Efficient Wireless Powering of Biomedical Sensor Systems for Multichannel Brain Implants
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Abstract—This paper describes the complete mathematical optimization process of an inductive powering system suitable for the application within implanted biomedical systems. The optimization objectives are thereby size, energy efficiency, and tissue absorption. Within the first step, the influence of the operational frequency on the given quantities is computed by means of finite element simulations, yielding a compromise of power transfer efficiency of the wireless link and acceptable tissue heating in terms of the specific absorption rate. All simulations account for the layered structure of the human head, modeling the dielectric properties with Cole–Cole dispersion effects. In the second step, the relevant coupling and loss effects of the transmission coils are modeled as a function of the geometrical design parameters, enabling a noniterative and comprehensible mathematical derivation of the optimum coil geometry given an external size constraint. Further investigations of the optimum link design also consider high-permeability structures being applied to the primary coil, enhancing the efficiency by means of an increased mutual inductance. Thereby, a final link efficiency of 80% at a coil separation distance of 5 mm and 20% at 20 mm using a 10-mm planar receiving coil can be achieved, contributing to a higher integration density of multichannel brain implanted sensors. Moreover, the given procedure does not only give insight into the optimization of the coil design, but also provides a minimized set of mathematical expressions for designing a highly efficient primary side coil driver and for selecting the components of the secondary side impedance matching. All mathematical models and descriptions have been verified by simulation and concluding measurements.

Index Terms—Biomedical electronics, brain–computer interfaces, coils, design optimization, energy efficiency, impedance matching, implants, inductive power transmission.

I. INTRODUCTION

NEUROLOGICAL disorders as Parkinson’s disease or epilepsy are topics of vivid ongoing research. Understanding the underlying mechanisms and treating them accordingly has been a challenge to an interdisciplinary research community, leading to sophisticated systems capturing the neural activity [1]. To acquire information in terms of electrical signals, electrocorticography (ECoG) sensing systems are directly applied to the cerebral cortex, providing a high spatial resolution of the neural activity with up to several hundred channels [2]. Moreover, the sensor output data can also be used within brain–computer interfaces, establishing a long-term connection between nervous system and different assistance facilities like computer cursors or robot arms, giving paralyzed patients the ability to interact with their environment [3].

With power consumption levels ranging from some microwatts for highly specialized system-on-chip devices up to 100 mW for multichannel systems also enabling the stimulation of the tissue (see [3]–[5]), considerable attention has to be attributed to a flexible and efficient power supply: while primary batteries being placed in the region of the chest suffer from additional surgery to implant the wired connection, from limited battery life and from the susceptibility of cable failure [6], energy harvesting devices do not provide sufficient energy to power the present generation of read-out devices [7]. In contrast, wireless powering has established as a reliable solution to overcome these challenges. Hereby, an inductive link consisting of two coils mutually coupled by their near field is used, as shown in Fig. 1.
case of cortical implants, three basic requirements have to be balanced:
1) a **high power transfer efficiency** to increase battery life while decreasing external battery size;
2) **small overall dimensions** not to compromise the patient’s comfort and to provide a high integration density of the sensor system;
3) a low specific absorption rate (SAR) to avoid tissue damaging.

Within the last few years, the optimization of the transmission coils for the best power transfer efficiency has been an important topic. Among others, Jow and Ghovanloo [8] presented an optimization algorithm for small planar spiral coils (PSCs) within biological tissue, achieving an efficiency of about 31% in the environment of muscle tissue with centimeter-sized coils. However, the approach suffers from the complexity of the iterative algorithm, while not revealing the best possible efficiency for the given size [9]. Poon et al. [10] investigated the use of frequencies in the gigahertz range, proving that systems with millimeter-sized receiver coils reach the optimum efficiency in the UHF range, although energy loss due to dielectric losses within the tissue is more significant there. An implementation of such a system was shown in [11], finally delivering microwatts of power at efficiency levels below 0.1%. The concept of modeling the link as general two-port was also taken by Zargham and Gulak [9], who performed numerical simulations on different case studies, achieving high efficiency levels at small coil size. However, their optimization procedure remains unknown and their coils show resistances of some milliohms, being close to parasitic resistances (which then highly degrade performance and aggravate the proposed impedance matching). Most recent publications focus on transmission systems using three or four coils [12], [13], claiming higher power transfer efficiency for greater coil separation distance. This concept is applicable if the additional coils are located in-between the transmitter and the receiver (i.e., decreasing the effective coil distance), while the effect resembles that of a simple impedance transformer for the case that the additional coils are directly placed on top of the transmitter and the receiver. Therefore, two out of four coils mainly perform an impedance matching, reducing the transmission system effectively to a two-coil system.

The beneficial effect of high-permeability materials being applied to the transmission coils has been investigated in terms of coil coupling and inductance by Hurley and Duffy [14] and Acero et al. [15]. However, their final impact on the power transfer efficiency of wireless links has been rarely analyzed and exploited.

Within this paper, we will present a comprehensive and straightforward procedure to obtain a wireless powering system with optimum efficiency for a given size constraint, also taking tissue absorption effects into account. After the introduction of important link quantities, we will deduce the most suitable operational frequency for a required output power from a basic finite element (FEM) simulation. Outgoing from the modeling of the coil parameters as a function of the geometrical dimensions, the coils will be optimized to yield a maximum power transfer efficiency. Finally, we elaborate the design of the required coil driver, rectification, and matching stages in the view of impedance matching. Measurements on implemented prototypes will verify the mathematical modeling and optimization process, also indicating the impact of biological tissue. This paper is a technical extension of [16], improving the coil optimization and giving additional insight into the optimum design of the required power electronics.

### II. Important Link Quantities

#### A. Efficiency Parameters

A general two-port network can be fully described by the four complex scattering parameters $S_{11}$, $S_{12}$, $S_{21}$, and $S_{22}$. Typically, electromagnetic field simulators as well as network analyzers can directly provide the $S$-parameters, which require a direct relation to the maximum link efficiency. Assuming conjugate complex matching of the two ports on the input and output, a quantity called maximum available gain (MAG) will directly address this best case transducer efficiency, which is defined as [17]

$$\text{MAG} = \max \left( \frac{P_{\text{out}}}{P_{\text{in}}} \right) = |S_{21}| \left( K - \sqrt{K^2 - 1} \right)$$

where $K$ is the Rollett stability factor

$$K = 1 + \frac{|S_{11}S_{22} - S_{12}S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}||S_{12}|}$$

An equivalent circuit of the coil system is given in Fig. 1(c) in the form of a pi-network. For coils being operated far below their resonant frequency, the capacitors $C_1$ and $C_2$ are negligible and the link can be represented by the remaining T-circuit. In this case, the MAG can be rewritten to the following expression using the well-known transformation formulas from $S$- to $Z$-parameters [17]:

$$\text{MAG} = \frac{1}{R_{12}^2 + X_{12}^2} \frac{1}{2\sqrt{(R_{11}R_{22} - R_{12})(R_{11}R_{22} + X_{12})}}$$

Here, $R_{ij}$ and $X_{ij}$ represent the real and the imaginary parts of the impedance matrix elements. For the inductive link of Fig. 1, it holds that $R_{ii} = R_i$, $X_{ii} = \omega L_i$, $X_{12} = \omega M_{12}$, and $R_{12} \approx 0$. Inserting these definitions into the general expression (3) and rearranging, the well-known formula for maximum link efficiency $\eta_{\text{link,opt}}$ is obtained, showing the equivalence of MAG and the formula frequently used in the literature

$$\eta_{\text{link,opt}} = \frac{\zeta}{(1 + \sqrt{1 + \zeta})^2} \quad \text{with} \quad \zeta = \frac{\omega^2 M_{12}^2}{R_1 R_2}$$

Note that this maximum link efficiency is monotonic function of the argument $\zeta$, i.e., it is sufficient to optimize $\zeta$ to yield maximum $\eta_{\text{link,opt}}$. When powering the link by a controlled voltage source of voltage $U_{\text{in}}$, the link efficiency is maximized for a particular complex load at the secondary side, as given by [18]

$$R_{L,\text{opt}} = \sqrt{\frac{R_2}{R_1}} \left( R_1 R_2 + \omega^2 M_{12}^2 \right)$$

$$|X_{L,\text{opt}}| = \omega L_2$$
B. Tissue Absorption

Besides the link efficiency, tissue absorption is an important quantity indicating the heating of the tissue due to the existing electromagnetic fields. It is expressed in terms of the SAR, which relates the overall tissue conductivity $\sigma_{\text{total}}$ at the frequency of interest, the magnitude of the E-field $|E|$ and the mass density of the tissue $\rho_m$.

$$\text{SAR} = \frac{\sigma_{\text{total}} |E|^2}{2\rho_m}. \quad (7)$$

Typically, the SAR value is averaged over a cubic volume containing a specific mass and needs to fulfill specified requirements: for example, $\text{SAR}_{\text{avg}}$ must not exceed $\text{SAR}_{\text{max}} = 1.6 \ \text{W/kg}$ averaging over 1 g in the United States, while European standards declare a limit of 2 W/kg for 10 g of averaging mass [17].

As SAR linearly scales with the active coil input power $P_{\text{in}}$, we define a normalized figure of merit

$$\text{SAR}_{\text{norm}} = \frac{\text{SAR}_{\text{avg}}}{P_{\text{in}}} = \frac{\text{SAR}_{\text{avg}}}{(1 - |S_{11}|^2)P_{\text{source}}} \quad (8)$$

Using the link efficiency $\eta_{\text{link}}$, we can define the maximum receivable power at the secondary side in order not to exceed the maximum allowed SAR

$$P_{\text{out,max}} = P_{\text{in,max}} \cdot \eta_{\text{link}} = \frac{\text{SAR}_{\text{max}}}{\text{SAR}_{\text{norm}}} \cdot \eta_{\text{link}}. \quad (9)$$

To model the tissue at an angular frequency $\omega$, a series of Cole–Cole dispersion terms accounts for the delay of the tissue’s electric dipoles in relation to the applied electromagnetic fields. The relative permittivity can consequently be written using numerical parameters, which can be found, e.g., in [19]

$$\varepsilon_r(\omega) = \varepsilon_\infty + \sum_{n=1}^{4} \frac{\Delta \varepsilon_n}{1 + (j\omega\tau_n)^{1-\alpha_n}}. \quad (10)$$

Using this description of a complex permittivity, the total conductivity taking dielectric losses into account is given by

$$\sigma_{\text{total}}(\omega) = \sigma_{\text{static}} + \sigma_0 \cdot \varepsilon_0 \cdot \text{Im}(\varepsilon_r(\omega)) \quad (11)$$

where the imaginary part of $\varepsilon_r$ contributes to a $\sigma_{\text{total}}$ rising with increased frequency.

III. TRANSMISSION INTERFACE DESIGN

A. Selection of the Operational Frequency

The optimization of the transmission coil systems starts with choosing an operational frequency which will provide a good tradeoff between efficiency and SAR. To see the general frequency behavior, we set up an FEM simulation following the model of Fig. 2: a small loop coil with a radius of $r_2 = 5 \ \text{mm}$ (representing the implant coil) is situated 20 mm within a stacked layer of biological tissue, each modeled with a permittivity being defined by the Cole–Cole dispersion according to (10) with numerical parameters being taken from [19]. Another loop coil, being situated 2 mm above the tissue, serves as transmission structure. The copper traces were chosen to have a width of 200 $\mu\text{m}$ and a height of 50 $\mu\text{m}$. Ansoft HFSS is used as a simulation tool.

From the given simulation setup, the two-port $S$-parameters and the local SAR distribution were extracted for the frequency range from 1 MHz to 1 GHz and for different primary coil radii $r_1$. Subsequently, the maximum link efficiency $\eta_{\text{link,opt}}$, the SAR per input power $\text{SAR}_{\text{norm}}$, and the maximum receivable power $P_{\text{out,max}}$ are computed following (1), (8), and (9), respectively. The corresponding results are shown in Figs. 3–5. The SAR is thereby averaged over the 1-g volume with the highest field strength being directly located underneath the primary coil trace, assuming a maximum value $\text{SAR}_{\text{max}} = 1.6 \ \text{W/kg}$.

The simulated efficiency curves show the tendency implied by (4): in general, a larger primary coil radius leads to increased mutual inductance $M_{12}$ and thus to increased efficiency. For frequencies in the lower megahertz range, the efficiency is highly improved for increased frequency, following $\zeta \propto f^2$. At around 5 MHz, skin effect ($R \propto \sqrt{f}$) starts to become dominant for the given conductor cross section, thus the slope of $\eta_{\text{link}}$ within the double logarithmic plot decreases.
due to $\zeta \propto f$. At a certain point, dielectric losses within the tissue start to influence the behavior. As shown by the simulation, larger coil radius means that this loss type becomes dominant at lower frequencies. Therefore, the frequency of optimal efficiency is lower for coils of extended size.

The SAR per input power strongly increases with frequency due to augmented dielectric losses and electromagnetic fields. As a result, the maximum power that can be received at the secondary side without violating the SAR limits decreases with frequency as the gain in efficiency does not compensate for the lower primary side input power that is allowed to be emitted.

In terms of efficiency, a frequency in-between 10 and 200 MHz would be acceptable, peaking around 50–100 MHz when considering coils with a low number of windings. If, however, aspiring for available power levels around 100 mW, frequencies below 20 MHz must be considered. Thus, the Industrial Scientific Medical band at 13.56 MHz is a suitable choice for these medium power levels and is therefore chosen within this paper.

At this place, the attentive reader might ask if the numerical simulations on efficiency and absorption are still valid if considering multiturn coils. As both efficiency and the tissue loss scale approximately linear with the number of turns (for small $\zeta$), the receivable power will approximately stay constant. Thus, the operational frequency is still to choose by the power requirements, but the resulting efficiency levels will be increased.

When using coils of greater trace width, only the efficiency will be enhanced, also increasing the receivable power. Thus, the analysis can be seen as a worst case estimation it terms of trace width.

**B. Simple Coil Model**

For the design of the coils, we follow the approach of using PSCs, as shown in Fig. 1. In contrast to classical litz wire coils, PSCs are very reproducible, which is mandatory for performing mathematical analysis and optimization. For the given frequency range, multiturn coils are highly preferable against loop coils as going to a higher number of turns will scale both $M_{12}$ and $R_t$ in the same direction, but as $\zeta$ scales with the square of $M_{12}$, the efficiency is likely to be increased. In the following, the important coil equivalent circuit parameters will be expressed by the geometrical coil parameters of Fig. 6(a), which are the respective number of turns $N_n$, the outer coil radius $r_{n,max}$, the trace width $w_n$, and the trace pitch $p_n$ (with $n$ denoting the number of the coil). From this, these parameters will be related to the maximum link efficiency, allowing a numerical optimization to yield the most efficient coil system.

1) **Mutual Inductance:** Under the condition that the transmitter coil is much larger than the receiver coil, the flux $\Phi_{12}$ defining the mutual inductance $M_{12}$ can be approximated by multiplying the external coil magnetic field $B_{1,total}$ at the location of the receiver coil by the effective area of the small receiver coil, resulting in

$$M_{12} = \frac{\Phi_{12}}{I_1} \approx B_{1,total} \cdot A_{2,total} = \left(\sum_{i=1}^{N_1} B_{1,i} \right) \cdot \left(\sum_{j=1}^{N_2} A_{2,j} \right)$$

$$= \left(\sum_{i=1}^{N_1} \frac{\mu_0 \cdot r_{1,i}^2}{2 \left(r_{1,i}^2 + a_{1, coils}^2\right)^{3/2}}\right) \cdot \left(\sum_{j=1}^{N_2} \pi \cdot r_{2,j}^2\right) := M_A \cdot M_B. \quad (12)$$

Here, $B_{1,i}$ is the magnetic field on the axis of coil 1 at a distance $d_{c, coils}$ (stemming from turn $i$ of coil 1) and $A_{2,j}$ is the area enclosed by turn $j$ of coil 2.

The corresponding error due to the given simplification is shown in Fig. 7. Moreover, it can be stated that the mutual
inductance can be decomposed into two factors $M_A$ and $M_B$, separating the influence of primary and secondary coil.

For a PSC, a particular turn does not have a constant radius. Therefore, we approximate the radius $r_{n,i}$ of coil $n$ and turn $i$ by

$$r_{n,i} = r_{n,\text{max}} - (i - 1/2)\rho_n.$$  \hspace{1cm} (13)

2) Loss Resistance: For the description of the resistance, a semiempirical model partially based on simulation data was set up. It uses the typical dc resistance according to $\rho \cdot l_n / (w_n \cdot h_n)$ for lower frequency, while assuming that the effective conducting cross section at higher frequency is proportional to the product of $w_n$ and the skin depth $\delta(f) = (\rho/(\pi f \mu_0 \mu_r))^{1/2}$. The proportionality constant $k$ is dependent on the ratio of line width $w_n$ and height $h_n$ and is determined by simulation. It follows the fit function extracted from an HFSS simulation:

$$k(w_n/h_n) \approx 0.9674 \cdot e^{-0.0858 w_n/h_n} + 0.5.$$  \hspace{1cm} (14)

Finally, the resistance can be summarized to

$$R_n(f) = \begin{cases} \frac{\rho}{w_n \cdot h_n} \sum_{i=1}^{N_n} 2\pi r_{n,i}, & \text{for } f < f_T \\ \frac{\rho}{w_n \cdot k(w_n/h_n) \cdot \delta(f)} \sum_{i=1}^{N_n} 2\pi r_{n,i}, & \text{for } f > f_T. \end{cases}$$  \hspace{1cm} (15)

Here, $f_T$ is the frequency at which the skin effect becomes dominant, resulting from $R_n(f)$ being a steady function at this transition frequency

$$f_T = \frac{k(w_n/h_n)^2 \cdot \rho}{h_n^2 \pi \mu_0 \mu_r}.$$  \hspace{1cm} (16)

C. Advanced Coil Parameters

Although not used for the optimization of the power transfer efficiency (see next chapter), coil inductance $L_n$ and the parasitic coil capacitance $C_n$ determine the frequency behavior of the coil [see the coil equivalent circuit of Fig. 6(c)]. This is important for the system design when interfacing the coil with electronics and when performing impedance matching. Finally, the coil impedance is given by the parallel combination of the $RL$-branch and the capacitance

$$Z_{\text{coil},n}(f) = \frac{L_n + R_n(f)}{2\pi f C_n \cdot (L_n + R_n(f)) + 1}. $$  \hspace{1cm} (17)

Therefore, we will give insight how inductance and capacitance can be derived.

1) Inductance: Considering the operation well below the self-resonant frequency, the self-inductance can be extracted by the well-known expression

$$L_n = \sum_{i=1}^{N_n} \frac{N_n}{i} \sum_{j=1}^{N_n} M_{ij} \cdot (1 - \delta_{ij})$$  \hspace{1cm} (18)

with $L_i$ being the self-inductance of winding $i$ according to [20], $M_{ij}$ the mutual inductance of turn $i$ and $j$ [15], and $\delta_{ij}$ the Kronecker delta function, given by

$$L_i = \mu_0 \pi r_i \int_0^\infty J_1(k) J_1 \left(\frac{r_i - w_j}{r_i} \right) dk$$  \hspace{1cm} (19)

$$M_{ij} = \mu_0 \pi r_i r_j \int_0^\infty J_1(r_i) J_1(r_j) dk$$  \hspace{1cm} (20)

$$\delta_{ij} = \begin{cases} 0, & \text{for } i \neq j \\ 1, & \text{for } i = j. \end{cases}$$  \hspace{1cm} (21)

2) Parasitic Coil Capacitance: Especially when using the multiturn setup, parasitic capacitive effects being represented by the lumped capacitors $C_1$ and $C_2$ shown in Fig. 1 need to be considered for the correct impedance matching. The turn-to-turn capacitance per length of adjacent traces can be derived using the coplanar stripline model of [21], resulting in

$$C_{\text{unit},n} = e_0 \left( \varepsilon_{r,\text{vol}} \frac{K(\sqrt{1-k^2})}{K(k)} + (\varepsilon_{r,\text{sub}} - \varepsilon_{r,\text{vol}}) \frac{n_{\text{side}} K(\sqrt{1-k^2})}{K(k_1)} \right)$$  \hspace{1cm} (22)

with $K(x)$ being the complete elliptic integral of the first kind and

$$k = \frac{p_n - w_n}{p_n + w_n} \quad k_1 = \frac{\sin \left( \frac{2(p_n-w_n)}{4n_{\text{sub}}} \right)}{\sin \left( \frac{2(p_n+w_n)}{8n_{\text{sub}}} \right)}.$$  \hspace{1cm} (23)

All situations are simplified to the arrangement shown in Fig. 6(b), where the coil traces are surrounded by an infinite volume with the permittivity $\varepsilon_{r,\text{vol}}$, either being wrapped completely ($n_{\text{side}} = 0$), on a substrate of $\varepsilon_{r,\text{sub}}$ and height $h_{\text{sub}}$ ($n_{\text{side}} = 1$) or in a coating of $\varepsilon_{r,\text{sub}}$ and one-sided height $h_{\text{sub}}$ ($n_{\text{side}} = 2$). For a first estimation of the capacitance, it is sufficient to multiply $C_{\text{unit}}$ with the length of the gap in-between the traces and to divide the result by the number of turns $N_n$. However, for a more accurate modeling, the distributed nature of these capacitive elements has to be considered, following the equivalent circuit model of Fig. 6(d). The corresponding parameters for the turn-to-turn capacitance $C_{i,i+1}$ and the effective turn inductance $L_{\text{turn},i}$ can be calculated using

$$C_{i,i+1} = C_{\text{unit},n} \cdot r_{n,i}$$  \hspace{1cm} (24)

$$L_{\text{turn},i} = L_i + \sum_{j=1}^{N_n} M_{ij} (1 - \delta_{ij}).$$  \hspace{1cm} (25)

Setting up the corresponding model within a simulation tool and capturing the resonance frequency $f_{\text{SRF}}$ (the frequency of the imaginary part’s singularity), the effective coil capacitance can be estimated

$$C_n \approx \frac{1}{(2\pi f_{\text{SRF}})^2 L_n}. $$  \hspace{1cm} (26)
3) Nearby High-Permeability Material: If applying a planar layer of high-permeability material (with \( \mu = \mu_0 \mu_r - j \mu_0 \mu_i \) and negligible conductivity, i.e., we consider a ferrite) to the primary coil as shown in Fig. 8, loss resistance, self-inductance, and mutual inductance to the secondary coil will increase. Following the mathematical derivation in [14], the mutual inductance will be supplemented by:

\[
\Delta M_{12, \text{PSC}} \approx \sum_{i=1}^{N_1} \sum_{j=1}^{N_2} \mu_0 \pi r_i r_j \left( \frac{\mu_r - 1}{\mu_r + 1} \right) \cdot \int_0^\infty J_1(kr_i) J_1(kr_j)  \left( 1 - e^{-2k h_{\text{HP}}} \right) \left( 1 - \frac{\mu_0 - 1}{\mu_r + 1} e^{-k d_{\text{coils}} + 2h_{\text{sub}}} \right) dk \]

(27)

where \( h_{\text{HP}} \) is the height of the high-permeability material and \( h_{\text{sub}} \) corresponds to the substrate thickness separating the copper traces from the ferrite. At the same time, the loss resistance will increase due to two effects.

1) There will be magnetic switching losses within the additional material.

2) The increased magnetic fields around the current conducting traces will alter the current distribution over the conductor cross section so that, e.g., proximity losses will be increased.

Due to this very complex behavior, the loss resistance is preferably computed by means of an FEM simulation.

D. Coil Optimization

Finally, the geometrical parameters should be selected to yield the highest power transfer efficiency. To simplify the design process, the coils shall be optimized without considering the high-permeability material, which is simply applied in the second step to boost the performance. As indicated earlier, the optimization of \( h_{\text{link}} \) according to (4) just requires maximizing \( \zeta = \frac{\omega^2 M_{12}^2}{(R_1 \cdot R_2)} \). We recall that \( M_{12} \) can be decomposed into \( M_A \) and \( M_B \), separating the influence of primary and secondary coil. Therefore, \( \zeta \) can also be decomposed, yielding the optimization quantities \( \zeta_1 \) and \( \zeta_2 \), optimizing both coils independently from each other

\[
\zeta = \frac{\omega^2 M_{12}^2}{R_1 \cdot R_2} = \left( \frac{\omega M_A^2}{R_1} \right) \left( \frac{\omega M_B^2}{R_2} \right) = \zeta_1 \cdot \zeta_2. \]

(28)

When seeking the correct expression of \( R_1 \) and \( R_2 \) to be inserted, there are two possibilities:

1) the coil loss resistance, as given by (15);

2) the complete resistive (real) part of the coil impedance, also including the contribution of the coil capacitance (this is, e.g., considered by [8]).

Taking into consideration that the parasitic parallel capacitance [as shown in Fig. 1(b) and (c)] does not introduce any lossy element and that there might even be some matching strategies using parallel resonant circuits, it is mandatory to neglect the capacitance when calculating the maximum efficiency and when optimizing the coil geometries, i.e., expression (15) is directly to be used as a loss resistance. Thus, self-resonant behavior (i.e., low \( Q = \text{Im}(Z_{\text{coil}})/\text{Re}(Z_{\text{coil}}) \)) measured at the coils terminals) does not necessarily sacrifice efficiency and the parasitic capacitance must be considered in the step of impedance matching, but not when optimizing the transmission interface for maximum efficiency. Subsequently, the trace pitch \( p_n \), which increases capacitance when being decreased, does neither influence loss resistance nor maximum efficiency (as long as the proximity effect can be neglected).

Therefore, each coil is left with three parameters: 1) \( r_{n, \text{max}} \); 2) \( N_n \); and 3) \( w_n \). As increased coil size is always favorable as long as the currents are kept uniform along the line, \( r_{n, \text{max}} \) is chosen as large as possible under the given size constraints. Then, \( \zeta_1 \) and \( \zeta_2 \) can be plotted over the number of windings \( N_n \) and the trace width \( w_n \) for both cases by inserting (12) and (15) into (28). The combination of number of windings and trace width which yields in the highest possible \( \zeta \) is chosen as coil dimension. Although \( C_n \) does not affect theoretical performance, a lower \( C_n \) means that the coil
TABLE I

<table>
<thead>
<tr>
<th>Design parameter</th>
<th>Coil 1</th>
<th>Coil 2</th>
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<tbody>
<tr>
<td>r_{tr,\text{max}}</td>
<td>15 mm</td>
<td>5 mm</td>
</tr>
<tr>
<td>N_{tr}</td>
<td>3</td>
<td></td>
</tr>
<tr>
<td>w_{tr}</td>
<td>1.0 mm</td>
<td>0.4 mm</td>
</tr>
<tr>
<td>d_{tr}</td>
<td>1.5 mm</td>
<td>0.7 mm</td>
</tr>
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will be operated further away from its resonance frequency, i.e., approaching tissue will be less likely to influence $C_n$ and therefore the effective input impedance. So, a smaller number of turns is favorable if delivering the same value of $\zeta$.

For the given scenario of the brain implant, we chose $r_{1,\text{max}} = 15 \text{ mm}$ and $r_{2,\text{max}} = 5 \text{ mm}$, with trace spacings of $s_1 = p_1 - w_1 = 0.5 \text{ mm}$ and $s_2 = p_2 - w_2 = 0.3 \text{ mm}$ (being chosen according to the available manufacturing technology). The coil distance also influencing $\zeta_1$ is set to $d_{\text{coils}} = 20 \text{ mm}$, being a typical implant depth for the brain environment. The corresponding results of $\zeta_1$ and $\zeta_2$ are shown in Figs. 9 and 10, with the optimum design parameters being extracted to Table I. After acquiring the dimensions, the ferrite structures that are being depicted in Fig. 8 are applied to the primary coil. Thereby, a ferrite material showing low losses at the operational frequency is highly recommended. For the given setup, we use a Würth WS-FSFS364 material. The shielding structure consists of a planar layer of height $h_{\text{HP}} = 0.6 \text{ mm}$ and a cylindrically shaped core with a diameter of $d_{\text{core}} = 8 \text{ mm}$.

IV. ELECTRONIC DESIGN

As mentioned within the technical introduction, besides coil optimization, a suitable impedance matching is required for an energy efficient system. Within this section, we will provide a guideline to the design of a simple electronic interface using an exemplary implant device. Therefore, we consider a custom-made eight-channel electrophysiological readout-system (e.g., to capture signals from ECoG electrodes) using a Texas Instruments ADS1298 signal conditioning chip with an MSP430FR5738 microcontroller. Its current consumption is approximately $I_{\text{DD2}} = 5 \text{ mA}$ at a supply voltage of $U_{\text{DD2}} = 3.3 \text{ V}$. For the power supply system design, we consider a setup of Fig. 11: while the primary side is powered by a class E amplifier, the secondary side uses an L-matching network consisting of the reactive elements $X_{M1}$ and $X_{M2}$, followed by a Greinacher voltage doubling rectifier and a low-dropout regulator (we will neglect the parasitic capacitor $C_2$, as it is small for small coils). The system design is started at the secondary side, coping with the power transmitter in the second step.

A. Implant Electronics

First of all, the equivalent load impedance has to be determined. Therefore, the dc load resistance being present at the rectifier’s output can be approximated by

$$R_{\text{Load}} = \frac{U_{\text{DD2}} + \Delta U_{\text{LDO}}}{I_{\text{DD2}}}$$

with the anticipated regulator voltage drop $\Delta U_{\text{LDO}}$. In principle, the input impedance of rectifier and load can also be represented as a parallel combination of a resistance $R_{\text{par}}$ and a capacitance $C_{\text{par}}$. This transformation process is shown for different values of $R_{\text{Load}}$ in Fig. 12, allowing one to determine the effective ac input resistance $R_{\text{par}}$ for the given rectifier setup. This consists of a voltage doubler rectifier, being composed out of two Schottky BAT754 diodes. In our

**Fig. 11.** Optimized inductive link being interfaced by a class E coil driver and an L-matched Greinacher rectifier.

**Fig. 12.** Impedance transformation of the Greinacher rectifier using BAT754 diodes at a frequency of 13.56 MHz for different rectifier input power levels. The data are acquired by a large scale S-parameter simulation using Agilent Advanced Design System 2013 and the corresponding SPICE diode models.
case example, we set $\Delta U_{LDO} = 0.3$ V, therefore resulting in $R_{\text{Load}} = 720$ $\Omega$ and $R_{\text{par}} \approx 140$ $\Omega$ (the last step being deduced from Fig. 12).

To design the matching network, we simply require solving the following equations for $X_{M1}$ and $X_{M2}$, with $R_L, opt$ and $X_L, opt$ being calculated with (5) and (6):

$$R_{L, opt} = \Re \left( jX_{M1} + \frac{jX_{M2} \cdot R_{\text{par}}}{jX_{M2} + R_{\text{par}}} \right)$$

$$X_{L, opt} = \Im \left( jX_{M1} + \frac{jX_{M2} \cdot R_{\text{par}}}{jX_{M2} + R_{\text{par}}} \right).$$

(30)

(31)

For a powering distance of 15 mm ($M_{12} \approx 11$ nH), we obtain the reactances $X_{M1} \approx −1.78$ $\Omega$ and $X_{M2} \approx −7.79$ $\Omega$, resulting in capacitors of $C_{M1} = 6.6$ nF and $C_{M2} = 1.51$ nF.

B. External Power Transmitter

For the design of the primary side coil driver, the highly efficient class E topology is chosen. By a suitable dimensioning, either current through or the voltage over the switching transistor $T_1$ is zero, allowing efficiency levels of up to 90%. Our design guidelines are following the approach of [22], but with a smaller set of equations. For the class E design, we utilize the primary coil as inductor of the class E RLC output network (see [18] for detailed information). To start, the input impedance of the link has to be determined. For the class E design, the following equations for $\sin(\phi)$ and $\cos(\phi)$ are given

$$\sin(\phi) = \frac{2y^2 - 1}{\sin(y)} - 2 \sin(\phi) \sin(y)$$

$$\cos(\phi) = \frac{2y \cot(y) - 1}{\csc^2(y)} - 3y \cot(y) - 1.$$

(36)

(37)

The influence of biological tissue being present within the inductive interface is also considered within this research. Therefore, a measurement of the full two-port $S$-parameters is performed for a coil separation distance of 20 mm within the parallel capacitor $C_{E2}$ and the gate duty cycle $D$ are given by

$$C_{E2} = \frac{2y^2 + 2y \sin(\phi) - 2 \sin(\phi) \sin(y)}{\cos(y) g^2 R_{in}}$$

$$D = 1 - y / \pi.$$
Fig. 13. Measured versus calculated coil impedance over frequency for the designed primary (left) and secondary (right) side coil structure.

Fig. 14. Measured efficiency and primary coil quality factor $Q_1$ over frequency for a coil distance of 20 mm. Two setups are characterized: the transmission interface in air and the secondary coil being embedded within a 13-mm layer of pork muscle tissue.

It is noteworthy that the efficiency is barely influenced by the tissue up to 30 MHz, while slightly dropping above due to additional dielectric losses. However, the primary coil $Q$-factor is starting to be influenced from about 10 MHz, which is because of the parasitic capacitance $C_1$ being slightly increased. Moreover, it is clearly shown that although $Q_1$ is strongly decreasing when approaching the self-resonant frequency, the efficiency does not follow this behavior. Thus, the tissue’s influence, especially at 13.56 MHz, is exclusively limited to the parasitic capacitance, which has to be accounted for within the process of impedance matching. It is therefore reasonable to coat the coil with a material of defined permittivity (as suggested by [8]), but for the purpose of a controlled capacitance.

Fig. 15. Measured best case link efficiency versus coil separation distance at an operational frequency of 13.56 MHz, including additional ferrite structures being applied to the primary coil. The dashed curves are calculated using the model developed within this paper.

Fig. 16. Power transfer efficiency for state-of-the-art coil systems reported in the literature. The coloring represents different coil spacings. The sources are: 1 $\rightarrow$ [23], 2 $\rightarrow$ [24], 3 and 4 $\rightarrow$ [8], 5 $\rightarrow$ [11], 6–8 $\rightarrow$ [9], 9 $\rightarrow$ [12], and star marker $\rightarrow$ this paper.

C. Link Performance

Finally, the link performance of the coil system, especially accounting for the different types of ferrite structures, is evaluated. Therefore, the S-parameters are captured for different distances adjusted by plastic spacers, allowing to calculate the optimum efficiency [see (1)] for the case of optimum impedance matching. The measurement results are shown in Fig. 15, together with the efficiency values derived by the given modeling. Hereby, the coil loss resistance including the ferrite was computed by an HFSS simulation, yielding in $R_{1,\text{ferrite}} \approx 1.4 \Omega$. Again, an excellent agreement of calculation and measurement can be registered. For small coil distances corresponding to a subcutaneous placement of the secondary coil, efficiency levels exceeding 80% can be realized. For the implantation under the skull ($d_{\text{coils}} \approx 20 \text{ mm}$), 20% are still achievable. Therefore, the coils can be considered to provide a good ratio of efficiency and size, also being proven by the state-of-the-art comparison of Fig. 16. The improvements due to the ferrite structure itself are most beneficial at higher coil separation distance (around 20 mm), where the efficiency is relatively increased by a factor of about 1.2. Another advantage of the ferrite can be seen in its shielding effect, allowing to place conductive elements (e.g., the primary side reader electronics) on top of the coil without sacrificing efficiency.

Fig. 17 shows the distribution of the SAR within the tissue for the designed ferrite enhanced planar spiral coil as well...
electrophysiological read-out, together with additional electronics for communication purpose. The system performance in terms of efficiency is evaluated both in simulation and measurement for a broad range of load impedances, as shown in Fig. 19. While the link efficiency is maximized around the nominal design goal of around 800 $\Omega$ because of the corresponding matching network, the total efficiency peaks at slightly higher load resistances. This is due to the rectifier efficiency, which will steadily increase for even higher load. The class E amplifier exhibits an efficiency of around 90% at the load of interest. Again, both simulation and measurement are in very good agreement, also validating the given design approach.

VI. CONCLUSION

Within this paper, a general design approach to wireless powering systems targeting the application of brain implanted sensor has been given. Starting from the fundamental question of the operational frequency, a straight-forward noniterative way of optimizing the transmission interface was deduced. For the advanced system designer, additional equations to determine inductance and parasitic coil capacitance are given; they have been proven to yield good accuracy in terms of parameter prediction. Besides, the approach does not only include coil design, but also provides generalized and compact design equations to perform a suitable impedance matching. While this is achieved for a constant coil separation distance, optimum load impedance and input impedance will vary with nonconstant coupling conditions. Allowing the best case efficiency in all possible coil alignment situations by a dynamic impedance matching is one of the most recent challenges within the wireless powering of biomedical implants and will therefore be addressed by the authors in future publications.

REFERENCES


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