q_2$: For a small number of turns per layer, increasing the number of turns improves the $Q$-factor. The number of turns in coil 2 is constrained by design requirements. In our example, $h_T$ of 5.5 $\text{mm}$ implies $N_1 = 11$ with $OD = 0.48\text{ mm}$. Fig. 14 shows the variation of $Q_2$ with the changing number of layers and operating frequency. Increasing the primary coil size increases the parasitic capacitance between its turns. To obtain high $Q$-factor at low frequency, the inductance of primary coil should be a large value. It results in a significant effect of parasitic capacitance on its self-resonating frequency. To reduce the parasitic capacitance between coil turns, low dielectric insulating material is inserted between layers. As a rule of thumb, the thickness of the dielectric layer should be varied until the self-resonating frequency of the secondary coil, $f_r$, is almost be the same as sense resistance to almost be the same as

Fig. 14. $Q_2$ versus the number of layers and operating frequency.

$Q_3$: With the constraint of $h_R = 2.5\text{ mm}$, five turns per layer can be accommodated and Fig. 15 shows the variation of $Q_3$ for different number of layers and operating frequencies. In this design, the self-resonating frequency of the secondary coil, without dielectric between layers is high enough compared to the operating frequency range so no dielectric layer is used.

$Q_4$: Given the load resistance (e.g., 100 $\Omega$) and single turn per layer, Fig. 16 shows the variation of $Q_4$ for a different number of layers and operating frequencies. To approximate the four-coil power-transfer model with a two-coil equivalent, (26) and (27) can be used to find the desired $Q_4$ from the graph. Additional consideration is to keep the overall size of the implantable coils in the four-coil system (i.e., combination of the driver coil and the primary coil) to almost be the same as

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Design Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmitter Outer diameter</td>
<td>$D_{\text{out}T}$</td>
<td>$\leq 80\text{mm}$</td>
</tr>
<tr>
<td>Transmitter Coil Thickness</td>
<td>$b_T$</td>
<td>5.5 mm</td>
</tr>
<tr>
<td>Implanted outer diameter</td>
<td>$D_{\text{out}T}$</td>
<td>22 mm</td>
</tr>
<tr>
<td>Implanted Coil Thickness</td>
<td>$b_H$</td>
<td>2.5 mm</td>
</tr>
<tr>
<td>Minimum Coil inner diameter</td>
<td>$D_{\text{in}T}$</td>
<td>8 mm</td>
</tr>
<tr>
<td>Coil relative distance</td>
<td>$d$</td>
<td>20 mm</td>
</tr>
<tr>
<td>Source resistance</td>
<td>$R_S$</td>
<td>5.6 $\Omega$</td>
</tr>
<tr>
<td>Load resistance</td>
<td>$R_L$</td>
<td>100 $\Omega$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Strand gauge</td>
<td>-</td>
<td>AWG 44</td>
</tr>
<tr>
<td>Number of Strands</td>
<td>$N_s$</td>
<td>40</td>
</tr>
<tr>
<td>Insulation thickness</td>
<td>$\zeta$</td>
<td>3 $\mu$m</td>
</tr>
<tr>
<td>Strand radius</td>
<td>$r_s$</td>
<td>25 $\mu$m</td>
</tr>
<tr>
<td>Operating Frequency</td>
<td>-</td>
<td>350-850 KHz</td>
</tr>
<tr>
<td>Outer Diameter</td>
<td>$OD$</td>
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</tr>
<tr>
<td>Max. DC resistance</td>
<td>$R_{\text{DC}}$</td>
<td>2.873/feet</td>
</tr>
<tr>
<td>Unit length DC resistance</td>
<td>$R_{\text{DC}}^{\text{unit}}$</td>
<td>0.0768/feet</td>
</tr>
<tr>
<td>Filling factor</td>
<td>$\beta$</td>
<td>0.4784</td>
</tr>
</tbody>
</table>
C. Optimizing Design Parameters

To implement coils, the driver and primary coils are made co-centric (and coaxial), and the number of turns per layer in the driver \((N_t(1))\) and primary \((N_R(2))\) coil are equal and kept close to \(N_t(1)\). Due to the small size of the secondary coil, the self-resonating frequency is high compared to the operating frequency so the number of turns are mainly limited by design constraints. The number of turns per layer in secondary coil \((N_t(3))\) and load coil \((N_t(4))\) are also chosen as equal.

Fig. 17 shows the design steps to obtain system parameters \((N_t(n))\) and \((N_R(n))\) for \(n \in \{1, 2, 3, 4\}\) to achieve the optimum efficiency at a given distance. The range for the number of layers in coil and turns per layer depends on the design constraints and may vary per the application requirement.

IV. RESONANCE-BASED POWER TRANSFER

Previous sections provide the models of different design parameters of the four-coil-based power-transfer system. Design steps help select the design parameters based on design constraints (Table I). Based on the design constraints, the power-transfer efficiency can be maximized for a targeted operating distance (e.g., \(d = 20\) mm) and operating frequency can be calculated for which the design provides the maximum efficiency. Table III shows the mechanical specifications of the optimized design by following the design flowchart (Fig. 17).

Based on simulation by following the design flowchart (Fig. 17), the operating frequency of 700 kHz is chosen. Depending on the application and the design constraints, the optimum design parameters can be calculated based on the design flowchart in Fig. 17. Table IV shows the simulated electric parameters for the optimized coil dimensions to obtain high-power-transfer efficiency. A source series resistance of \(R_s(= 5,6\Omega)\) is used to emulate the nominal output impedance of the power amplifier.


for driver coil. A load resistance of $R_L (= 100 \Omega)$ is used which is a realistic load for a targeted application.

Coupling between coils 1 and 2 ($k_{12}$) can be calculated using the coil dimensions and parameters. From simulation $k_{12} = 0.6335$, $k_{34} = 0.601$ and $k_{23} = 0.058$ for $d = 20$ mm.

For coaxial primary and secondary coil with physical dimensions per Table III, Fig. 18 shows the coupling coefficient $k_{23}$ along with the fitted curve based on regression model (95% confidence). The curve is modeled as

$$k_{23} = 148.2 \left( \frac{1}{\sqrt{d^2 + r_m^2}} \right)^{1.2} - 0.0002857$$ (44)

where $r_m = \sqrt{r_p r_s} = \sqrt{32 \times 11}$ and $d$ are the edge-to-edge minimum distance between the coils. $r_p$ and $r_s$ are the radius of the primary (coil 2) and the secondary coil (coil 3), respectively. For other applications based on design constraints and dimension of coils (calculated based on the flowchart), coefficients of the coupling model will change.

**V. EXPERIMENTAL SETUP**

To demonstrate the validity of the presented modeling techniques and the design flow, a prototype four-coil wireless power-transfer system is designed and implemented. Table III shows the optimized dimension of the coils based on the design constraints. Multi-strand Litz wire ($N_s = 40$) of strand AWG 44 is used to implement the coils. The HP4194A impedance/gain-phase analyzer is used to measure the electrical parameters of the coils which are reported in Table V. The $Q$-factor of each coil is limited due to the high ac resistance for the operating frequency above $f_b$ (10) and the increase in effective resistance due to low self-resonant frequency (SRF) [32]. $k_{12}$ and $k_{34}$ are measured to be 0.56 and 0.59, respectively, which are close to simulated coupling factors. $k_{12}$ and $k_{34}$ do not change during the operation of the system (since coils 1 and 2 and coils 3 and 4 do not move with respect to each other). A 50-$\Omega$ sinusoidal source is used to generate a signal with a 10.2-V amplitude at 700 kHz. A resistance of 5.6 $\Omega$ is used in series with the driver coil to measure the current of this coil. Since most energy is dissipated at an internal source resistance of 50 $\Omega$, a supply with a lower impedance should be used to improve the efficiency of the system. In this setup, the efficiency is calculated from the output terminal of the signal source and the effect of the realistic power-amplifier source resistance is taken into account by using 5.6-$\Omega$ sense resistance. When an input voltage source with a constant amplitude is used as the input source, the input power supplied to the system depends on the load impedance seen by the source. Changing the distance between the primary and secondary coils will result in the variation of the impedance seen by the source and, thus, the input power to the system will change accordingly. The plots of the measured and simulated input and output power of the four-coil-based power-transfer system are presented later (Figs. 24 and 25). For the conventional two-coil-based system when driven by a fixed-amplitude input voltage source, the input and output powers show similar trends as those of the four-coil system. The simulated and measured values of the power-transfer efficiency of the conventional two-coil-based system are provided in Fig. 22.

For this system, primary coil is wound over plastic tube with a height of 5.5 mm with side walls. After making the primary coil, the driver coil is wrapped over the primary coil to obtain high coupling between these coils ($k_{12}$). Similarly, secondary coil is made on a smaller tube, and the load coil is wrapped over secondary coil (Fig. 19). Fig. 21 shows the structure and cross section of the primary (or secondary) coil, position of the driver (or load) coil, and the order of the turns in each coil. The structure of coil Fig. 20 shows the relative dimensions of the coils. Since plastic does not affect the magnetic field, the effect of tube on the operation of the system can be neglected. In our experiment, all coils are centered around the same axis and the

<table>
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<th>Table IV</th>
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<tr>
<td><strong>COILS’ ELECTRICAL SPECIFICATION</strong></td>
</tr>
<tr>
<td>Coil (Number)</td>
</tr>
<tr>
<td>---</td>
</tr>
<tr>
<td>1</td>
</tr>
<tr>
<td>2</td>
</tr>
<tr>
<td>3</td>
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<td>4</td>
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Fig. 19. Power-transfer system.
horizontal distance between the primary and the secondary coils varies to change their coupling factor ($k_{23}$).

VI. RESULTS AND ANALYSIS

The distance between the primary and secondary coils varies from 6 mm to 52 mm in steps of 2 mm. Measured efficiency of the power-transfer system (Fig. 19) for four-coil-based and the conventional two-coil-based power-transfer systems is illustrated in Fig. 22 and shows that even with a relatively large distance between the coils of $d \geq 20$ mm (equivalent $r_m = 1.07$), high power-transfer efficiency is achieved. The measured results match very well with the SPICE simulations and the analytical equations derived to calculate power-transfer efficiency. The four-coil-based power-transfer system provides stable power-transfer efficiency over a long operating range. When the primary coil and the secondary coils are close (≈5-10 mm), experimental results are slightly different from the approximated analytical model derived for the power-transfer model (21). This slight discrepancy is due to the assumption of low $k_{34}$, $k_{24}$, and $k_{13}$ to derive (21). When coils are close, these parameters cannot be neglected. For SPICE simulations, the effect of $k_{34}$, $k_{24}$, and $k_{13}$ is taken into account and, hence, closer matches are obtained with respect to measured data.

In the traditional two-coil inductively coupled power-transfer system [10], the decrease in $\eta$ with respect to distance is more pronounced (see Fig. 22). This is because the coupling coefficient of the two coils drops rapidly with the distance ($\alpha$) and the coils have a small $Q$-factor (in this case, the loaded $Q$-factor for the external coil is $\approx 236$ and for the implanted coil, which is loaded by a 100 $\Omega$ load is 1.28). Fig. 23 shows the effect of source resistance on power-transfer efficiency. For a low value of series resistance, a high efficiency can be obtained. A $Q$-factor of the driver coil changes the optimum efficiency point and shifts with the distance as shown by simulation (Fig. 9). For
between primary and secondary coils. Using a constant-amplitude voltage source (10.2 V), the measured and simulated input power of the four-coil-based prototype power-transfer system is plotted in Fig. 24. The output power at 100-Ω load resistance of the system is measured and plotted in Fig. 25. As can be seen from these figures, the measured and simulated values closely match.

VII. COMPARISON WITH PREVIOUS WORK

The design based on the proposed technique is compared with previously reported power-transfer methods applied for implanted devices. Table VI summarizes the results. To make a fair comparison with different designs, efficiency at a normalized distance is presented, where \( r_p \) and \( r_s \) are the radius of primary and secondary coils. \( r_m \) is the geometric mean of \( r_p \) and \( r_s \).

VIII. CONCLUDING REMARKS

In this paper, the design and optimization steps for resonance-based four-coil wireless power delivery systems are described. The focus of this paper is on power delivery in implantable devices. However, the method is general and can be applied to other applications that use wireless power transfer [9], [10]. Experimental results show that significant improvements in terms of power-transfer efficiency are achieved (compared to traditional inductively coupled two-coil systems). Measured results are in good agreement with the theoretical models and match well with the simulation results. Efficiency is enhanced by using high-Q factor coils (\( \sim 300 \)) and high coupling between driver and primary coils (\( k_{12} \approx 0.56 \)) and secondary and load coils (\( k_{34} \approx 0.50 \)). The prototype four-coil system achieves at least 2× more efficiency compared to prior art inductive links operating with comparable size and operating range. With external and implanted coils with diameters of 64 mm and 22 mm, respectively, and at an operating distance of 20 mm, a power-transfer efficiency of 82% is achieved. For an operating distance of 32 mm, efficiency slightly drops to 72%, which confirms the robustness of the four-coil-based power-transfer system when operating at long range.

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