# Designing Efficient Inductive Power Links for Implantable Devices

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***Abstract*—Due to limited battery life and size limitations, many implantable biomedical devices must be powered inductively. Because of weak coupling between implanted and external coils, obtaining high power efficiency is a challenge. Previous authors have addressed the issue of optimizing power efficiency in these systems. In this paper, we further this analysis for the case of planar spiral "pancake" coils at low RF frequencies (100 kHz – 10 MHz). We consider practical design constraints such as component variation, power amplifier limitations, and coil voltage limits. We introduce a new, complete expression for total power link efficiency.**

## I. INTRODUCTION

Implantable biomedical devices are powered either from a long-life battery (e.g., pacemakers) or a wireless inductive link (e.g., cochlear implants). As the size of implantable devices shrinks, it is difficult to use battery technology to achieve a multi-year lifespan. Even rechargeable batteries have a limited number of recharge cycles before they become ineffective. For new cortical recording devices under development (e.g., [1]), it is important to minimize the size and mass of the implant so that the device "rides along" with the malleable brain tissue. With modern micromachining processes, small high-quality coils can be mass produced [2]. These coils have low mass, consume very little volume, and do not contain toxic chemicals present in many batteries.

Inductive power links for biomedical applications have been studied extensively during the past few decades [3-4]. Most of these early analyses considered the use of solenoidal coils, which are useful for powering small devices in limbs. For cortical recording devices resting on the surface of the brain, planar spiral "pancake" coils are a better choice given the limited headroom between cortex and skull. In the external power unit, planar coils can fit flat against the body and be integrated into clothing, providing a cosmetic advantage. In this paper, we develop a method for analyzing and then optimizing an inductive power link between two planar spiral coils.

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## II. INDUCTIVE POWER LINK OPTIMIZATION

### A. Transmit and Receive Circuit Analysis

An inductive power link consists of a transmitting coil having inductance $L_T$ and a receiving coil $L_R$ (see Fig. 1). Some magnetic flux is shared between the coils, resulting in a coupling coefficient $k$, where $0 < k < 1$. An ac voltage of amplitude $V_T$ is applied across the transmitting coil, and this induces an ac voltage $V_R$ on the receiving coil. The receiving coil is connected to a load $R_L$. Thus, the power delivered to the load is given by

$$P_L = \frac{V_{R\text{pk}}^2}{2R_L} \tag{1}$$

and the power drawn from the supply is $P_S = V_S I_S$. The overall efficiency of the inductive link is $\eta = P_L/P_S$.

Since narrowband operation is typically used, capacitors $C_T$ and $C_R$ can be used to create resonant circuits that boost the voltages across the coils. In the absence of magnetic coupling, the quality factor of the series *RLC* transmitting circuit is given by $Q_T = \omega_0 L_T/R_T$, where $\omega_0 = 1/(L_T C_T)^{1/2} = 1/(L_R C_R)^{1/2}$ is the frequency of oscillation and $R_T$ is a combination of the transmit coil series resistance, equivalent series resistance (ESR) of $C_T$, and the output resistance of the power amplifier. The quality factor of the receiving circuit *in the absence of a load* is given by $Q_R = \omega_0 L_R/R_R$, where $R_R$ is a combination of the receive coil series resistance and the ESR of $C_R$.

The receiver resonant circuit can be approximated using a narrowband equivalent circuit model [see Fig. 1(b)] and assuming $Q_R^2 >> 1$ so that $Q_R^2 + 1 \approx Q_R^2$. Using this model, we observe that when $R_L$ is added, some power will be wasted as heat due to losses in the resonant circuit. We can define the *receive efficiency* $\eta_R$ for the receiver circuit as

$$\eta_R = \frac{1}{1+\dfrac{R_L}{Q_R^2 R_R}} = 1 - \frac{Q_R'}{Q_R} \tag{2}$$

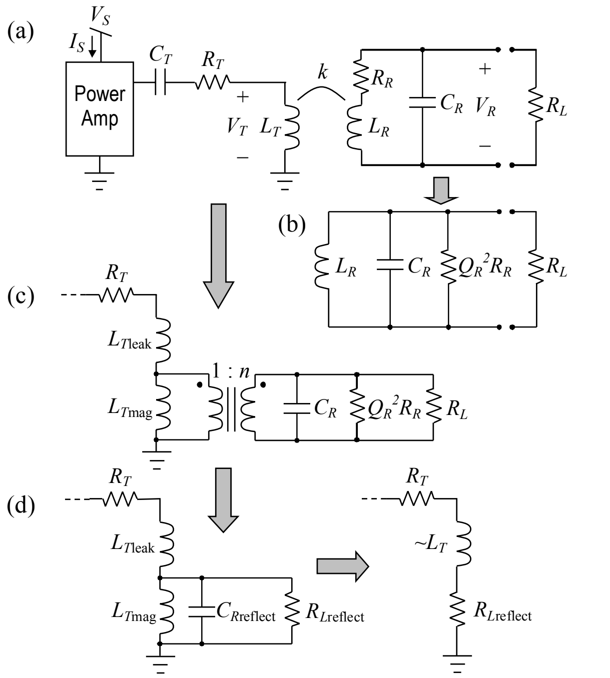


Figure 1. Schematic and equivalent circuit diagrams for inductive power link.

where

$$Q'_R = \frac{\left(Q_R^2 R_R\right) \| R_L}{\omega_0 L_R}. \tag{3}$$

$Q'_R$ is the quality factor of the receiving circuit *with load*. Thus, for high efficiency we should ensure that $Q_R^2R_R >> R_L$. For high receive efficiency, $Q'_R$ will be much less than $Q_R$.

The weak coupling between transmit and receive coils can be modeled as an ideal transformer, a leakage inductance $L_{T\text{leak}} = (1 - k^2)L_T$, and a magnetizing inductance $L_{T\text{mag}} = k^2L_T$ [5]. In the weakly-coupled case where $k^2 << 1$, we can approximate $L_{T\text{leak}} \approx L_T$. The "turns ratio" $n$ of the ideal transformer is given by $n = (1/k)(L_R/L_T)^{1/2}$. Fig. 1(c) shows the equivalent circuit using this model. Reflecting the capacitance $C_R$ and the resistance $R_L$ through the ideal transformer, we get values of $C_{R\text{reflect}} = (L_R/L_T)(C_R/k^2)$ and $R_{L\text{reflect}} = k^2(L_T/L_R)R_L = k^2Q_TQ'_RR_T$. It can be shown than $L_{T\text{mag}}$ and $C_{R\text{reflect}}$ resonate at $\omega_0$, resulting in a total reflected impedance that is purely resistive and equal to $R_{L\text{reflect}}$, as shown in Fig. 1(d).

The inductively coupled load can therefore be viewed as a resistance in series with the transmit coil; if we deliver power $P$ to $R_{L\text{reflect}}$, then $\eta_RP$ will be delivered to the load. The efficiency with which we deliver power to this reflected resistance is limited by the voltage divider formed by $R_{L\text{reflect}}$ and the transmit coil losses $R_T$. This allows us to derive the *transmit efficiency* $\eta_T$ with which we deliver power to $R_{L\text{reflect}}$, a result first derived in [3]:

$$\eta_T = \frac{k^2 Q_T Q'_R}{1 + k^2 Q_T Q'_R} = k^2 Q'_T Q'_R \tag{4}$$

where $Q'_T = \omega_0L_T/(R_T + R_{L\text{reflect}})$ is the quality factor of the *loaded* transmit circuit.

Finally we must consider the properties of the power amplifier driving the transmit coil. Like many previous designs, we use a class E amplifier to drive the coil [6]. The class E configuration requires only a single nMOS switch and has a theoretical efficiency of 100% [7]. In practice, efficiencies greater than 85% are routinely reported [8]. From equations (1) and (2) in [7], it can be shown that the peak ac voltage generated on a transmit coil powered by a class E amplifier operating from a dc supply $V_S$ is ideally

$$V_{T\text{pk}} = \frac{2}{\sqrt{\pi^2/4 + 1}} Q_T V_S = A_{\text{PA}} Q_T V_S \tag{5}$$

where the "voltage gain" of the power amplifier $A_{\text{PA}} \approx 1.07$. (In experiments, we observe $A_{\text{PA}}$ closer to 0.85.) The voltage gain from transmit coil to receive coil is given by

$$\frac{V_{R\text{pk}}}{V_{T\text{pk}}} = k\sqrt{\frac{L_R}{L_T}} Q'_R. \tag{6}$$

Thus, the total voltage gain from supply voltage $V_S$ to peak load voltage $V_R$ is

$$\frac{V_{R\text{pk}}}{V_S} = A_{\text{PA}} k \sqrt{\frac{L_R}{L_T}} Q'_T Q'_R. \tag{7}$$

We can now introduce a new, complete expression for the *total* efficiency of an inductive power link:

$$\eta = \eta_{\text{PA}} \eta_T \eta_R \tag{8}$$

where $\eta_{\text{PA}}$ is the efficiency with which the power amplifier delivers supply power ($V_SI_S$) to the total resistance in the loaded transmit coil: $R_T + R_{L\text{reflect}}$. In real class E circuits, some power is wasted in the switch MOSFET and by the circuit that drives the MOSFET gate at $\omega_0$ (although resonant gate driver circuits can lower this power). The receive efficiency $\eta_R$ can be close to one if the unloaded receive coil has a high $Q_R$. Thus, the total power efficiency of an inductive link is primarily dependent on the transmit efficiency $\eta_T$. To maximize $\eta_T$, $Q'_T$ and $Q'_R$ must be made as high as is practical.

However, there is a danger in using high values of $Q$. The bandwidth of a resonant circuit is given by $\omega_0/Q$, so high $Q$ implies a narrow operating region. Component variation, stray capacitances, and the presence of conductive materials near the coil can easily shift the resonant point of an $LC$ circuit by a few percent. Thus, $Q'_R$ should be limited to a value that allows for expected variation in the precise resonant frequency of the receive circuit. Similarly, $Q'_T$ of the loaded transmit circuit should be limited, and this places a bound on the maximum value of $Q_T$. (We could *vary* $\omega_0$ to track changes in $L_T$ as in [6], but this would require a lower $Q'_R$.) High values of $Q_T$ can also lead to dangerously high transmit coil voltages and limited bandwidth for telemetry over the power signal. With $Q_TQ'_R$ limited by potential component variation and safety considerations, we must focus on maximizing the coupling coefficient $k$ to achieve high transmit efficiency $\eta_T$.

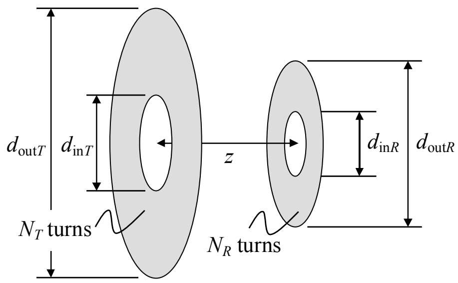


Figure 2. Geometry for two circular planar "pancake" coils.

### *B. Coupling Coefficient Optimization*

The mutual inductance of two aligned circular filaments (i.e., single-turn coils) having diameters $d_T$ and $d_R$ and separated by distance $z$ is given by the following exact expression [4]:

$$M = \frac{1}{2}\mu_0\sqrt{d_T d_R}\left[\left(\frac{2}{f} - f\right)K(f) - \frac{2}{f}E(f)\right] \qquad (9)$$

where $K(f)$ and $E(f)$ are the complete elliptic integrals of the first and second kind, respectively, $\mu_0 = 4\pi \times 10^{-9}$ H/cm, and

$$f(d_T, d_R, z) \equiv \sqrt{\frac{4d_T d_R}{(d_T + d_R)^2 + z^2}}\,. \qquad (10)$$

To estimate the mutual inductance of two planar spiral coils, each having outer diameter $d_{out}$ and inner diameter $d_{in}$ (see Fig. 2), we use Lyle's method from 1902 [9], which approximates each coil as two circular filaments having diameters $d_{avg} \pm (d_{out} - d_{in})/(8\sqrt{3})$, where $d_{avg} = (d_{out} + d_{in})/2$. If the transmit coil has $N_T$ turns and is represented by two circular filaments $A$ and $B$, and the receive coil has $N_R$ turns and is represented by two circular filaments $C$ and $D$, then the mutual inductance between the coils can be approximated as

$$M \approx N_T N_R\left(\frac{M_{AC} + M_{AD} + M_{BC} + M_{BD}}{4}\right) \qquad (11)$$

where mutual inductance between pairs of filaments is calculated using (9) and (10).

To estimate the self-inductance of a planar spiral coil with $N$ turns, we use the following semi-empirical equation, which has been shown to give results within 5% of field solvers [5]:

$$L \approx \frac{1}{2}\mu_0 N^2 d_{avg}\left[\ln\left(\frac{2.46}{\rho}\right) + 0.20\rho^2\right] \qquad (12)$$

where $\rho \equiv (d_{out} - d_{in})/(d_{out} + d_{in})$. The coupling coefficient $k$ between two magnetically linked coils is given by

$$k = \frac{M}{\sqrt{L_T L_R}}\,. \qquad (13)$$

Since $M$ is a function of $N_T N_R$, $L_T$ is a function of $N_T^2$, and $L_R$ is a function of $N_R^2$, $k$ is independent of the number of turns in each coil. Thus, $k$ is determined only by the gross geometry of each coil and the distance between the coils. Using equations (9)–(13), we typically obtain coupling coefficients within 5% of those obtained using a field solver (FastHenry2) with a computational speed-up of better than $10^5$.

Biomedical applications place an upper limit on the outer diameter of the receive coil ($d_{outR}$), which is typically implanted in the body, and a lower limit on the coil-to-coil spacing ($z$) since the transmit coil is located outside the body. For a particular ratio of $z/d_{outR}$ it is possible to find values of $d_{outT}$, $d_{inT}$, and $d_{inR}/d_{outR}$ that maximize $k$. Using equations (9)–(13), we simultaneously varied these three parameters for values of $z/d_{outR}$ from zero to four and identified values that maximized $k$. Figs. 3(a)-(c) show the optimum values for each parameter as a function of $z/d_{outR}$. Dotted lines show the range of parameter values that resulted in a coupling coefficient of $0.9k_{max}$ or greater.

From Fig. 3(a), it is clear that as the spacing between coils increases, the transmit coil should be made larger. Figs. 3(b) and 3(c) show that for values of $z/d_{outR}$ greater than one, coil coupling can be optimized by making $d_{inT} \approx 0.18d_{outT}$ and $d_{inR} \approx 0.75d_{outR}$. However, the dotted lines show that some variation in inner diameter will not have a large effect on $k$. The maximum value of $k$ attained is shown in Fig. 3(d), and drops off as $(z/d_{outR})^{3/2}$ if the coils are properly sized for each value of $z/d_{outR}$. Therefore, the maximum achievable transmit power efficiency $\eta_T$ given in (4) falls off as $(z/d_{outR})^3$ for all but very closely spaced coils (see Fig. 4).

## III. Experimental Results

To validate the design equations presented above, we built and tested an inductive power link that uses a class E amplifier to deliver power to a 1.0-cm receive coil at a distance of 1.5 cm ($z/d_{outR}$ = 1.5). This closely models the coil size and spacing we expect in a cortical recording system under development [1]. We operate the link in an FCC-approved ISM (Industrial, Scientific, Medical) band at 6.78 MHz.

The receive coil was hand wound from 30AWG wire with $d_{outR}$ = 1.0 cm, $d_{inR}$ = 0.5 cm, and $N_R$ = 8 turns. Our chosen value of $d_{inR} = 0.5d_{outR}$ is slightly lower than optimal, but allows us to achieve a higher inductance and higher $Q_R$. At 6.78 MHz, the receive coil has an inductance $L_R$ = 654 nH with a series resistance of 0.883 Ω (measured with an Agilent 4285A Precision LCR Meter), resulting in $Q_R$ = 32.

Using the charts from Fig. 2, we designed the transmit coil to have $d_{outT}$ = 5.2 cm and $d_{inT}$ = 1.0 cm, which is close to optimal for a spacing of 1.5 cm. The coil was fabricated on a printed circuit board with 1-oz copper traces and $N_T$ = 24 turns. At 6.78 MHz, the transmit coil has an inductance $L_T$ = 16.96 μH with a series resistance of 4.73 Ω, resulting in $Q_T$ = 153. A coupling coefficient of 0.036 is predicted from (9)–(13), and we measured $k$ = 0.040 experimentally.

For demonstration purposes, we designed the system to deliver 50 mW to a 200-Ω load, giving $Q'_R$ = 7. The resistance in the transmit circuit $R_T$ is given by

$$R_T = R_{coil} + R_{C2} + 0.212R_{C1} + 1.365R_{on} + R_{limit} \quad (13)$$

where $R_{coil}$ is the series resistance of the transmit coil (measured at $\omega_0$), $R_{C1}$ and $R_{C2}$ are the equivalent series resistances of the class E capacitors $C_1$ and $C_2$, and $R_{on}$ is the "on" resistance of the class E MOSFET switch [7-8]. The $R_{\text{limit}}$ term represents an optional resistor that may be added to limit $Q_T$ to reduce sensitivity to component variation (at the expense of power efficiency). In our transmit circuit, $R_{C2}$ = 1.5 Ω, $R_{C1}$ = 0.4 Ω, $R_{\text{on}}$ = 3.4 Ω, and $R_{\text{limit}}$ = 17.9 Ω, giving $R_T$ = 28.9 Ω, which set $Q_T$ to 25 and $Q'_T$ to approximately 21.

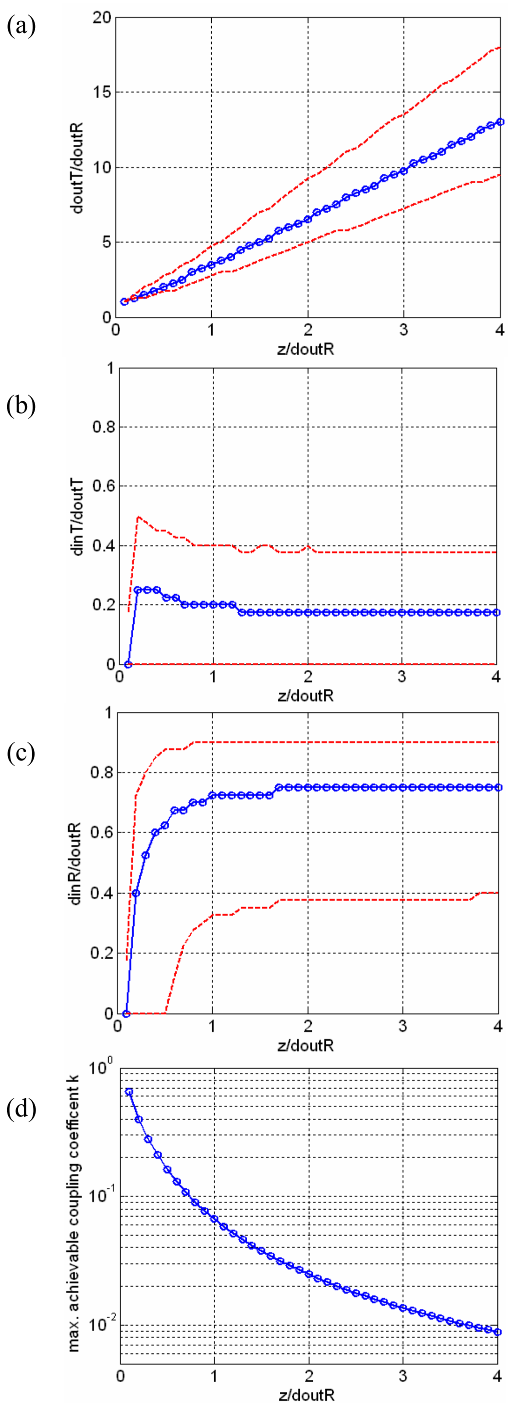


Figure 3. Optimum geometric values for planar coil coupling as a function of spacing ($z$) to receive-coil-outer-diameter ($d_{outR}$) ratio. (a) Optimum transmit coil outer diameter. (b) Optimum transmit coil inner diameter. (c) Optimum receive coil inner diameter. (d) Maximum achievable coupling coefficient $k$.

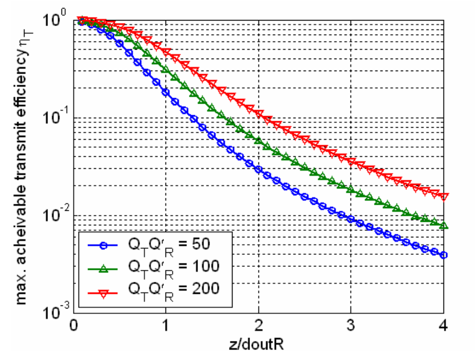


Figure 4. Maximum acheivable transmit efficiency $\eta_T$ as a function of spaing to receive-coil-diameter ratio, for different values of $Q_TQ'_R$.

Given the required load voltage and power, the receive coil should be designed to give $L_R = V_{R\text{pk}}^2/(2Q'_RP_L\omega_0)$ with $Q_R >> Q'_R$. The required class E voltage supply level is given by (7). The transmit coil voltage level can be predicted from (6).

We successfully delivered 50 mW to the load when the class E supply voltage was set to 4.85 V, which produced a transmit coil voltage of 62 $V_{\text{rms}}$. During operation, 53.5 mA of current was drawn from the supply, resulting in $P_S$ = 259 mW. This resulted in a measured efficiency of $\eta$ = 0.193, with $\eta_{\text{PA}}$ = 0.95, $\eta_T$ = 0.22, and $\eta_R$ = 0.82. A single-chip oscillator running from a 5-V supply was used to drive the gate of the class E MOSFET. This circuit consumed an additional 70 mW of power. When this is added to efficiency calculations, the overall system efficiency drops to $\eta$ = 0.152, with $\eta_{\text{PA}}$ = 0.75. Since implantable devices will consume much less than 100 mW, efficiencies in the 10-20% range are acceptable.

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