RF Powering of Millimeter- and Submillimeter-Sized Neural Prosthetic Implants

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Abstract—The size of the transducers for neural stimulation has shrunk steadily with application of thin film techniques to electrode design. In this paper, we examine the feasibility of designing millimeter- and submillimeter-sized power sources based on RF coupling that could be integrated into these new implants to provide power without a tethering power cable. The coupling between a transmitter coil and receiver coil when the coil diameters are markedly different is analyzed, and for these circumstances, a simple Thevenin equivalent model is developed to describe the power transmission between the transmitter and receiver. The equivalent circuit developed gives insight into the way that coil diameters, frequency, and turns affect coupling between large and small coils. Several examples demonstrate that milliwatt range power sources can be implemented with millimeter and submillimeter diameter receivers.

Introduction

There have been several analyses of transformer links for coupling power and information through the skin. Most recently, Soma *et al.* have given a thorough analysis of RF coupling for a variety of misalignment cases [3]. This paper examines the special case of transformer coupling to millimeter-sized or submillimeter-sized receiver coils by larger centimeter- and decimeter-sized transmitter coils. For this case, straightforward assumptions lead to a relatively simple expression for the transformer coupling. This expression is used to demonstrate that useful amounts of power can be coupled to millimeter and submillimeter coils for powering untethered sensor and stimulator implants. Finally, potential applications of implantable, untethered, integrated sensors and stimulators are briefly considered.

$$\begin{bmatrix} V_t \\ 0 \end{bmatrix} = \begin{bmatrix} \left(R_t + L_t jw + \dfrac{1}{C_t jw}\right) & -Mjw \\ -Mjw & \left(R_r + L_r jw + \dfrac{1}{C_r jw}\right) \end{bmatrix} \begin{bmatrix} I_t \\ I_r \end{bmatrix}. \tag{1}$$

Analysis

A loop of wire carrying a time-varying current produces time-varying magnetic and electric fields in the space around the wire. For the coil sizes and frequencies of interest here, the radiated energy is small and most of the energy in the field is stored energy. Under these conditions, the coupling of energy can be analyzed using lumped circuit theory.

Fig. 1 shows an inductively coupled transmitter–receiver circuit. The circuit is driven by a voltage source *Vt* that is sinusoidal with frequency *Wo*. *Lt* represents the inductance of the transmitter coil and *Ct* is chosen to be resonant with *Lt* at the transmitter driving frequency *Wo*. The resistance *Rt* represents the resistance in the inductor winding and other losses in the transmitter circuit including radiated energy and energy absorbed by the tissue. *Vt* represents the RF voltage supply to the transmitter. *Lr* and *Cr* are the inductor–capacitor pair that will be implanted and function as the receiver. *Rr* represents the resistive losses in the receiver coil circuit which, as in the case of the transmitter coil, consist of resistive losses in the coil and of radiated losses and tissue absorption losses. The receiver voltage *Vr* is the open circuit voltage across the capacitor *Cr*. This open circuit voltage *Vr* and an associated Thevenin equivalent impedance will characterize the behavior of the receiver under varying loads.

Maximum coupling between two coils of fixed geometry is achieved when both are tuned to resonate at the operating frequency. In the analysis that follows, we first determine the magnitude of *Vr* as a function of the transmitter voltage *Vt* at resonance. Then we determine the equivalent impedance associated with this receiver voltage. The open circuit voltage and source impedance give a complete Thevenin equivalent for the receiver power supply operated at resonance.

The loop equations for this circuit are given in (1)

At resonance, the impedance of the capacitor is the same magnitude but 180° out of phase from the impedance of the inductor, so these impedances cancel ($jwL + (1/jwC) = 0$). At resonance using sinusoidal sources, with the cancellation noted above, (1) reduces to

$$\begin{bmatrix} V_t \\ 0 \end{bmatrix} = \begin{bmatrix} R_t & -Mjw \\ -Mjw & R_r \end{bmatrix} \begin{bmatrix} I_t \\ I_r \end{bmatrix} \tag{2}$$

where $V = V(jw)$, $I = I(jw)$, $w = Wr$.

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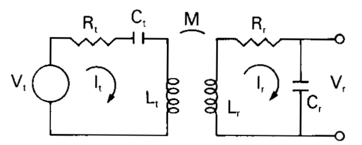

Fig. 1. Circuit representation of an RF coupled transmitter and receiver.

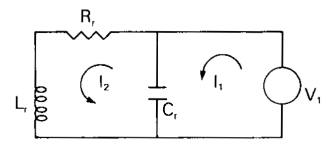

Fig. 2. Circuit to determine the equivalent impedance of the receiver power supply.

The determinant of the impedance matrix is

$$\Delta = R_t R_r + M^2 w^2. \tag{3}$$

For weakly coupled circuits, $R_t R_r >> M^2 w^2$ and the determinant is approximated by

$$\hat{\Delta} = R_t R_r. \tag{4}$$

(In the cases that follow, for example, the $R_t R_r$ product is 20–15 000 times greater than $M^2 w^2$.)

Then by matrix inversion,

$$\begin{bmatrix} I_t \\ I_r \end{bmatrix} = \frac{1}{\hat{\Delta}} \begin{bmatrix} R_r & Mjw \\ Mjw & R_t \end{bmatrix} \begin{bmatrix} V_t \\ 0 \end{bmatrix} \tag{5}$$

and by substitution,

$$V_r = I_r \cdot \frac{1}{C_r \cdot jw} = \frac{V_t Mjw}{R_t \cdot R_r \cdot C_r \cdot jw} \tag{6}$$

and

$$|V_r/V_t| = \frac{M}{R_t R_r C_r}. \tag{7}$$

Let Xl and Xc be the magnitude of the inductive and capacitive reactance (Lw and $1/Cw$) at resonance, respectively. Let Qt and Qr be defined as the ratio of the magnitude of the inductive or capacitive reactance to the resistance in the transmitter and receiver circuits, respectively. Then

$$|V_r/V_t| = X_M X_{Cr}/R_t R_r \tag{8}$$

and

$$|V_r/V_t| = \frac{M}{L_t} Q_t Q_r. \tag{9}$$

Equation (9) can be used to estimate receiver open circuit voltage Vr as a function of transmitter voltage Vt for coupled circuits with knowledge of the mutual inductance, transmitter coil inductance, transmitter coil Q, and receiver coil Q.

The equivalent impedance of the receiver power supply is determined by applying a voltage source $V1(jw)$ across the resonant capacitor Cr in Fig. 1 and determining the ratio of voltage to current when Vt is zero. For weakly coupled coils, however, the impedance coupled into the transmitter circuit does not contribute significantly to the receiver impedance and can be ignored to a first approximation. The circuit then simplifies as shown in Fig. 2.

Loop equations at resonance using the inductance–capacitance cancellation as in (2) give

$$\begin{bmatrix} V_1 \\ 0 \end{bmatrix} = \begin{bmatrix} \frac{1}{C_r jw} & -\frac{1}{C_r jw} \\ -\frac{1}{C_r jw} & \left(R_r + L_r jw + \frac{1}{C_r jw}\right) \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}.$$

$$\begin{bmatrix} V_1 \\ 0 \end{bmatrix} = \begin{bmatrix} \frac{1}{C_r jw} & -\frac{1}{C_r jw} \\ -\frac{1}{C_r jw} & R_r \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}. \tag{10}$$

The determinant is

$$\Delta = \frac{R_r}{C_r jw} - \frac{1}{(C_r jw)^2}.$$

Substituting the receiver capacitor reactance X_{Cr} and receiver Q, Qr gives

$$\Delta = \frac{1}{C_r w}\left(\frac{1}{C_r w} - jR_r\right)$$

$$\Delta = X_{Cr}(X_{Cr} - jX_{Cr}/Q). \tag{11}$$

For a tuned circuit with a moderate Q, the imaginary part of the determinant is small (for example $Q > 20$ implies the magnitude of the imaginary term is less than one twentieth of the magnitude of the real term) and the determinant is approximated as

$$\hat{\Delta} = X_{Cr}^2. \tag{12}$$

Inversion of the matrix in (10) using the determinant approximation of (12) and solving for $V1/I1$ gives

$$R_{eq} = V_1/I_1 = \frac{X_{Cr}^2}{R_r} = X_{Cr} : Q_r. \tag{13}$$

Equations (9) and (13) together give a complete description of the receiver power supply for our special case of weakly coupled, moderate Q circuits at resonance. The values of Qt, Qr, Lt, and XCr are easily calculated or measured.

An estimate of M, the mutual inductance, can be obtained from the following analysis. Assume two coplanar, coaxial coils with radii Yt and Yr (with $Yt >> Yr$) and turns Nt and Nr. This corresponds to a small diameter receiver coil being excited by a large diameter transmitter coil. At the center of the coils, a current It through the transmitter coil produces a magnetic field and flux given by

$$H_z = I_t N_t / 2Y_t \tag{14}$$

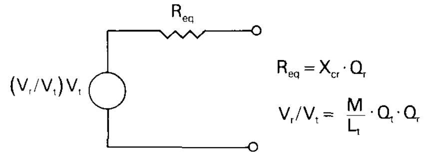

Fig. 3. Thevenin equivalent model for the receiver power supply in terms of the transmitter and receiver Q's, the transmitter and receiver inductances, the mutual inductance, and the transmitter supply voltage.

and

$$B_z = I_t N_t \mu / 2Y_t. \quad (15)$$

If we assume that H is relatively uniform over the small area of the receiver coil, then the flux linkage is

$$\hat{\phi} = \frac{N_t N_r I_t \mu}{2Y_t} \cdot \pi Y_r^2 \quad (16)$$

For the receiver coil, the voltage across the inductor Vl is

$$V_l = -\frac{\partial \phi}{\partial t} = \frac{N_t N_r \pi Y_r^2 \mu}{2Y_t} \cdot \frac{\partial I_t}{\partial t}. \quad (17)$$

But this is the expression for mutual inductance so for coaxial coplanar coils

$$\hat{M} = \frac{\pi}{2} N_t N_r \mu \frac{Y_r^2}{Y_t}. \quad (18)$$

The expression for mutual inductance as a function of coil plane and axial misalignment of the transmitter and receiver coils can be obtained by using a more general expression for the magnetic flux density in place of (15). This expression for the magnetic flux, normal to the plane of the transmitter coil, at the receiver due to a current It thru a loop of Nt turns at the transmitter is given by [4]:

$$B_z = \frac{1}{\left((Y_r + r)^2 + z^2\right)^{1/2}} \cdot \frac{\mu_0 I_t N_t}{2\pi} \cdot \left[K + \frac{(Y_r^2 - r^2 - z^2)}{(Y_r - r)^2 + z^2} \cdot E\right] \quad (19)$$

where z represents the displacement between the transmitter and receiver coil planes and r is the axial misalignment of the receiver coil. $K(k)$ and $E(k)$ are complete elliptic integrals of the first and second kind of the parameter k where k is given by

$$k = \left[4 \cdot Y_r \cdot r / \left((Y_r + r)^2 + z^2\right)\right]^{1/2}. \quad (20)$$

If a ferrite core is added to the small coil, it has the effect of increasing the self-inductance and the mutual inductance by a variable amount and can be represented by including an appropriate factor μ_f into (18). The magnitude of this factor depends on the relative length and cross section of the ferrite rod and typical values range from 4 to 15 [8].

The Thevenin equivalent model for the transmitter–receiver is given in Fig. 3.

Results

Five examples will now be examined that demonstrate the feasibility of powering millimeter- and submillimeter-sized devices. Case 1 assumes a transmitter coil of 90 mm diameter and receiver coils of 1.5 mm in diameter. This might correspond to powering several small implants in the forearm with a bracelet-type transmitter coil. Case 2 assumes the same sized transmitter and receiver as in case 1, but assumes that a ferrite core has been added to the receiver that increases the mutual inductance and the self-inductance of the receiver coil by 5. Case 3 assumes a transmitter coil of 320 mm and a receiver coil of 2 mm as might be used for spinal cord or abdominal devices with a belt-like transmitter coil. Case 4 assumes a transmitter coil of 140 mm diameter and receiver coils of 0.4 mm diameter that could be used for cortical stimulation with a single external transmitter coil and multiple implanted receivers. Case 5 is identical to case 4 but assumes a fivefold increase in receiver coil inductance and mutual inductance is achieved using a ferrite core in the receiver coil.

In each case, the inductance of the transmitter coil was chosen to provide a reactive impedance at the operating frequency of 500 Ω. The number of turns listed for the transmitter coil is the nominal number rounded upward to provide the desired reactance based on standard design tables [7] and assuming a diameter-to-length ratio of 10. The Q of the transmitter and receiver is assumed to be 40 in each case. Receiver coil inductances are calculated using a design formula [6] useful for diameter-to-length ratios of less than 2.5. The receiver coil is assumed to have 60 turns in each case. This corresponds to a 3 mm coil length if 50 μm spacing is assumed between successive turns of a single layer coil. The final column of Table I lists the rms power (calculated as the square of the rms output voltage divided by four times the equivalent impedance) that could be delivered to a matched load at at the receiver assuming that the transmitter voltage is 10 V peak. The 10 V peak driving voltage corresponds to an rms power delivered to the transmitter of 4 W for a Q of 40. The results are presented for each case for frequencies of 2 and 20 MHz.

Discussion

The results tabulated in Table I and in the equivalent circuit of Fig. 3 lead to several generalizations for millimeter- and submillimeter-sized receivers.

Most importantly, it is possible to power millimeter-sized and submillimeter-sized receivers using large diameter transmitter coils. Assuming active regulation of the power supply at the implanted device, variation in coupling is not important as long as the mutual inductance remains above a minimum level. If the receiver coil is within the circumference of the transmitter coil, the mutual inductance is minimal when the coils are nearly coaxial and increases as the coils are moved significantly from the axial alignment. This has been seen experimentally with test coils and is also shown by evaluation of the

TABLE I
SUMMARY OF THE POWER SUPPLY CHARACTERISTICS FOR FIVE CASES OF TRANSMITTER AND RECEIVER COILS. SEE TEXT FOR DETAILS.

case		Transmitter			Receiver					Power Supply		
	freq. Mhz	dia mm.	turns	L μH	dia mm.	len mm.	L μH	Xcr ohm	M nH	Vr/Vt	Req K ohm	power mw
1a	2	90	14	40	1.5	3	2.17	27	21	.83	1.1	7.9
1b	20	90	5	4	1.5	3	2.17	270	7.4	3	11	10
2a	2	90	14	40	1.5	3	10.85	135	104	4.1	5.4	39
2b	20	90	5	4	1.5	3	10.85	1350	37	14.8	54	50
3a	2	320	8	40	2.	3	3.64	45	8	.3	1.8	0.7
3b	20	320	3	4	2.	3	3.64	450	3	1.2	18	0.96
4a	2	140	11	40	0.4	3	.178	2.24	.54	.02	0.09	0.06
4b	20	140	4	4	0.4	3	.178	22.4	.2	.08	0.9	0.09
5a	2	140	11	40	0.4	3	.89	11.2	2.7	.11	0.45	0.32
5b	20	140	4	4	0.4	3	.89	112	1	.41	4.5	0.46

expression for B in (19) for different axial misalignments. Thus, the receiver will operate anywhere within the transmitter circumference with performance comparable or better than the performance suggested in Table I. Performance dropoff associated with z axis displacement between the transmitter and receiver coils has not been examined here but can be evaluated using (19). Thus, for example, the spatial region of operation for the receiver coil where the mutual inductance is at least 50 percent of the value given in (18) can be calculated using (19).

The voltage transfer ratio is seen to depend on few variables. It varies with the square of the receiver coil diameter and varies inversely with the transmitter coil diameter. It is proportional to the number of turns in the receiver coil and also varies directly with the increase in the effective permeability associated with ferrite cores in the receiver coil. It is proportional to the product of the transmitter and receiver Q's.

The equivalent impedance of the receiver power supply is directly proportional to the Q of the receiver circuit and to the inductive or capacitive reactance of the receiver. Because reactance is proportional to frequency, the impedance is also proportional to frequency for a given receiver coil. The frequency dependence of the receiver reactance gives insight into the selection of an operating frequency. The operating frequency must be high enough that the reactance of the receiver coil is at least 40 times greater than the resistance of the receiver coil if a Q of 40 is to be realized.

The variation of the equivalent circuit parameters at 2 and 20 MHz assumes that the Q's of the transmitter and receiver can be maintained at the frequencies of interest. The loaded Q of the transmitter coil in an application depends on the resistive coil losses and on power radiation by the transmitter coil but also depends on losses associated with power absorption by tissue and power absorption by the receiver coil(s). This technique, as well as safety considerations, requires that the tissue not absorb significant RF energy. Considering tissue losses, it is known that absorption varies approximately as the square of frequency in the range of 1–100 MHz [5]. Thus, for a given energy density within the transmitter coil and for a specified maximum power absorption by the tissue, there will be a high-frequency cutoff above which the power absorbed by tissue will be excessive. Tests done at frequencies up to 3 MHz using a 90 mm transmitter coil surrounding a water bath or operating in free air produced Q's well over 40. The Q of a 1.5 mm diameter ferrite core receiver coil operating in air and submerged in water after a dip coating of wax also exceeded 40. The loaded Q of the transmitter can also be reduced if numerous receiver coils are implanted. The loss of Q from this mechanism becomes significant only if the power extracted by the receiver coils is a significant fraction of the power consumed by the transmitter coil. Taking case 5a) as an extreme example, over 600 receivers could be implanted before the power used by the receivers totaled 10 percent of the power used to drive the unloaded transmitter coil. A final practical consideration for maintaining high Q is the requirement that the coils be tuned to resonance at the operating frequency. The inductance of the primary coil changes with small changes in its geometry and local environment. To maintain operation at the resonant frequency, the resonant capacitor may require active adjustment to compensate for these variations.

Bowman and Meindl, discussing the problems of making connections to implanted sensors state, "To date though, no durable and noncorrodible connection to implantable integrated sensors has been demonstrated, and the successful application of integrated sensors to implanted telemetry and closed-loop control awaits the solution of this particular micropackaging problem" [2]. Integrated power supplies may provide one approach to the packaging problem by eliminating external cables.

This analysis is based on minimizing the size of the receiver coil. As a result, other parameters such as efficiency have suffered. A tradeoff may be required between size and efficiency if these implants are to be used with portable battery powered transmitters. However, if relatively high transmitter power consumption can be tolerated, this analysis suggests that receiver–stimulators can be designed which can be implanted through a 14 gauge (1.6 mm i.d.) hypodermic needle.

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