Study on Magnetically Coupled Intrabody Communication Dynamic Optimization Regarding Two-Coil Power-Transfer System Using L-section Matching Network

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Received: March 29, 2019; Accepted: August 8, 2019; Published: August 31, 2019

**Abstract.** This study examined a novel intrabody communication method using magnetic coupling. For designing transceivers utilizing this communication method, it is important to understand the power-transfer characteristic between the sending and receiving coils of the transceivers and to develop a technique for maximizing the transceiver efficiency. This study targeted a two-coil wireless power-transfer system with an L-section matching network. The computer simulation results showed that 1) the maximum power-transfer efficiency (PTE) was obtained by adjusting the matching network, and 2) the optimal PTE was achievable. In addition, this study conducted experiments and derived similar results from part of the simulation. The proposed wireless power-transfer system is promising as a magnetic intrabody communication system.

**Keywords:** Intrabody Communication, Magnetic Coupling, Matching Network

1. Introduction

Intrabody communication is a new communication method that employs the human body as a transfer medium for an electric signal. Study of intrabody communication began to draw attention after T. G. Zimmerman of IBM published a paper entitled “Personal area networks: Near-field intrabody communication” in 1996 [1]. Thus far, various examinations, such as communication-mechanism clarification and application development, regarding intrabody communication have been conducted [2, 3]. Intrabody communication is one of the promising communication methods for constructing a body area network and has advantages over other
wireless systems regarding communication efficiency and confidentiality, as the human body and its surrounding environment are the main transfer medium.

Generally, intrabody communication is classified into two coupling methods: galvanic coupling and capacitive coupling [2, 3]. However, the communication quality of these two methods depends strongly on the surrounding environment. Recently, the magnetic-coupling method, which has less influence of the external environment, was proposed [4–7]. The signal-loss characteristic of this method depends extremely on the coupling method employed. The author has analyzed various proposed magnetic-coupling methods and examined the wireless power-transfer field in order to maximize the communication distance and minimize the signal loss [8]. The examined coupling methods were general inductive coupling (non-resonant coupling) and two kinds of resonant couplings [9]. A three-dimensional (3D) finite-element model incorporating circuit elements was used for the analysis. The frequency-characteristic results of the signal-transfer loss showed that the optimum frequency band of the inductive coupling was 2–3 MHz. Additionally, one of the resonant couplings was able to improve the signal loss by approximately 20 dB. This resonant coupling can be improved by 20 dB without specially adjusting the parameters even if the distance between the transmitter and receiver is changed, that is, the mutual inductance between the sending and receiving coils is altered. Therefore, the resonant coupling is considered to be fairly effective for extending the communication distance. However, this coupling method may not minimize the signal loss.

This study aims at realizing the system giving the minimum loss in the case where the two coils of the transceivers are dynamically positionally changed. For a wireless power-transfer system using two coils, where the self-inductance and parasitic resistance of the coils and mutual inductance between the coils are given, a method for configuring the matching network that gives the optimum transducer gain was proposed [10]. In pure wireless power transfer, the optimum transfer system can be realized by simply applying this method, as the two coils can be installed in a known state; that is, the mutual inductance between them can be obtained in advance. However, in intrabody communication, because the positional relation between the two coils is typically changed, the mutual inductance between them also varies. If the mutual inductance can be estimated in real time, while dynamically altering the parameters of the matching network, the transfer system that gives the optimum transducer gain at the present positional relation can be configured [11, 12]. The references 11 and 12 showed the feasibility of such a system through a simulation. Besides a simulation, through an experiment, this study investigates whether this type of system can be configured.

Chapter 2 outlines the two-coil wireless transfer system using a matching network. Chapter 3 describes the calculation formulae for the transducer power gain and matching parameters of the system incorporating the L-section matching network. Chapter 4 and 5 present the simulation and experimental methods and results, respectively, followed by Chapter
2. Two-coil wireless power-transfer system using matching network

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Figure 1 shows a two-coil wireless power-transfer system with an input–output matching network yielding the maximum power-transfer efficiency (PTE). $V_s$ in the figure represents the signal source voltage, and $Z_s$ and $Z_L$ represent the signal source and load impedances, respectively. If the optimum source and load conditions are satisfied, the wireless power-transfer system can achieve the ideal maximum PTE; the matching network transforms the signal source and load impedances into the optimum ones. The PTE can be estimated using the transducer power gain and calculated as described below [10].

The power input into the matching network on the signal source side can be computed as follows:

$$P_{in} = \frac{1}{2} |I_1|^2 \text{Re}(Z_{in}) = \frac{1}{2} \frac{V_s}{Z_{in} + Z_s}^2 \text{Re}(Z_{in}),$$

(1)

where $Z_{in}$ represents the input impedance to the matching network on the signal source side. On the other hand, the actual power input to the load is computed as follows:

$$P_L = \frac{1}{2} |I_2|^2 \text{Re}(Z_L).$$

(2)

The transducer power gain $G_T$ is thus expressed as

$$G_T = \frac{P_L}{P_{av}} = \frac{P_L}{P_{in}|_{Z_o=Z_s}},$$

(3)

where $P_{av}$ represents the available power of the source and becomes equal to $P_{in}$ when the network is conjugate-matched to the source. Therefore, the available power can be computed as follows:

$$P_{av} = \frac{1}{8} \frac{|V_s|^2}{\text{Re}(Z_s)}.$$  

(4)

As the transducer power gain is defined as the ratio between the power supplied to the load and the available power of the signal source, it can be used to evaluate the PTE under the condition with the signal source and load. As shown in Fig. 1, when the self-inductances $L_1$ and $L_2$, the mutual inductance $M$, and the parasitic resistances $R_1$ and $R_2$ for the two coils are given, the optimum signal source and load conditions for $Z_{s,\text{opt}}$ and $Z_{L,\text{opt}}$ maximizing the transducer power gain are given by (5) and (6), respectively [10].

$$Z_{s,\text{opt}} = R_1 \sqrt{1 + \frac{\omega^2 M^2}{R_1 R_2} - j\omega L_1}$$

(5)
\[ Z_{L,\text{opt}} = R_2 \sqrt{1 + \frac{\omega^2 M^2}{R_1 R_2}} - j \omega L_2, \tag{6} \]

where \( \omega \) denotes the angular frequency.

\[ Z_{\text{in}} = Z_s^* = R_1 \sqrt{1 + \frac{\omega^2 M^2}{R_1 R_2}} - j \omega L_1 \]
\[ Z_{\text{out}} = Z_L^* = R_2 \sqrt{1 + \frac{\omega^2 M^2}{R_1 R_2}} - j \omega L_2 \]

Figure 1: Two-coil wireless power-transfer system with an input–output matching network satisfying the maximum-PTE condition.

3. Two-coil wireless power-transfer system using L-section matching network

Matching networks can be configured in various ways, such as transformers, lumped elements, and transmission lines. This study utilized an L-section matching network employing lumped elements. Figure 2 depicts the two-coil wireless power-transfer system with an L-section matching network.

3.1. Derivation of transducer power gain \( G_T \) and load voltage \( V_2 \)

In this section, the transducer power gain \( G_T \) and load voltage \( V_2 \) on the system are derived.

The \( Z \)-parameters of the inductive link can be determined as follows:

\[ Z_{11} = R_1 + j \omega L_1 \tag{7} \]
\[ Z_{22} = R_2 + j \omega L_2 \tag{8} \]
\[ Z_{12} = Z_{21} = j \omega M. \tag{9} \]

Using the \( Z \)-parameters, the output impedance \( Z_{\text{out}} \) can be expressed as

\[ Z_{\text{out}} = Z_{C_{2R}} + \frac{Z_{C_{2p}} \times Z_{\text{out}_c}}{Z_{C_{2p}} + Z_{\text{out}_c}}, \tag{10} \]

where \( Z_{C_{2s}} \) and \( Z_{C_{2p}} \) represent the impedances for the capacitors \( C_{2s} \) and \( C_{2p} \), respectively. \( Z_{\text{out}_c} \) is the output impedance looking into the inductive link from the matching net-
work of the load side and is given as

\[ Z_{\text{out}_c} = Z_{22} - \frac{Z_{12}^2}{Z_{11} + Z_{s_\perp}}, \quad (11) \]

where \( Z_{s_\perp} \) is the source impedance including the matching network of the signal source side. 
\( Z_{s_\perp} \) is given as

\[ Z_{s_\perp} = \frac{Z_{C1p} \times (Z_{C1s} + Z_s)}{Z_{C1p} + (Z_{C21} + Z_s)}, \quad (12) \]

where \( Z_{C1s} \) and \( Z_{C1p} \) represent the impedances for the capacitors \( C_{1s} \) and \( C_{1p} \), respectively. The input impedance \( Z_{\text{in}} \) can be calculated in the same manner. The current flowing through the load, \( I_2 \), can be expressed as

\[ I_2 = \frac{Z_{C2p}Z_{12}Z_{C1p}V_s}{(Z_{CZZ} + Z_{C2p})(Z_{\text{out}_c} + Z_{L_\perp})(Z_{11} + Z_{C1p})(Z_s + Z_{CCRL1})}, \quad (13) \]

where \( Z_{L_\perp} \) is the load impedance including the matching network of the load side; \( Z_{CZZ} \) is the synthetic impedance of the capacitor \( C_{2s} \) and the load impedance \( Z_L \); and \( Z_{CCRL1} \) is the synthetic impedance of the signal source side, excluding the signal source impedance. Thus, the load voltage \( V_2 \) can be calculated as

\[ V_2 = I_2 \cdot Z_{L_\perp}. \quad (14) \]

The transducer power gain \( G_T \) can be computed by substituting (2), (4), and (13) into (3).

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3.2. Computation of matching parameters

As described in Section 2, the optimum signal source and load conditions are given by (5) and (6), respectively. The matching parameters—the series and parallel capacitors \( C_s \) and \( C_p \)—are adjusted so that the signal source and load impedances including the matching network, \( Z_{s_\perp} \) or \( Z_{L_\perp} \), are equal to \( Z_{s_{\text{opt}}} \) or \( Z_{L_{\text{opt}}} \), respectively. The exact values of \( C_s \) and \( C_p \) can be determined using the analytical equation of the L-section matching network [13].

Figure 3 shows a schematic of the L-section matching network of the load side. The
terms $X$ and $B$ represent the reactance and susceptance of the matching network, respectively. The impedance-matching condition is as follows:

$$\frac{1}{R_{L,\text{opt}} + jX_{L,\text{opt}}} = jB + \frac{1}{R_L + jX}$$

(15)

where $Z_{L,\text{opt}} = R_{L,\text{opt}} + jX_{L,\text{opt}}$ and $Z_L = R_L + jX = R_L + j0$. By separating (15) into two equations for the real and imaginary parts and solving these for $X$ and $B$, respectively, the following is obtained:

$$X = \pm \sqrt{\frac{R_{L,\text{opt}}R_L(R_{L,\text{opt}}^2 - R_{L,\text{opt}}R_L + X_{L,\text{opt}}^2)}{R_{L,\text{opt}}}}$$

(16)

$$B = \frac{-R_LX_{L,\text{opt}} \pm \sqrt{R_{L,\text{opt}}R_L(R_{L,\text{opt}}^2 - R_{L,\text{opt}}R_L + X_{L,\text{opt}}^2)}}{R_L(R_{L,\text{opt}}^2 + X_{L,\text{opt}}^2)}$$

(17)

Thus, although two combinations of solutions are obtained, because one combination $X$ is positive and can only be realized with coils, this combination can be ignored. The series and parallel capacitors $C_{2s}$ and $C_{2p}$ are calculated as follows:

$$C_{2s} = -\frac{1}{\omega X}$$

(18)

$$C_{2p} = \frac{B}{\omega}$$

(19)

The series and parallel capacitors $C_{1s}$ and $C_{1p}$ are calculated in the same manner.

![Figure 3: Schematic of the L-section matching network of the load side.](image)

4. **Computer simulation**

The network parameters for the transfer system with the L-section matching network (Fig. 2) are listed in Table 1. As the coupling coefficient is $k = 0.006$, the mutual inductance in this simulation is $M = k\sqrt{L_1L_2} = 14.7$ nH. The simulation program was developed and executed in MATLAB.
4.1. Computation of network parameters and efficiency

The matching parameters—the series and parallel capacitors $C_s$ and $C_p$—were computed under the conditions of Table 1, yielding $C_{1s} = C_{2s} = 0.161 \text{ nF}$ and $C_{1p} = C_{2p} = 2.43 \text{ nF}$. Next, using the above parameters, the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$ were calculated with a coupling coefficient $k$ of 0.00–0.03. Figures 4 and 5 depict the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$ with respect to the coupling coefficient $k$. As shown in Fig. 4, although the transducer power gain $G_T$ exhibits a relatively high efficiency of 0.57 when the coupling coefficient $k$ is 0.006, it is maximized (0.62) at $k = 0.008$. Thus, the coefficient value maximizing the transducer power gain $G_T$ deviated from the setting. As shown in Fig. 5, the magnitude of the load voltage was maximized (0.39 V) at the coupling coefficient of $k = 0.008$, whereas it was 0.38 V at $k = 0.006$.

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**Table 1: Network parameters.**

| Parameters | Value     | Description                                      |
|------------|-----------|--------------------------------------------------|
| $L_1, L_2$ | 2.45 μH   | Inductance of primary and secondary coil          |
| $R_1, R_2$ | 0.05 Ω    | Parasitic resistance of primary and secondary coil |
| $Z_s, Z_L$ | 50 Ω      | Source and load impedances                        |
| $k$        | 0.006     | Coupling coefficient                              |
| $f_0$      | 2 MHz     | Carrier frequency                                 |
| $V_s$      | 1 V       | Source voltage                                    |

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**Figure 4:** Transducer power gain $G_T$ with respect to the coupling coefficient $k$. 
Figure 5: Magnitude of the load voltage $|V_2|$ with respect to the coupling coefficient $k$.

4.2. Computation of efficiency at each coupling coefficient

In this section, the matching parameters $C_s$ and $C_p$ are calculated for coupling coefficients of $k = 0.001, 0.002, ..., 0.03$. Furthermore, the mutual inductance $M$ that yields the optimum transfer power gain $G_T$ at the corresponding matching parameters is estimated. Figure 6 shows the matching parameters $C_{1s}, C_{1p}, C_{2s},$ and $C_{2p}$ at each coupling coefficient. As the coupling coefficient increases, the parameters $C_{1s}$ and $C_{2s}$ continuously monotonically increase from 0.09 to 0.36 nF, while the parameters $C_{1p}$ and $C_{2p}$ continuously monotonically decrease from 2.5 to 2.24 nF. Figure 7 shows the mutual inductances at each coupling coefficient. The term “true M” represents the mutual inductance used for estimating the matching parameter, and the term “optimum M” represents the mutual inductance yielding the optimum transducer power gain $G_T$ at the corresponding matching parameters. For all the coupling coefficients, the “optimum M” is 4.5 nH greater than the “true M”. Additionally, the deviation appears to increase when the coupling coefficient is extremely small.

Figure 6: Matching parameters $C_{1s}, C_{1p}, C_{2s},$ and $C_{2p}$ with respect to the coupling coefficient $k$. 
4.3. Computation of network parameters and efficiency considering deviation

As described in the previous section, when the matching parameters calculated via (18) and (19) are used, the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$ are maximized not at the coupling coefficient of 0.006 (mutual inductance = 14.7 nH) but rather at the larger coupling coefficient of 0.008 (mutual inductance = 19.2 nH). Thus, in this section, the matching parameters $C_{1s}, C_{1p}, C_{2s}$, and $C_{2p}$ are calculated considering this deviation, yielding $C_{1s} = C_{2s} = 0.137$ nF and $C_{1p} = C_{2p} = 2.45$ nF. Figures 8 and 9 show the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$. As shown, the transducer power gain $G_T$ was almost maximized (0.53) at $k = 0.006$ and decreased with the increasing deviation of the set coupling coefficient. It was confirmed that the matching parameters were successfully obtained. Furthermore, as depicted in Fig. 9, the magnitude of the load voltage $|V_2|$ was also maximized at the coupling coefficient of 0.006. However, the gain $G_T$ and the load voltage magnitude $|V_2|$ at $k = 0.006$ were smaller than the ones of section 4.1.
Fig. 9: Magnitude of the load voltage $|V_2|$ considering the deviation.

### 4.4. Estimation of mutual inductance

This section confirms whether the mutual inductance corresponding to each coupling coefficient can be estimated when the coupling coefficient is 0.00–0.03 under the conditions of $C_{1s} = C_{2s} = 0.137$ nF and $C_{1p} = C_{2p} = 2.45$ nF. The magnitude of the load voltage $|V_2|$ was used for the estimation. As shown in Fig. 9, because the magnitude of the load voltage $|V_2|$ depends only on the coupling coefficient $k$, that is, the mutual inductance $M$, the mutual inductance $M$ can be estimated by measuring the magnitude $|V_2|$. The steps of the estimation are as follows: 1) the receiver alters $Z_{12}$ of the Z-parameters gradually and then calculates the load voltage $|V_2|$ using (14); 2) the receiver compares the calculated load voltage $|V_2|$ with the measured load voltage to estimate the mutual inductance using $M = Z_{12}/\omega$ from $Z_{12}$ with the least error. However, because there are two mutual inductances $M$ giving the same magnitude $|V_2|$, two estimates are obtained.

Figure 10 depicts the true and estimated value of the mutual inductance at each coupling coefficient. As shown in Fig. 10(a), as the coupling coefficient $k$ increases, the mutual inductance $M$ increases from 2.45 to 73.5 nH. Figure 10(b) shows the estimated values of the mutual inductance $M$ when there are no errors in the matching parameters. Thus, two estimated values are obtained, and it is confirmed that one of them represents the actual mutual inductance. This discrimination can be performed by setting the matching parameters to the values computed by respective estimated value of the mutual inductance $M$ and selecting the one with the larger voltage magnitude. Figure 10(c) shows the estimated values of the mutual inductance $M$ when there are uniform errors of <1% in the matching parameters. It was confirmed that the mutual inductance $M$ was satisfactorily estimated with practical precision, although the two estimated values could not be distinguished in the coupling-coefficient range of $k = 0.004–0.010$, because the two values were close.
Figure 10: True and estimated values of the mutual inductance with respect to the coupling coefficient $k$.
5. Experiment

To examine the validity of the simulation conditions, we investigated the relationship between the coupling coefficient $k$ and the distance between the primary and secondary coils. This is described in Section 5.1 that follows. Sections 5.2 and 5.3 explain how we confirmed that the L-section matching network could realize the maximum transducer power gain. For this, the maximum transducer power gain was calculated at each coupling coefficient $k$. In addition to conducting measurements, we also calculated the theoretical values.

The primary and secondary coils used in this experiment were composed of 0.65-mm heat-resistant PVC wire. For both coils, the radius was $r = 10.5$ cm and the number of turns was $N = 3$. The self-inductances $L_1$ and $L_2$ and the parasitic resistances $R_1$ and $R_2$ for the two coils were measured using an LCR meter (HIOKI IM3536). The primary coil was $L_1 = 2.67 \mu H$ and $R_1 = 0.06 \Omega$; the secondary coil was $L_2 = 2.65 \mu H$ and $R_2 = 0.06 \Omega$.

5.1. Relationship between the coupling coefficient $k$ and the distance between the two coils

The coupling coefficient $k$ at a distance of 1 to 30 cm between the primary and secondary coils was determined by theoretical calculations and measurements. The primary and secondary coils were arranged in parallel and concentrically (Fig. 11). The coupling coefficient $k$ was calculated using the Neumann formula [14, 15]. In the measurements, S-parameters were measured using a vector network analyzer. They were then converted to Z-parameters, and the coupling coefficient $k$ was calculated [16].

First, calculations using the Neumann formula are explained as follows. In general, mutual inductance between two circuits $C_s$ and $C_r$ is expressed by the following equation.

$$M = \frac{\mu_0}{4\pi} \int_{C_s} \int_{C_r} \frac{dl_s \cdot dl_r}{l_{sr}}$$

(20)

where $\mu_0$ is the vacuum permeability, $dl_s$ and $dl_r$ are minute elements on the circuit, and $l_{sr}$ is the distance between $dl_s$ and $dl_r$. The mutual inductance in the geometric positional relationship of the primary and secondary coils as shown Fig. 11 is expressed by the following equation.

$$M = N^2 \frac{\mu_0 r^2}{4\pi} \int_0^{2\pi} \frac{\cos(\phi_s - \phi_r) d\phi_s d\phi_r}{\sqrt{2r^2 \{1 - \cos(\phi_s - \phi_r)} + D_{sr}^2}}$$

(21)

where $D_{sr}$ is the distance between the two coils. The coupling coefficient is given by $k = M/\sqrt{L_s L_r}$.

Next, the conducted measurements are described as follows. The S-parameter matrix $S$ can be converted to the Z-parameter matrix $Z$ using the following equation [17].
where \( \mathbf{I} \) represents the identity matrix. The real and imaginary parts of \( \mathbf{Z} \)-parameter matrix \( \mathbf{Z} \) can be represented as follows.

\[
\mathbf{Z} = \begin{bmatrix}
Z_{11} & Z_{12} \\
Z_{21} & Z_{22}
\end{bmatrix} = \begin{bmatrix}
R_{11} & R_{12} \\
R_{21} & R_{22}
\end{bmatrix} + j \begin{bmatrix}
X_{11} & X_{12} \\
X_{21} & X_{22}
\end{bmatrix}
\]  
(23)

The coupling coefficient \( k \) is then obtained from the following equation using the imaginary part of the \( \mathbf{Z} \) parameter [15].

\[
k = \frac{|X_{21}|}{\sqrt{X_{11}X_{22}}}
\]  
(24)

For the coupling coefficient \( k \), Fig. 12 shows the theoretical values calculated using the Neumann formula and the obtained values derived from a vector network analyzer (KEYSIGHT E 5063A). As shown in Fig. 12, the theoretical and measured values were nearly identical. Section 4 describes simulations that were conducted with a coupling coefficient \( k = 0.002 - 0.03 \), which roughly corresponds to a coil distance of 10–30 cm \( (k = 0.002 - 0.03) \). In addition, the coupling coefficient \( k = 0.006 \) for which the matching parameter was calculated corresponds to the distance 18–19 cm between the two coils.

Figure 11: Positional relationship of the primary and secondary coils.
5.2. Maximum transducer power gain at each coupling coefficient \( k \)

The maximum transducer power gain \( G_{T,\text{max}} \) at the distance between the primary and secondary coils of 10–30 cm (i.e., corresponding to each coupling coefficient \( k = 0.002–0.03 \)) was obtained by theoretical calculations and measurements. The matching parameters \( C_{1s}, C_{1p}, C_{2s}, C_{2p} \) were calculated for each coupling coefficient \( k \), and the maximum transducer power gain \( G_{T,\text{max}} \) was determined using the theoretical calculation described in Section 3.

In the measurements, S-parameters were measured using the vector network analyzer and calculated using the following method [18]. Through the S parameter matrix \( S \),

\[
S = \begin{bmatrix}
S_{11} & S_{12} \\
S_{21} & S_{22}
\end{bmatrix}
\]  

(25)

the maximum transducer power gain \( G_{T,\text{max}} \) was obtained by the following equation:

\[
G_{T,\text{max}} = \frac{1}{1 - |\Gamma_S|^2} \cdot |S_{21}|^2 \cdot \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2}
\]  

(26)

where \( \Gamma_S \) and \( \Gamma_L \) are the reflection coefficients on the source and load sides, respectively, which can be calculated by the following equations:

\[
\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1}
\]  

(27a)

\[
\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2}
\]  

(27b)

The variables \( B_1, C_1, B_2, C_2 \) are defined by the following equations:

\[
B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2
\]  

(28a)
In addition, \( \Delta \) is given by:

\[
\Delta = S_{11}S_{22} - S_{12}S_{21}
\]  

(29)

Equation (26), in which the maximum transducer power gain \( G_{T,max} \) is given, can be rewritten as:

\[
G_{T,max} = \frac{|S_{21}|}{|S_{12}|} \left( K - \sqrt{K^2 - 1} \right)
\]  

(30)

where \( K \) is given by:

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

(31)

In the range of coupling coefficient \( k = 0.002 - 0.03 \), \( K > 1, |\Delta| < 1 \) for all coupling coefficients. In addition, the maximum transducer power gain \( G_{T,max} \) could be obtained. I also confirmed that the results calculated using (26) and (30) matched.

Figure 13 shows the theoretical and measured values of the maximum transducer power gain \( G_{T,max} \) at each coupling coefficient \( k \). As shown in Fig. 13, the theoretical and measured values coincided to some extent, but the measured value tended to be smaller. Under actual measurement conditions, the transducer power gain using the L-section matching network with coupling coefficient \( k = 0.006 \) was expected to be approximately 0.4–0.5.
5.3. Realization of maximum transducer power gain using L-section matching network

We constructed a two-coil transfer system using an L-section matching network. Through measurements, we then confirmed that the system could realize a transducer power gain $G_T$ similar to that described in the previous section. The coupling coefficient was set to $k = 0.006$ as in the previous calculation. The matching parameters (i.e., the series and parallel capacitors $C_s$ and $C_p$) were measured using the LCR meter, yielding $C_{1s} = C_{2s} = 0.160 \, \text{nF}$ and $C_{1p} = C_{2p} = 2.23 \, \text{nF}$. The distance between the primary and secondary coils was $10\text{–}30 \, \text{cm}$. In other words, each coupling coefficient $k$ was equal to $0.002\text{–}0.03$. We also calculated the transducer power gain $G_T$ using the theoretical formula described in Section 3. In the theoretical simulation, the capacitor values were set to the theoretical values $C_{1s} = C_{2s} = 0.155 \, \text{nF}$ and $C_{1p} = C_{2p} = 2.23 \, \text{nF}$

Figure 14 shows the theoretical and measured values of the transducer power gain $G_T$ at each coupling coefficient. At coupling coefficient $k = 0.006$, the measured value of transducer power gain $G_T$ was approximately 0.11, which was lower than the theoretical value. The power gain was maximized when $k = 0.011$. The reason for the differences between the theoretical and measured characteristics may be considered 1) the equivalent series resistance of the matching capacitors and so on and 2) the deviation from theoretical values of the self-inductance and parasitic resistance for the two coils. It is considered that relatively severe parts selection and initial adjustment are necessary.

![Figure 14: Variations in transducer power gain $G_T$ based on the coupling coefficient.](image)

6. Discussion

As shown in Figs. 4 and 5, when the matching parameters $C_{1s}, C_{1p}, C_{2s},$ and $C_{2p}$ are determined so as to satisfy the optimum signal source and load conditions $Z_{s,\text{opt}}$ and $Z_{L,\text{opt}}$ re-
garding the set coupling coefficient $k = 0.006$ (mutual inductance $M = 14.7 \text{nH}$), although the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$ have relatively good values compared to those without optimization, they are maximized at a coupling coefficient (mutual inductance) 0.002 larger than the given one. As shown in Fig. 7, because this deviation value is almost constant within the coupling-coefficient range of $k = 0.01$ to 0.03, the above problem can be resolved by correcting the matching parameters according to the deviation value in advance. As shown in Fig. 6, because the matching parameters change continuously, the correction value can be easily determined. Then, when the matching parameters are actually corrected according to the deviation value, as shown in Figs. 8 and 9, the transducer power gain $G_T$ and the magnitude of the load voltage $|V_2|$ can be maximized at the set coupling coefficient of $k = 0.006$ (mutual inductance $M = 14.7 \text{nH}$).

As shown in Fig. 10(c), when uniform errors of <1% are added to the set matching parameters, there is a possibility that a certain value in the range of $k = 0.004$–0.010 is estimated as the coupling coefficient $k = 0.006$. As shown in Fig. 8, when the matching parameters are determined at the coupling coefficient of $k = 0.006$, the transducer power gains $G_T$ at $k = 0.004$ and 0.010 are approximately 0.44 and 0.43, respectively. Because the optimum transducer power gain $G_T$ is approximately 0.53 when the exact coupling coefficient is estimated, even if there are errors in the set matching parameters, it is possible to achieve 83% and 81% with respect to the optimum condition. If the matching parameters can be set in the aforementioned range, the two-coil power-transfer system can operate with sufficient efficiency for practicable applications.

In this study, to evaluate the appropriateness of the simulation conditions and the feasibility of the obtained results, the coupling coefficient and power gain were measured. The self-inductance $L$ and parasitic resistance $R$ of the primary and secondary coils were similar as the simulation conditions described in Section 4. As shown in Fig. 12, the theoretical and measured values of the coupling coefficient were nearly the same, and both results were considered to be correct. The coupling coefficient $k = 0.002$–0.03 corresponded to approximately 10 to 30 cm of the distance between the two coils, and with respect to the actual application, corresponded to near-field communication between the chest and upper arm or between the forearm and upper arm. In addition, as shown in Fig. 13 and 14, although the measured values of the power gain tended to be smaller than the theoretical values, the basic characteristic shapes agreed.

7. Conclusion

A two-coil wireless power-transfer system with an L-section matching network was examined in order to realize the optimum power transfer for magnetic-coupling intrabody communication. First, the matching parameters using the optimum signal source and load conditions
were computed, and it was confirmed that the calculated transducer power gain and magnitude of the load voltage were not optimum at the set coupling coefficient. Second, the matching parameters and the mutual inductance yielding the optimum transducer power gain at each coupling coefficient were obtained. The results revealed that the matching parameters changed smoothly and monotonically with respect to the coupling coefficient and that the mutual inductance giving the optimum transducer power gain did not match the set value and had nearly constant deviation. Third, it was confirmed that the nearly optimum transducer power gain could be obtained at the actual set coupling coefficient by correcting the matching parameters according to the deviation value. Furthermore, it was shown that the magnitude of the load voltage was nearly maximized at this time. Lastly, it was shown that the mutual inductance could be estimated and the nearly optimum power transfer could be realized even if the positional relation between the sending and receiving coils was dynamically changed. Through measurements, we clarified that the coupling coefficient set by the simulation corresponded to the short distance communication, and the transducer power gain could be improved by the L-section matching network. In a subsequent study, a four-coil wireless power-transfer system will be analyzed in the same manner. Additionally, we want to develop a transceiver to examine whether dynamic optimization is possible.

Acknowledgement

This work was partially supported by JSPS KAKENHI Grant Number JP18K12152 and UNION TOOL foundation. The author wishes to thank M. Nakajyo and A. Morita for his help in measuring S-parameters.

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