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Abstract. Galvanic coupling intrabody communication (IBC) is a novel technique that uses the human body as a transmission medium, and it is suitable for connecting wearable and implantable sensors and devices. If galvanic coupling IBC transmitters can be used on an insulating layer such as cloth, not only does it become easy to use but also its application range is dramatically increased. In order to examine its communicability, this study analyzed the signal path loss in partially noncontact galvanic coupling IBC using a four-terminal circuit and a finite element model. In addition, the effect of an LC series-parallel interface circuit, which effectively injects the signal into the human body, was examined. The attenuation of the transmitted signal was approximately 20 dB higher than that of contact galvanic coupling IBC at 2 MHz. The addition of the interface circuit improved the attenuation by approximately 20 dB at a resonant frequency of 2 MHz and could compensate its increment of the attenuation.

Keywords: Galvanic coupling, Intra-body communication, Circuit model, FEM model, noncontact electrode

1. Introduction

Intrabody communication (IBC) is a new transmission technology that uses the human body as its transmission medium [1]. IBC is a promising communication technology for the construction of body area networks (BANs) [2]. In general, there are two types of IBCs—namely,
capacitive coupling type and galvanic coupling type [3]. We have investigated the signal path loss in galvanic coupling IBC using two models of the human arm, with special attention given to the return path and optimization of the interface circuit [4]. These studies revealed that it is necessary to consider the capacitance of the parasitic return path when designing IBC transceivers and that an optimization interface circuit is effective in reducing signal path loss.

In galvanic coupling, the two electrodes of the transceiver must be placed directly on the skin. From the perspective of real-world applications, direct placement on the skin is a considerable constraint. If communication can be realized by placing electrodes on an insulating layer such as clothing, without requiring direct skin contact, its applicability increases. To verify the communicability and select the signal frequency, this study analyzed the signal path loss of a human arm under the condition that the two electrodes of the transmitter are located on an insulating layer; we also examined the optimization of the interface circuit, which was proposed by Okamoto et al. [5]. These electrodes are called noncontact electrodes because their conductive material is not in direct contact with the skin. Although several groups have attempted to investigate the propagation mechanism of galvanic coupling IBC [6, 7, 8, 9, 10], the possibility of noncontact communication has not been examined. For simulation analysis, we used two models. The first model is a four-terminal circuit model based on a simplified equivalent circuit representation of the human arm. The second model is a 3D finite element method (FEM) model of the human arm. These models include circuit elements such as a transmitter and receiver in addition to the body element.

The rest of this paper is organized as follows. Section 2 describes an optimized interface circuit that utilized an LC series-parallel circuit. Sections 3 and 4 describe the four-terminal circuit model and circuit-coupled FEM model of the human arm, respectively. Section 5 describes numerical simulations and in vivo measurements, followed by a discussion of the results in Section 6. Section 7 summarizes the conclusions of this paper.

2. Interface circuit utilizing LC series-parallel circuit

In order to optimize the transmission efficiency, we employed an LC series-parallel circuit as an interface circuit of the transmitter [4, 5, 11]. This LC circuit increases the signal voltage to be injected using electrical resonance. Figure 1 shows the LC series-parallel circuit and its equivalent circuit. When the capacitor component $C_{\text{arm}}$ of the human arm is much smaller than the capacitance of a capacitor $C_2$, the resonant frequency $F_r$ is determined by the inductor $L$ and the capacitors $C_1$ and $C_2$, and it is represented as follows:
The quality factor $Q$, which is a measure of the quality of a resonant circuit and is determined by a ratio of the resonant frequency to the bandwidth of the resonant circuit, depends on the resistor component $R_{\text{arm}}$ of the human arm. The LC series-parallel circuit has the additional functions of preventing electric shock and band limiting the transmitted signal.

\[
F_e = \frac{1}{2\pi \sqrt{L \left( \frac{C_1 \cdot C_2}{C_1 + C_2} \right) + C_1 + C_2}}
\]

3. Four-terminal circuit model

3.1. Model outline

In this study, the frequency range selected for the investigation was 100 kHz to 10 MHz. This region can be modeled using static circuit models. Figure 2 shows a four-terminal circuit model, which was developed by adding noncontact electrode components to the previously proposed circuit model [4]. Symbols $Z_n$, $Z_{o1}$, $Z_{o2}$, $Z_{b1}$, and $Z_{b2}$ are the impedances of the human arm; their values were determined by the application of an improved version of Song’s method [7]. The model includes the parasitic return path, the impedance of which is represented as $Z_{cc}$. As described in section 2, we employed an LC series-parallel circuit as an interface circuit for the transmitter. When incorporating the resonant interface circuit, the receiver was assumed to use an LC parallel tuned circuit. The noncontact electrodes of the transmitter are modeled as circular plates with silicon as an insulating material (which has good insulating properties—namely, a conductivity $\rho_n$ of $1.0 \times 10^{-12}$ S/m and a relative permittivity $\varepsilon_n$ of 11.7). The symbol $Z_n$ in the circuit model represents the noncontact electrode impedance shown in Fig. 2. The resistance $R_n$ and capacitance $C_n$ of the silicon portion of the electrode were calculated using the following formulas:
where $L_n$ and $S_n$ are the thickness and area of the silicon portion of the electrode, respectively. $\varepsilon_0$ is the vacuum permittivity. The remaining impedance symbols used in the models are listed in Table 1.

Figure 2: Four-terminal circuit model of the arm for galvanic coupling IBC.

Table 1: Impedance symbol

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Meaning</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_{ceo}$</td>
<td>input impedance of the receiver</td>
</tr>
<tr>
<td>$R_o$</td>
<td>output resistance of the transmitter</td>
</tr>
<tr>
<td>$Z_l, Z_{c1}, Z_{c2}$</td>
<td>impedances of the LC series-parallel circuit</td>
</tr>
<tr>
<td>$Z_{reso}$</td>
<td>impedance of the LC parallel circuit of the receiver</td>
</tr>
<tr>
<td>$Z_c$</td>
<td>coupling impedance</td>
</tr>
</tbody>
</table>

3.2. Formulation of input-output characteristic

The output voltage is calculated by applying the mesh current law to the circuit and solving the linear equations with seven unknown currents $i_1, i_2, ..., i_7$ (Fig. 2) [8]. The seven linear
equations for a single frequency are shown in (4) in a matrix form.

\[
\begin{bmatrix}
V_i \\
0 \\
0 \\
0 \\
0 \\
0 \\
0
\end{bmatrix}
= 
\begin{bmatrix}
R_o + Zc_1 + Zc_2 + Zl & -Zc_2 & 0 & 0 \\
-Zc_2 & 2Zn + Zc_2 + Zl & -Zl & -Zl \\
0 & -Zl & Zb_2 + Zl + Zt_1 & Zb_2 + Zl \\
0 & 0 & -Zl & Zb_2 + Zl & Zb_2 + Zl + Zl + Zo \\
0 & 0 & 0 & 0 & -Zo \\
0 & 0 & -Zn & -Zb_2 & -Zb_2 \\
0 & 0 & 0 & -Zb_2 & -Zb_2 - Zo
\end{bmatrix}
\begin{bmatrix}
i_1 \\
i_2 \\
i_3 \\
i_4 \\
i_5 \\
i_6 \\
i_7
\end{bmatrix}
\]

(4)

where \( V_i \) is the output voltage of the transmitter and \( Z_{cero} \) denotes a series connection between \( Z_c \) and the parallel connection of \( Z_{ceo} \) and \( Z_{reso} \):

\[
Z_{cero} = Z_c + \frac{Z_{ceo}Z_{reso}}{Z_{ceo} + Z_{reso}}
\]

(5)

The output voltage \( V_o \) can be calculated as

\[
V_o = (i_5 - i_6) \times \frac{Z_{ceo}Z_{reso}}{Z_{ceo} + Z_{reso}}
\]

(6)

The signal attenuation and phase shift can be evaluated according to the following equations:

\[
H_c = -20\log_{10} \left( \frac{|V_o|}{|V_i|} \right) \text{ dB}
\]

(7)

\[
\theta_{rad} = \tan^{-1} \left( \frac{\text{imag}(V_o)}{\text{real}(V_o)} \right)
\]

(8)

In addition, we evaluated the input voltage \( V_h \) for the human arm using the formula

\[
V_h = |(i_1 - i_2) \times Z_{c2}|
\]

(9)

4. Circuit-coupled FEM Model

4.1. Model outline

Figure 3 shows a circuit-coupled FEM model, which was also developed by adding noncontact electrode components to the previously proposed FEM model [4]. The model is designed using COMSOL 5.0 software [12]. The human arm is modeled as a cylinder with a 50 mm radius and a 500 mm length. It consists of five concentric and homogeneous layers: skin, fat, muscle, cortical bone, and bone marrow. The thicknesses of these different layers are 1.5 (skin), 8.5 (fat), 27.5 (muscle), 6 (cortical bone), and 6.5 mm (radius, bone marrow). The dielectric properties of the human tissues can be expressed by their conductivity and rel-
ative permittivity. These values are determined using the website calculation tool Calculation of the Dielectric Properties of Body Tissues [13]. The symbols $R_o$, $Z_l$, $Z_{c1}$, $Z_{c2}$, $Z_{ceo}$, $Z_{reso}$, and $Z_{cc}$ are the same as those of the circuit model. The coupling impedance between the electrode and skin is modeled using the contact impedance, which is one of the functions of COMSOL 5.0, rather than the circuit element $Z_c$. The noncontact circular electrodes are modeled in the same manner as the circuit model and are placed in contact with the skin layer; the conductive material does not touch the skin layer.

Figure 3: Circuit-coupled FEM model of the arm for galvanic coupling IBC. Simplified cylinder model of a human arm with five layers, including circuit elements.

4.2. Numerical implementation

Because the inductive effect and wave propagation can be neglected in biological tissues in the frequency range selected, Maxwell’s equations can be simplified using the continuity equation and constitutive relations, leading to the following equation for the quasi-static electric field [9, 14]:

$$-\nabla \cdot \left\{ \left( \sigma + j \omega \varepsilon \right) \nabla \nu \right\} = 0$$  (10)

where $\sigma$, $\omega$, $\varepsilon$, and $\nu$ are tissue conductivity, field angular frequency, tissue relative permittivity, and electric potential, respectively. On the skin surface, a zero-current flux is considered, and the Neumann boundary condition is employed (11), where $n$ is the surface normal and $J$ is the current density distribution. On all interior layer boundaries, the continuity of the current flux is applied (12):

$$n \cdot J = 0$$  (11)
The geometry of the human arm cylinder was meshed using tetrahedral and triangular elements. A finer discretization was used around the electrodes. The number of the elements was approximately 60,000.

5. Simulation and measurement

5.1. Simulation and measurement methods

In this simulation, the influence on the signal attenuation of the signal transmission distance, parasitic return path amplitude, and insulating layer (silicon) thickness was examined. The effectiveness of the interface circuit of the LC series-parallel circuit was also verified. The resonant frequency of the LC series-parallel circuit was set at 2 MHz. Table 2 lists the parameters used in both models for this simulation.

Table 2: Simulation parameters. The underlined values are defaults.

<table>
<thead>
<tr>
<th>Name</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electrode distance</td>
<td>100, 300 [mm]</td>
</tr>
<tr>
<td>Electrode radius</td>
<td>5 [mm]</td>
</tr>
<tr>
<td>Electrode interval</td>
<td>30 [mm]</td>
</tr>
<tr>
<td>Insulating layer (silicon) thickness Ln</td>
<td>0.5, 1.0, 1.5 [mm]</td>
</tr>
<tr>
<td>Parasitic return path Zcc</td>
<td>100, 200, 300 [pF]</td>
</tr>
<tr>
<td>Impedances of the LC series-parallel Zl</td>
<td>100 [H]</td>
</tr>
<tr>
<td>Impedances of the LC series-parallel Zc1</td>
<td>126 [pF]</td>
</tr>
<tr>
<td>Impedances of the LC series-parallel Zc2</td>
<td>126 [pF]</td>
</tr>
</tbody>
</table>

In order to validate the simulation results, we measured signal path loss of the human arm using non-contact electrodes on the transmission side. Figure 4 shows the measurement setup that consists of a battery-powered notebook PC, a DDS function generator, and an Anritsu MS8608A spectrum analyzer (input impedance: 50Ω). Electrode pair separation and the transmitting and receiving distances were set to 3 cm and 10 cm, respectively. The function generator provides sinusoidal signals with a power of 3 dBm (50 Ω load). The frequency range selected for the investigation of signal attenuation characteristics was 100 kHz to 10 MHz, and the resonant frequency of the LC series-parallel circuit was set at 2 MHz. The sig-
nal path loss was measured in two directions: from arm to wrist, and from wrist to arm.

Figure 4: Measurement setup.

5.2. Simulation and measurement results

Figure 5 shows the signal attenuation and phase values when the signal transmission distance was 10 and 30 cm in the four-terminal circuit model and circuit-coupled FEM model. As shown in Fig. 5(a), the attenuations in both models decreased as the signal frequency increased, but after 8 MHz, the decrease became more gradual. Although the absolute values differed by roughly 10 dB at most, the results for both models exhibited similar tendencies. As shown in Fig. 5(b), the phase in both models did not change significantly from 100 kHz to 1 MHz but decreased as the signal frequency increased further, and the phase and phase shift values in the FEM model were greater than those in the circuit model. An increase in the transmission distance from 10 to 30 cm caused only a slight increase in the signal phase in a range greater than approximately 1 MHz. Figure 6 shows the signal attenuation and phase values when the capacitance of the parasitic return path was 100, 200, and 300 pF. A variation in the capacitance of the parasitic return path caused almost no change in the attenuation and phase values. Figure 7 shows the signal attenuation and phase values when the insulating layer thickness is 0.5, 1.0, and 1.5 mm. The signal attenuation of both models decreased by roughly 10 dB at most in the overall frequency range as the thickness decreased and the phase shift increased, especially in the higher frequency range.

Figure 8 shows the signal attenuation and phase values with and without the resonant circuits. As shown in Fig. 8 (a), the signal attenuation was smallest at the resonant frequency of 2 MHz, and the signal attenuation values were –11.2 dB in the circuit model and –4.7 dB in the FEM model. Without the resonant circuits, the values of the signal attenuation at 2 MHz were –31.5 dB in the circuit model and –25.6 dB in the FEM model, as shown in Fig. 5.
The attenuation was improved by 20.3 and 20.9 dB, respectively. As shown in Fig. 8(b), the signal phase drastically changed by approximately 140–160° near the resonant frequency. Figure 9 shows the input voltage applied to the human arm with the resonant circuits. Both input voltages experienced a tenfold increase.

Fig. 10 shows the signal attenuation values measured with and without resonant circuits. In both directions, the signal attenuation measured without the resonant circuits decreased as the signal frequency increased, but around 8 MHz, the attenuation increased. The signal attenuation measured with the resonant circuits was smallest at the resonant frequency of 2 MHz. The attenuation was improved approximately by 12 dB.

Figure 5: Frequency characteristics of signal attenuation and phase shift when signal transmission distance is 10 and 30 cm; cir.: circuit model, FEM: finite element model; capacitance of the parasitic return path: 200 pF; insulating layer thickness: 1.0 mm.

Figure 6: Frequency characteristics of signal attenuation and phase shift when capacitance of parasitic return path is 100, 200, and 300 pF; cir.: circuit model, FEM: finite element model;
signal transmission distance: 10 cm; insulating layer thickness: 1.0 mm.

Figure 7: Frequency characteristics of signal attenuation and phase shift when insulating layer thickness is 0.5, 1.0, and 1.5 mm; cir.: circuit model, FEM: finite element model; signal transmission distance: 10 cm; capacitance of the parasitic return path: 200 pF.

Figure 8: Frequency characteristic of signal attenuation and phase shift with and without resonant circuits; cir.: circuit model, FEM: finite element model; signal transmission distance: 10 cm; capacitance of the parasitic return path: 200 pF, insulating layer thickness: 1.0 mm.
6. Discussion

As shown in Figs. 5 through 7, although the absolute values differed to some extent, the results with the two models had similar tendencies: the attenuations in both models decreased as the signal frequency increased. Because they decreased gradually in a higher frequency range, it is thought that there is a frequency band in which the attenuation is at its minimum in the range of more than 10 MHz. In addition, the quantities of the attenuation and phase hardly changed because of the signal transmission distance and return path capacitance. These results indicate that the characteristics are thought to be negligibly affected by the outer environment fluctuation. In contrast, the quantities varied by the insulating layer thickness. When the electrode is thinner, because the impedance is reduced, it is thought that the signal easily passes through the human body and that the overall attenuation decreases, as shown in Fig. 7. Table 3 lists the resistance \( R_n \) and capacitance \( C_n \) for various insulating layer thick-
nesses of the transmitting electrode, which were calculated using equations (2) and (3). The resistance value was very large with a teraohm magnitude, and only the capacitance component affected the frequency characteristics.

Table 3: Resistance $R_n$ and capacitance $C_n$ of the silicon portion of the electrode

<table>
<thead>
<tr>
<th>Thickness [mm]</th>
<th>Resistance [T]</th>
<th>Capacitance [pF]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>6.37</td>
<td>16.27</td>
</tr>
<tr>
<td>1</td>
<td>12.7</td>
<td>8.14</td>
</tr>
<tr>
<td>1.5</td>
<td>19.1</td>
<td>5.42</td>
</tr>
</tbody>
</table>

As shown in Figs. 8 and 9, although the values of the signal attenuation and input voltage at 2 MHz as well as the manner of the change in the phase around 2 MHz differed to some extent, the results of the two models also had similar tendencies: the signal attenuation was at its minimum at 2 MHz with the maximum input signal found at 2 MHz; the phase drastically shifted around 2 MHz. Adding the resonant interface circuits into the transceiver and receiver circuits led to a signal attenuation improvement of 20.3–20.9 dB, and it is thought that this would be a highly effective method. In this simulation, the resonant frequency was set at 2 MHz. If inductor $L$ is 100 µH and external capacitor $C_1$ equals $C_2$, then $C_1$ and $C_2$ are ideally calculated to be 126 pF using (1). Because the capacitor component of the insulating layer was very low as listed in Table 3, the resonant frequency became 2 MHz using these values in both models. Also, the quality factor $Q$ depends on the resistor component of the insulating layer. The resistance value is very high as listed in Table 3, and the quality factor $Q$ is predicted to be very high. The tenfold increases in the input voltages and the drastic shift in the signal phase support the prediction (Figs. 8(b) and 9). Thus, because the resonant frequency can be easily set up and high amplification of the input voltage can be surely gained, the LC series-parallel circuit as an interface circuit of the transmitter is thought to be very suitable for galvanic coupling IBC using noncontact electrodes.

As shown in Figs. 8(a) and 9, a similar tendency can be observed in both simulation and measurement results. Without the resonant circuits, the simulated and measured signal path loss decreased as the signal frequency increased and its trend has changed into the worsening state around 8 MHz. With the resonant circuits, the signal path loss was the smallest at the resonant frequency, and the attenuation was improved drastically. The external capacitors $C_1$ and $C_2$ used in the measurement were set at 100 pF by a few trial-and-error measurements. These results suggest that the two models were able to simulate the signal path loss in partially noncontact galvanic coupling IBC relatively precisely and the LC series-parallel circuit
Finally, we compared the signal path loss characteristics between the noncontact electrodes and the conductive electrodes. The results were calculated using the same models and conditions except that the electrodes were different. The attenuation of the transmitted signal was minimized within a range of 2–7 MHz when using the conductive electrodes of the transmission side [4]. Also, the attenuation was lower than that of the noncontact electrodes by up to approximately 20 dB at 2 MHz. However, the improvement of the attenuation by adding the LC series-parallel circuit was 1.9–5.8 dB and was not very large at the resonant frequency (2 MHz). With regard to the signal frequency, a frequency of several megahertz was suitable for the conductive electrodes and 10 MHz or more was suitable for the noncontact electrodes. It is preferable that the transceivers be able to communicate with each other in the same signal frequency in either case of conductive and noncontact electrodes. In light of the frequency characteristics, it is thought that a frequency of several megahertz is more suitable as the signal frequency of the contact and noncontact shared galvanic coupling IBC for the following reasons: the frequency band was minimized in the conductive electrodes, and the LC circuit can expect improvement of attenuation to some degree; although the attenuation of the noncontact electrodes is larger than that of the conductive electrodes, that amount can be compensated by the LC circuit; at high frequencies such as 10 MHz and higher, the human body radiates signals to the air as electromagnetic waves [8], and thus it is thought that confidentiality and communication efficiency, which are advantageous aspects of galvanic coupling IBC, are reduced. In addition, in the state using the conductive electrodes, the resonant frequency does not match the theoretical value, which is calculated using equation (1) [4]. The transceivers require a function for changing the inductor $L$ and the capacitors $C_1$ and $C_2$ of the LC circuit depending on contact or noncontact conditions. For this improvement method, we propose that the transceivers equipped with a simple impedance measurement function measure the input impedance of the human body before communication and determine the resonant constants depending on the impedance value. If the input impedance is high, the resonance constant is set to the theoretical value; if the input impedance is low, the resonance constant is set lower than the theoretical value.

7. Conclusions

This study investigated the signal path loss of galvanic coupling IBC using noncontact electrodes on the transmission side in the frequency range of 100 kHz to 10 MHz using two models of a human arm to examine the possibility of noncontact communication and the effectiveness of the optimization circuit. The simulation results revealed that although the signal attenuation in the noncontact state is greater than that in the contact state, it can be improved.
up to that in the contact state using the optimization circuit. Therefore, it is thought that communication in the noncontact state can also be achieved at the same quality as the contact state. Considering the coexistence of the contact and noncontact states, the most suitable frequency is thought to be approximately several megahertz.

Acknowledgement

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References


