Gaped Two-Loop Antenna-Based Magnetic Transceiver With an Empirical Model for Wireless Underground Communication

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ABSTRACT

This paper presents a practical low-power, low-frequency magnetic induction (MI) transceiver based on symmetrical gaped two-loop antennas with an empirical circuit model for wireless underground communication. Because of the quasi-two-dimensional loop antennas for the transmitter (Tx) and the receiver (Rx), the tailored hardware platform is not only simple, but can be readily extended to a mutually orthogonal 3-D scheme. Through a mixed model of actual measurement data and theoretical parameters, we developed the two-loop antenna with the space of the antenna radius optimized for energy efficiency. The Tx power almost doubles that of the single Tx one, which provides channel expansion as well as system stability in terms of the Tx current level. Accordingly, the Rx signal is considerably increased by 9.5 dB in contrast with a typical single-Tx single-Rx configuration. Moreover, we used empirical-based modeling for a reliable electrical model. The empirical mutual inductance, which was modeled in a 25 m corridor, resulted in a channel correction factor of 0.86, showing that the signal intensity was reduced by 14% in the underground corridor compared to the free space. A maximum range of 68 m under a SNR of 9.5 dB and a path loss of 100.7 dB on a load of 50 Ω at 34.5 dBm were predicted from the model. In addition, this study confirmed the system feasibility by a SNR of 8.7 dB in a practical lossy environment between two floors, where its value was converted into a MI channel at a distance of 70 m in the underground corridor.

INDEX TERMS

Magnetic fields, electromagnetic induction, electronic circuits, modeling, wireless communication.

I. INTRODUCTION

Wireless Internet of Things (IoT) platforms are applied to a specific environment by means of cutting-edge technology to monitor and control infrastructure in limited or inaccessible circumstances. Wireless underground sensor networks (WUSNs) based on classical electromagnetic (EM)-wave technology in an air medium, e.g., Bluetooth or ZigBee [1]–[3], are effective owing to their high-data rates [1], low-cost competitiveness [2], and low-power consumption [3]. However, unlike normal or steady state systems, these EM wave systems cannot operate if the propagation path is blocked by solid materials with high absorption of EM waves. For the mmWave bands, EM waves deteriorate rapidly in solid channels due to the skin effect [4], [5]. For harsh environments, using a very low frequency (VLF, 3 kHz to 30 kHz) band with a wavelength of 10 km or more, can be a solution [5], [6]. However, conventional transceivers of the WUSNs in the VLF band may suffer from large dipole antenna sizes [6]. Additionally, two-way wireless communication or localization is necessary to connect and rescue people who are suddenly trapped without power in a building or mine collapse. Then, a battery-powered mobile platform at the physical layer of the WUSNs is essential for the underground emergency response [7], [8].

Many magnetic induction (MI)-based communication systems have been found in the literature for underground [4], [5], [7]–[10] and underwater [11]–[13] applications. In particular, low-frequency two MI link systems, MI communication [5]–[7], [9], [10], [14]–[16] and MI
localization [2], [8], [17]–[21], comprise a promising method in extreme underground environments for rescue apparatuses for two-way connection. The two applications are divided into the physical layer and the data link/node networking [6]–[8]. Here, their physical layer is generally similar for the case of an emergency disaster, so the hardware design requirement is limited to the MI-based link system. The hardware design of WUSNs leads to the crucial factors [6]: the antenna volume, the operating frequency, the transmitter (Tx) power, and the receiver (Rx) readout module. Firstly, two-way communication requires a small antenna scheme with wavelength independence, and configurations are implemented in the two types: the narrow band resonant type (high efficiency) or the wideband flat coil type (low efficiency) [3]–[5]. The resonant type is mainly used due to power efficiency. The resonant loop antenna induces a magnetic field directly related to the data at an operating frequency that depends only on the matching capacitance, not the diameter of the loop antenna [3], [4]. This feature allows a loop antenna to be small and makes the mobile MI-link feasible [2], [8], [15], [16]. Secondly, the fundamental necessity to readily connect between underground channels is the operating frequency [5], [8], [14]. The operation frequency can strongly affect both the bandwidth of the system and the skin depth [4]–[8] of the magnetic channel. The higher the operating frequency is, the larger the bandwidth is with respect to the hardware system design, but this also reduces the skin depth. The long-wavelength (VLF) band or near (2 kHz or 125 kHz) can penetrate solid media, such as rock, aqueous soil, concrete, air, or wood, etc., in the condition of the quasi-static field (the reactive near field) rather than the propagation wave (the far field) causing multipath loss or fading influence [5]–[8], [11], [16]. For this reason, the operating frequency that satisfies the skin depth must be selected. For example, the skin depth at the VLF band is in the range of 29 m to 92 m when the conductivity of complex media is assumed as 0.01 S/m. Thirdly, when the above two conditions are satisfied, the Tx power range determines the ability of bidirectional communication [2], [3], [8]. Regarding the feasibility of mobile sensor nodes, low-power small transceivers benefit from the loop antenna scheme. Here, the magnetic intensity decreases rapidly with an increasing distance (path loss of 60 dB/decade) [8], because the magnetic intensity is inversely proportional to the cube of the distance between two coupled antennas in the near field [2]. Given the low operation frequency, the coverage within tens of meters is subsequently controlled by the transmitter (Tx) power (more than 30 dBm). In [2], the MI-based WSNs were presented using a commercial off-the-shelf (COTS) product. However, it is not simple to control the output power or the minimum received signal (80 μV) [2] due to the COTS system. Thus, dealing with tailored Tx power is important to enhance the overall performance. Lastly, the receiver (Rx) features, such as the signal-to-noise ratio (SNR), path loss (PL), and magnetic flux density [6], [8] etc., can be characterized from the magnetic medium and the Tx power along with the Rx readout module (Rx-RM). The signal integrity of Rx antennas is improved by the deployment configuration of a MI guided scheme [4], [17] or multi-loop coils [7], [19]. The lifetime of a mobile sensor node is determined by the budget of the power management. For example, at a 5 V (USB 3.0 voltage) bias, a 10,000 mAh rechargeable battery (Xiaomi, PLM132M, 37 Wh) can continuously handle a 37 dBm Tx module for 7.4 hours.

An equivalent circuit model [2]–[4], [9]–[11], [14], [15] reflecting practical boundary conditions is necessary to design a MI data link system. This model has the advantage of properly predicting the performance of the entire hardware system, including solid media. However, some parameters cannot be obtained by theoretical models. For a real environment, it seems complicated or not easy to estimate the mutual inductance [3], [4], [17] in terms of the connectivity between the Tx and the Rx. In [4], the mutual inductance was modeled by the theoretical equation, but the deviation in real circumstances is not negligible (in our case: 14% for the Tx part). The modeled mutual inductance is too simplified as an ideal assumption. Moreover, the evaluation of the Tx parasitic resistance related to the Tx power level is impossible by theoretical calculation due to a practical characteristic. The Tx parasitic resistance of a closed loop circuit can only be extracted from empirical data to determine the proper Tx power. If this parameter is not clear, it should be omitted [3], [4], [17]. In this case, the circuit model tends to be simplified with limited reliability. For example, the ratio of the parasitic to intrinsic resistance in our Tx electrical circuit is 60%, which shows that the parasitic resistance has to be defined and evaluated. Thus, when modeling equivalent circuits, empirical variables in the real world must be considered in order to accurately design the performance of the hardware system.

In this paper, we provide a detailed description of a symmetrical gaped two-loop antenna-based low-frequency MI transceiver with an empirical circuit model in order to realize the hardware system of WUSNs for practical two-way underground connectivity, and we investigate its feasibility in real underground environments (two-story space). We focus only on issues related to the node design (magnetic field formation, degree of influence of magnetic channel, magnetic field sensing), leaving the network/node networking to the system designer [6]. The features and contributions are listed as follows.

- We give a practical low-power MI transceiver based on symmetrical gaped two-loop antennas for WUSNs, which is applicable to two-way communication as the hardware system of a mobile sensor node.
- We devise a highly efficient quasi 2-D antenna block. Since this antenna module can be readily extended to a mutually orthogonal 3-D scheme [2], [3], [21], it can be used as a core antenna block for various applications of WUSNs.
- We provide an empirical circuit model for each module (Tx, Rx) including a practical magnetic channel, and
evaluate the performance of the whole system using the model. This reliable model helps one intuitive to understand the MI link system.

- We realize whether a MI-based link system could be used in buildings with concrete walls, and show the preliminary data as a reference. Conveniently, researchers can design and implement MI-based link systems from many given key parameters.

The outline of the paper is as follows. Section II briefly describes the theoretical background of both MI between two coils and the proposed MI transceiver. Section III discusses the Tx part while Section IV presents the Rx part for detecting signals in a MI based channel. Section V describes the proposed MI-based link system and its equivalent circuit. Section VI discusses several channel effects involving underground environments. The paper is concluded in Section VII.

II. LOW FREQUENCY MI LINK SYSTEM

This section explains the MI principle of a coupled loop antenna and describes a MI-based link system for underground applications. The MI theoretical background is presented by the structure of two coupled circular loop antennas. Each Tx and Rx characteristic is analyzed into four cases. Through this comparison, we introduce the proposed MI transceiver.

A. MI OPERATION PRINCIPLES

When a loop coil is operating in a band much lower than the self-resonant frequency, it can be approximated by a series-connected circuit with a resistor and an inductor [3], [4], [9]–[11]. If two loop coils are deployed at some interval $d_{ch}$, they are coupled by mutual inductance $M_{ch}$, changing to the equivalent impedance for each other. As one of the two loop coils with radius $a_t$ and $a_r$, respectively, is applied to the Tx part with a signal current $i_t$, as shown in Fig. 1, the signal from $i_t$ is detected in the other as the Rx part at distance $d_{ch}$ on the basis of Biot–Savart’s law. Here, $P$ is a point located on the Rx coil with angle $\theta_r$ relative to the Z axis in a cartesian coordinate system. The magnetic vector potential $A_t(\phi)$ at $P$ with regard to the Tx coil is given by [3], [4]

$$A_t(\phi) = \frac{\mu_0 i_t a_t^2 \phi}{4(d_{ch}^2 + a_t^2)^{3/2}}$$

where $\mu_0$ is the space permeability. The magnetic flux is given by integrating $A_t(\phi)$ around the Rx coil with $a_r$. $M_{ch}$ is obtained from the magnetic flux at the Rx coil divided by $i_t$. Since $M_{ch}$ for each other is symmetrical, it is given by [3], [4], [11]

$$M_{ch} = \frac{\mu_0 N_t N_r a_t^2 a_r^2}{2(d_{ch}^2 + a_t^2)^{3/2}}$$

where $N_t$ and $N_r$ are the number of coil turns on each side. Thus, increasing the radius of both coils to the upper limit is more effective than raising $N_t$ and $N_r$ under $d_{ch} \gg a_r$ conditions [3], [4]. For two-way MI connectivity, we chose the maximum radius of two coupled antennas to 0.25 m.

For an intuitive understanding, we selected four cases from a variety of loop coil configurations and compared them, as shown in Fig. 2. It is noted that all of the four cases as a core antenna block can be extended or applied to a mutually-orthogonal effective 3-D antenna scheme. The first case is the simplest way: the conventional configuration for one Tx and one Rx (Case 1: single-Tx single-Rx) as seen in Fig. 2(a). It communicates through two coupled coils at $d_{ch}$. Case 2 is a method of adding a coil at the Rx part in Case 1. If two identical Rx coils with one Tx coil (Case 2: single-Tx two-Rx) are applied at the conditions of $d_{ch} \gg a_r$, the induced voltage $V_{ch}$ in the Rx part can be doubled due to the two Rx coils, as shown in Fig. 2(b). At a given Tx power, it can be effective without a large Tx power loss; for instance, in our case, we needed an additional Tx power of at least 3 dBm to double the induced voltage at a distance of 25 m. Conversely, Case 3 is an inverse combination of Case 2. Fig. 2(c) is a method of Case 3 (two-Tx single-Rx) extending the transmission distance almost twice compared to Case 1. This structure is suitable for broadcasting systems between large base stations and small sensor nodes. Case 4 is a structure that improves both Tx and Rx characteristics. In order to simultaneously extend the coverage and improve the sensitivity of the Rx part, Case 2 should be combined

![FIGURE 1. Detailed illustration of a mutual inductance between two circular coils: orientation, coil scheme, and coupling principle.](image)

![FIGURE 2. Loop antenna configuration of (a) single-Tx single-Rx (Case 1) (b) single-Tx two-Rx (Case 2) (c) two-Tx single-Rx (Case 3) (d) gaped two-Tx gaped two-Rx (Case 4).](image)
with Case 3, which creates Case 4 (gaped two-Tx gaped two-Rx) as shown in Fig. 2(d). Here, an important factor is \( M_{\text{ch}} \).

The tight coupling between the two coils interferes with the two coils’ current flow. Obviously, some spacing \( g_t \) and \( g_r \) between them on both sides can be adjusted for both high Tx power and an enhanced received signal. Thus, Fig. 2(d) shows the optimized scheme.

### B. GAPED TWO-LOOP ANTENNA-BASED MI SYSTEM

Fig. 3 shows a proposed MI transceiver based on gaped two-loop antennas, which is divided into three parts: the Tx module (magnetic field formation), the magnetic channel (degree of influence of magnetic channel), and the Rx module (magnetic field sensing) [3], [4], [6]–[8]. Each item considered for design in the three areas of the physical layer is also shown. The design of the Tx module is related to an antenna volume and field coverage as well as the operation frequency and the bandwidth, while the issues of the Rx module are a transceiver lifetime, an antenna scheme, and Tx characteristics. Here, all the variables will be investigated except for the lifetime issue which is mainly the domain of the networking design. Also, for the magnetic channel, both the skin depth [5] and the background noise [8] are dominantly working as a design parameter. The skin depth works as the parameter of the upper limitation for the operation frequency, whereas the background noise is considered as the bottom limitation. We chose the lowest frequency band (19 kHz - 21 kHz) of the International Telecommunication Union-Radio communication Sector/Study Groups 1 (ITU-R/SG1) standard dealing with wireless power transmission compatible with loop antenna and its driver.

The Tx module consists of four blocks: two Tx drivers with input voltage \( V_{t1} \) and output voltages \( V_{m1}, V_{m2} \) from a signal source, series resistors \( R_{t1}, R_{t2} \) and series inductors \( L_{t1}, L_{t2} \) consisting of a loop antenna, matching capacitors \( C_{t1}, C_{t2} \) controlling an operating frequency, and parasitic resistance \( R_{\text{par}} \) in a closed electrical circuit. The Rx module is composed of two loop antennas, which are symmetrical to the Tx part and an Rx-RM. Like the Tx part, the two antennas are consisting of series resistors \( R_{r1}, R_{r2} \) and series inductors \( L_{r1}, L_{r2} \). The Rx-RM detects and amplifies two tiny input signals to an analog-to-digital converter (ADC). The magnetic channel of MI-based RF link system is linked by \( M_{\text{ch}} \) of symmetrical antennas at \( d_{\text{ch}} \). The Tx module is usually relevant to a Tx power, whereas the Rx module are related to \( V_{\text{ch}} \) (or sensitivity). The magnetic channel is mostly associated with path loss \( PL \). For the underwater case [11], 40 dBm is required because of the inverse proportion of the field intensity to the cubic of the distance. It is noted that Joule heating can degrade the system. This means that the transmission distance and the system stability are at odds. Therefore, the design of Tx power plays an important role in determining the performance of the overall system.

In turn, we chose two loop coils at an interval (Case 4), which not only improved the Tx power, but also mitigated the negative effects of \( M_{\text{ch}} \), while maintaining the stability of the system. The parallel structure equally shared the total current. The two coils for the two sides were considered to be equally effective coils because \( d_{\text{ch}} \) was much wider than the spacing between two adjacent coils \( g_t \) or \( g_r \). In addition, to simplify the entire link system, the antenna of the Rx part was designed with the same structure (Case 4) as the Tx part, which produced a symmetrical scheme. Moreover, the Rx-RM was developed to effectively improve the sensitivity from the instrumentation amplifier.

### III. TRANSMITTING SYSTEM

This section describes a high Tx power system for MI-based RF links. A Tx module based on Case 4 is equipped with a Tx driver. Characterization of the Tx driver is first done to design the Tx power. In addition, the antenna for Case 4 is analyzed for electrical and mechanical models. As mentioned in the
previous section, it is designed for the class of about 40 dBm power to implement mobile two-way communication.

**A. Tx DRIVER AND GAPED TWO-LOOP ANTENNAS**

Prior to developing the quasi 2-D antenna, a Tx driver is modeled to estimate the Tx power, especially for the impedance of the signal source \( R_{in} \) and \( R_{par} \). In [2]–[4], [17], the equivalent model was oversimplified by omitting \( R_{in} \) and \( R_{par} \) even though the magnitude was not negligible for \( R_{in} \) or not predictable for \( R_{par} \) due to empirical data. Thus, we modeled the equivalent circuit on the basis of empirical parameters. The Tx driver was chosen for the BJT-based differential amplifier where negative feedback was applied to control the gain and frequency response. The theoretical power conversion efficiency of the push-pull amplifier for Class-AB is 78.5% due to empirical data. Although the magnitude was not negligible for \( R_{par} \) model was oversimplified by omitting \( R_{par} \). However, the ideal assumption may deviate from the practical wire-based coils. Thus, we chose an empirical method to rearrange the initial number of turns as an effective number of turns \( N_{eff} \), showing that \( N_{eff} \) is 23.29 from the LCR meter. In accordance with the correction factor \( \alpha_{d} \), the modeled \( L_{t1} \) was determined as \( \mu \alpha_{d} \pi N_{eff}^{2} a_{1}/2 \), where \( \alpha_{d} \) was 1.37. At 20.84 kHz, \( L_{t1} \) and \( R_{t1} \) were 268 \( \mu \)H and 0.5 \( \Omega \), respectively. As with \( L_{t1} \), it was also difficult to model the mutual inductance from the theoretical mutual inductance \( M_{t-the} \) on account of the complex boundary conditions. Thus, the empirical mutual inductance \( M_{t}(= \alpha_{d} M_{t-the}) \) was adopted, where \( \alpha_{d} \) was the correction factor of Tx. A loop antenna with \( C_{t1} \) had the first resonant frequency at \( (2\pi f_{res})^{2} = 1/L_{t1}C_{t1} \). If one coil with \( C_{t1} \) was coupled to

where \( R_{d1} \), \( R_{d2} \), \( C_{d1} \), \( C_{d2} \) were 1 k\( \Omega \), 14 k\( \Omega \), 3.3 nF, 47 pF, respectively. The simulated and measured data are shown in Fig. 4(b). With (3), the theoretical dominant zero was 3.45 kHz whereas the two poles were 48.25 kHz and 241.3 kHz, respectively. As a result, we tailored the output signal under 10 kHz with \( C_{d1} \) of 3.3 nF rather than 33 nF to control the background noise. The measured data agreed well with the simulation data. Fig. 5 shows the implemented Tx driver module on the printed circuit board (PCB).

For a high Tx power, we developed a small loop antenna with a radius of 0.25 m and 17 turns as shown in Fig. 6. Here, the diameter of each turn was 3 mm consisting of 700 strains of 40 AWG litz wire. Theoretically, the self-inductance of a circular coil can be determined as \( \mu \alpha_{d} \pi N_{eff}^{2} a_{1}/2 \) [4], [9], [10]. However, the ideal assumption may deviate from the practical litz wire-based coils. Thus, we chose an empirical method to rearrange the initial number of turns as an effective number of turns \( N_{eff} \), showing that \( N_{eff} \) is 23.29 from the LCR meter. In accordance with the correction factor \( \alpha_{d} \), the modeled \( L_{t1} \) was determined as \( \mu \alpha_{d} \pi N_{eff}^{2} a_{1}/2 \), where \( \alpha_{d} \) was 1.37. At 20.84 kHz, \( L_{t1} \) and \( R_{t1} \) were 268 \( \mu \)H and 0.5 \( \Omega \), respectively. As with \( L_{t1} \), it was also difficult to model the mutual inductance from the theoretical mutual inductance \( M_{t-the} \) on account of the complex boundary conditions. Thus, the empirical mutual inductance \( M_{t}(= \alpha_{d} M_{t-the}) \) was adopted, where \( \alpha_{d} \) was the correction factor of Tx. A loop antenna with \( C_{t1} \) had the first resonant frequency at \( (2\pi f_{res})^{2} = 1/L_{t1}C_{t1} \). If one coil with \( C_{t1} \) was coupled to

\[
\frac{V_{n1}(j\omega)}{V_{n2}(j\omega)} \approx \frac{1 + j\alpha_{d}R_{d1}C_{d1}}{1 + j\alpha_{d}R_{d2}C_{d2}} \quad (3)
\]
the identical one in parallel, $M_t$ could be extracted from the measured $f_{res}$ due to the coupling influence.

**B. IMPLEMENTATION FOR THE TX MODULE**

The equivalent circuit of a Tx module for Case 4 was modeled as shown in Fig. 7(a). The KVL equations are given as

$$V_{in} = (R_{1-tot} + j\omega L_{1} + 1/j\omega C_{1}) I_1 + j\omega M_t I_2$$

(4a)

$$V_{in} = j\omega M_t I_1 + (R_{2-tot} + j\omega L_{2} + 1/j\omega C_{2}) I_2$$

(4b)

where $R_{1-tot} = 2R_{in} + 2R_{par} + R_{11}, R_{2-tot} = 2R_{in} + 2R_{par} + R_{21}, I_1$ and $I_2$ are the currents for the two Tx coils, and $I_t$ is the Tx current. As $I_1$ is equal to $I_2$ and $I_1 = I_2 + I_t$, $Z_{e-1} = V_{in}/2I_1 = R_{in} + R_{par} + R_{g-1} + j\omega L_{e-1} + 1/j\omega C_{e-1}$ where $R_{e-1} = R_{11}/2, L_{e-1} = [L_{1} + M_t(g_t)]/2$ and $C_{e-1} = 2C_t$. $R_{par}$ is determined from a measured loop current as well as $V_{in}$ and $R_{in}$ when $V_{in}$ of a Tx driver is measured at a given $V_t$ in Fig. 4(b). Also, from the series resonant condition, $M_t(g_t)$ for Case 2 or 4 as a function of spacing can be extracted as

$$M_t(g_t) = 1/[f_{res}(g_t)^24\pi^2C_t] - L_{tot}.$$  

(5)

The dependence of $M_t$ on the space to the maximum extent of $2a_t$ was evaluated with (3) and (5), respectively, at a distance of 5 m by using a commercial loop antenna (ETS-LINDGREN, ETS-6509) and a signal analyzer (Agilent, 35670A) as shown in Fig. 7(b). While $M_t$ without a gap was 268.0 $\mu$H, $M_t$ was 142.5 $\mu$H, making the deviation error of 88%. Obviously, $M_t$ should be modeled from empirical data. From $M_t$, the coupling coefficient $k$ was extracted to the maximum value of 0.53. In addition, a bandwidth $BW$ and $V_{ch}$ from the Tx system in Case 4 were characterized to optimize the interval of the two coils, as seen in Fig. 8(a). $BW$ was 0.46 kHz at $g_t$ of 0.25 m. When $g_t$ was equal to $a_t$, the intensity was the maximum. Also, the energy efficiency (V/J) as a figure of merit (FoM) was investigated by using the current probes (TELEDYNE, 2016) of the oscilloscope (TELEDYNE, Surfer 10M). As shown in Fig. 8(b), despite increasing DC current, it saturated at radial distance. Accordingly, the interval of two coils was determined to the radius of the loop antenna, resulting in an optimized scheme: gaped two Tx-gaped two Rx (Case 4).

![FIGURE 7](image-url)

**FIGURE 7.** (a) Electrical circuit for Two loop antennas connected with two Tx drivers and its effective equivalent circuit (b) empirical and theoretical mutual inductances for a Tx module.

![FIGURE 8](image-url)

**FIGURE 8.** (a) Induced voltage and Bandwidth characteristics and (b) Figure of merit (FoM) and DC bias current performances for the two loop antennas as a function of distance.

To evaluate whether Case 4 had better output efficiency for Case 1, the power efficiency (mV/W) denoted as the ratio of the induced voltage to the intrinsic Tx power was characterized. At a bias of 5 V, the Tx powers of Case 4 and Case 1 were 2.82 W and 1.41 W, respectively, whereas the voltage intensities were 1.73 mVrms and 0.91 mVrms at the distance of 5 m, respectively. The power efficiency was similar to each other at 0.61 and 0.65, despite an additional 5% loss.

**IV. RECEIVING SYSTEM**

The design issue of a receiving part is not tangled due to the symmetrical system. The symmetrical correlation makes the design margin of the Rx part much simpler. In this section, we explain the Rx module: the Rx antenna and the
Rx-RM, and conduct electrical modeling based on empirical parameters.

**A. TWO OPPOSITELY COUPLING Rx COILS**

To explain the effect of two oppositely-coupled inductors from the same voltage source, Fig. 9(a) shows a two mutually coupled inductor circuit that uses dot symbols which define the sign of the mutually induced voltage. Due to the influence of mutual inductance, half of the voltage induced in the gaped two Rx must be less than the magnitude of a single Rx. The KVL equations for the induced voltage $V_{ch}$ are defined as

\[ V_{ch} = (R_{r1} + j\omega L_{r1})I_{r1} - j\omega M_{r1}I_{r2} \]  \hspace{1cm} (6a)

\[ V_{ch} = j\omega M_{r1}I_{r1} - (R_{r1} + j\omega L_{r1})I_{r2} \]  \hspace{1cm} (6b)

where $I_{r1}$ and $I_{r2}$ are the currents for the two Rx coils. If $I_{r1}$ is equal to $I_{r2}$ under the condition of $d_{ch} \gg r$, one terminal voltage $V_{r1}$ is $j\omega(L_{r1} - M_{r1})I_{r1}$. In the opposite dot symbol, $V_{ch}$ is reduced by the effect of $M_{r}$, resulting in the effective self-inductance of $L_{e-r} = L_{r1} - M_{r}$. If a differential amplifier with two inputs is used in spite of the slight attenuation owing to $M_{r}$, $V_{ch}$ is detected almost twice. As a result, $V_{ch}$ can be maximized through Case 4.

![Figure 9](image-url)

**FIGURE 9.** (a) Two-coupled inductor circuit using dot symbols (b) the circuit block of the Rx readout module, and (c) the fabricated Rx readout module through the PCB.

**B. RX READOUT MODULE AND ITS MODEL**

To sense a signal from a gaped two-loop Rx antenna for tiny input signals, an instrumentation amplifier with a customized large voltage gain can be used. It was realized with two commercial operation amplifiers (TI, NE5532) as shown in Fig. 9(b) where $R_{a1}$ was 1 kΩ, $R_{a2}$ was 100 kΩ. Next, a band pass filter (BPF) was used to shape the input signal using an operation amplifier (TI, OPA2743UA) with the front-end high-pass filter ($C_{a1} = 22$ nF, $R_{a3} = 1$ kΩ) and the rear-end low-pass filter ($R_{a5} = 330$ Ω, $C_{a2} = 15$ nF), where the close loop gain ($R_{a4} = 1$ kΩ and $R_{a5} = 100$ kΩ) from the negative feedback was 100 V/V. The dominant 3-dB high frequency was 7.2 kHz while the dominant 3-dB low frequency was 32.2 kHz. Remarkably, the inserted BPF was not strongly effective due to the high gain configuration of the commercial operation amplifier to detect a mere intensity of signal, which had the gain-bandwidth product of about 10 MHz, showing that the voltage gain around the 1 kHz was the maximum. We added the gain control stage for tuning to the voltage gain $A_{gain}$ of 60 dBV. It was composed of an operation amplifier (TI, OPA2743UA) with a control loop ($R_{e8}$ of 4.7 kΩ and $R_{e7}$ of the varactor of 1 kΩ) to the output terminal of the BPF because the deviation of the characteristic for lumped commercial components exists in practical field. Here, each input impedance $R_{pre}$ of the Rx-RM was 300 kΩ. The fabricated Rx-RM is shown in Fig 9(c).

The equivalent electrical circuit for the Rx-RM is presented to appropriately investigate the Rx module of Case 4, which predicts the channel features reflecting the circumstances. As shown in Fig. 10, each input voltage of the Rx-RM is determined by $L_{e-r}$ with respect to the effect of $M_{r}$. If possible, $R_{pre}$ must be high to reduce the signal loss. Although $V_{ch}$ seems to theoretically have a nearly doubled intensity owing to a high $R_{pre}$, $L_{e-r}$ needs to be determined by empirical data. Due to the opposite connection, each $L_{e-r}$ is featured to have a reduced value by $M_{r}$ from $L_{r1}$, which produces much weaker coupling than Case 1. Nevertheless, according to the gaped two-loop Rx scheme, $V_{ch}$ can be made much higher. The $V_{ch}$ of Case 4 is improved by 3 times compared to that of Case 1.

![Figure 10](image-url)

**FIGURE 10.** Electrical circuit block for Two-loop Rx antennas connected with the Rx-RM and its equivalent circuit.

**V. EMPIRICAL EQUIVALENT CIRCUIT MODEL**

An equivalent electrical model is useful for intuitive interpretation, since it can predict critical performances, i.e., the Tx power $P_t$, SNR, received power $P_r$, and pass loss $PL$, as well as the magnetic flux density $B_{out}$. In this section, we discuss the process of modeling and its equivalent circuit, along with the characteristics of a modeled MI channel.

Fig. 11 shows the flow chart for an empirical-based electrical modeling of a MI-based transceiver. Although the model
is optimized for Case 4, the process for the other cases can be much simpler due to the optional extraction step. Considering the whole process, the model has the following characteristics: it sets the initial data of the theoretical value and determines the correction factor from the measured one. To this end, this model can be easily and accurately applied to practical applications because it can analyze characteristics of the magnetic channel.

As in the flowchart, \( L_{11} \) and \( R_{11} \) are measured first, and the effective number of turns \( N_{11} \) based on the empirical data is determined. Here, \( N_{11} \) is identical to \( N_{12}, N_{r1} \) and \( N_{r2} \). Next, as mentioned in the previous section, \( M_l \) is extracted by the measurement of \( f_{res} \) as a function of the interval between two coils. For Case 3 and 4, the correction factor \( a_t \) and the effective number of turns \( N_{e-r} \) are determined from modeled \( M_t \) as \( M_t = a_t M_{t-the} \). Then, \( R_{par} \) is extracted by the measured \( V_{in} \) at the output of a Tx driver and measured \( I_t \) in Tx coils. This plays a significant role in determining the Tx power. Unlike \( M_t \), \( M_{ch} \) and \( M_r \) are modeled by \( V_{ch} \) at the Rx point where \( V_{ch} = -j \omega M_{ch} I_t \). It is clear that \( M_{ch} \) cannot be obtained if \( I_t \) in \( R_{in} \) and \( R_{par} \) is not accurately characterized. \( M_{ch} \) reflects the boundary conditions as a specific channel circumstance. For \( M_{ch} \), there are two correction factors \( \alpha_{ch} \) and \( \alpha_{Rx} \), where \( \alpha_{ch} \) is about to the magnetic channel effect and \( \alpha_{Rx} \) is related with an Rx scheme itself. For Case 4, \( M_{ch} \) is to be \( \alpha_{ch} \alpha_{Rx} M_{ch-the} \approx \alpha_{ch} \alpha_{Rx} \mu_0 \pi N_{e-r} N_{e-r} (a_{r}^2 a_{r}^2 / 2d_{ch}^2) \) under the conditions of \( d_{ch} \gg a_r \), where \( N_{e-r} \) is the effective number of turns for the Rx part. \( \alpha_{ch} \) for Case 4 should be extracted by Case 3 due to one Rx coil with no effect of \( M_{r} \). Obviously, \( \alpha_{ch} \) is a critical factor showing magnetic field channel characteristics containing all of the actual surrounding environment. \( \alpha_{Rx} \) is a correction factor related only to Case 4, which is set to 2 for Case 4. After extracting \( \alpha_{ch} \) by Case 3, \( N_{e-r} \) is modeled from \( V_{ch} \) of Case 4. Then, \( M_{r} \) is modeled from \( L_{11} - L_{e-r} \), where \( L_{e-r} = \mu_0 \pi (N_{e-r})^2 a_r / 2 \). \( N_{e-r} \) is less than \( N_{11} \) due to the effect of \( M_{r} \). The fundamental process of modeling is completed by extracting \( M_{r} \). After that, characterization of the MI transceiver is carried out using this model, leading to \( P_{t}, P_{r}, \text{SNR}, \text{PL} \) and \( B_{out} \).

Fig. 12 shows an empirical-based equivalent circuit model applicable in all cases mentioned in the previous section. This is possible because the circuit is applied to both the Tx and Rx modules. Since the model has a signal source including \( R_{in} \) as the Thevenin’s equivalent circuit as well as \( R_{par} \), it can evaluate accurately the performance. The voltage gain is determined by

\[
\frac{V_{out}(\omega)}{V_{in}(\omega)} = \left( \frac{Z_{t-ant} + Z_{rt}}{Z_{t-tot} + Z_{rt}} \right) \left( -j\omega M_{ch} \right) \left( \frac{R_{pre}}{Z_{rt} + Z_{r} + R_{pre}} \right) \times A_{gain} \left( \frac{R_{load}}{R_{mod} + R_{load}} \right)
\]

where \( Z_{t-tot} = R_{in} + R_{par} + R_{e-1} + j\omega L_{e-1} + 1/j\omega C_{e-1} \), \( Z_{t-ant} = R_{e-1} + j\omega L_{e-1} + 1/j\omega C_{e-1} \), \( Z_{rt} = \omega^2 M_{ch}^2 / (Z_r + R_{pre}) \), \( Z_r = \omega^2 M_{ch}^2 / Z_{t-tot} \), \( Z_r = R_{e-1} + j\omega L_{e-1} \), \( A_{gain} = V_{mod} / (V_{in} - V_{in}) \). Also, \( R_{mod} \) is the output impedance of Rx-RM and \( R_{load} \) is the load resistance. Since SNR is proportional to \( M_{ch} \), the parameters of \( M_{ch} \), especially \( a_r \), can be tuned to improve channel performance. In addition, from \( \alpha_{ch} \), the channel influence is confirmed as much as the difference to the theoretical value \( M_{ch-the} \).

VI. PERFORMANCE EVALUATION

We perform whether a MI-based hardware system could be used in buildings with concrete walls as well as an underground hallway. The four cases were analyzed and compared with each other, where all the loop antennas had the 0.25 m radius and the effective 23.3 turns. The Tx power was characterized first in the semi-anechoic chamber, and then compared with values evaluated in an underground corridor. Actually, \( V_{ch} \) was compared through magnetic channel characteristics between the floors of a real building.

A. MI OPERATION PRINCIPLES

Fig. 13 shows figures of the experimental set up for Case 4 in a semi-anechoic chamber. Measured \( I_1 \) and \( I_2 \) at 20.84 kHz
related to a magnetic field of 12.1 mA/m whereas the Tx and the Rx with the interval of 5 m was 0.66 mV and the Rx antennas up to the interval range of 25 m which the line of sight was aligned to each center between the Tx in an underground hallway was evaluated as shown in Fig. 14.

The current of Case 1 was 2.42 A at DC power of 8.14 W. Here, $V_{in}$ and $2R_{in}$ for each Tx driver were 1.342 V$_{rms}$ and 21 mΩ at 5 V, respectively. $R_{par}$ was extracted to 139.5 mΩ from $I$, $V_{in}$, $R_{in}$, $R_{ant}$ and $R_{par}$. $V_{ch}$ was detected by both ETS-6509 with an antenna factor of 18.4 S/m at an operating frequency and a signal analyzer with an input impedance of 1 MΩ. It was determined to 0.87 mV$_{rms}$ from the half value of the measured voltage with the interval of 5 m in the semi-anechoic chamber. It means a magnetic field at the point to be 15.9 mA/m. Also, the magnetic channel influence in an underground hallway was evaluated as shown in Fig. 14. The line of sight was aligned to each center between the Tx and the Rx antennas up to the interval range of 25 m which was a hundred times the radius of the antenna. $V_{ch}$ between the Tx and the Rx with the interval of 5 m was 0.66 mV$_{rms}$ related to a magnetic field of 12.1 mA/m whereas $V_{ch}$ in the semi-anechoic chamber was 0.87 mV$_{rms}$. Obviously, surrounding environment made the difference by 0.21 mV$_{rms}$ from the reference condition on 0.87 mV$_{rms}$, compared with that of the semi-anechoic chamber.

As shown in Fig. 15(a), the $V_{ch}$ was compared in different cases at $d_{ch}$ of 25 m. $V_{ch}$ at Case 1 was 40.5 $\mu$V$_{rms}$ at the Tx power of 31.5 dBm. Here, on the signal analyzer’s two-channel output, the top of ETS-6509 and the bottom of Case 1 represented 8.1 $\mu$V$_{rms}$ and 40.5 $\mu$V$_{rms}$, respectively. $V_{ch}$ for ETS-6509 was half of the initial value, leading to 4.05 $\mu$V$_{rms}$. From the antenna factor of 18.4 S/m for ETS-6509, that of Case 1 was extracted to be 1.84 S/m, which represented a magnetic field through $V_{ch}$. The magnetic field for Case 1 was 104 $\mu$A/m at 31.5 dBm. For Case 3, $V_{ch}$ was 75.0 $\mu$V$_{rms}$ at 34.5 dBm, as shown in Fig. 15(b). It is clear that $V_{ch}$ was not 81.0 $\mu$V$_{rms}$, which was twice 40.5 $\mu$V$_{rms}$, but 75.0 $\mu$V$_{rms}$ due to the influence of $M_{r}$. The magnetic field and $M_{ch}$ for Case 3 were 195 $\mu$A/m and 171 pH, respectively, at 34.5 dBm. Here, $\alpha_{ch}$ was modeled to 0.86. It proved that $V_{ch}$ was reduced by 14% in the underground corridor compared to the free space. Fig. 15(c) demonstrates the validity of the Rx-RM because $V_{ch}$ of the same system in Fig. 15(b) was amplified by 60 dBV to 75.9 mV$_{rms}$. Conversely, for Case 2, $V_{ch}$ was enhanced 1.67 times from Case 1, as shown in Fig. 15(d). Fig. 15(e) and 15(f) have the same structure except for the interval of the antenna’s radius. As expected, owing to $M_{r}$, $V_{ch}$ for the two Rx coils with no gap was 13.6% lower than that of Case 4. $M_{ch}$ was 274 pH at 34.5 dBm. $N_{r}$ was extracted to 18.7, which determined $M_{r}$ to be 96 $\mu$H. $M_{r}$ in Case 4 was 1.5 times higher than Case 1.

Fig. 16(a) shows the performances of $M_{ch}$ up to the distance of 25 m. The measured and modeled $M_{ch}$ at 25 m were 276 pH and 274 pH, respectively, with little difference at the operating frequency of 20.84 kHz. Thus, the slight difference value validates the empirical electrical model. As seen in Fig. 16(b), $V_{ch}$ for Case 4 was 171 $\mu$V$_{p}$ (1210 $\mu$V$_{rms}$) in a 25 m at the resonant frequency, giving almost the same result as measured in Fig. 15(f). Case 4 had three times higher voltage than Case 1, which was equal to the SNR of 9.5 dB. If $R_{pre}$ was replaced to 50 $\Omega$, $P_{r}$ could be modeled. $P_{r}$ for Case 4 was −66.2 dBm while $P_{r}$ for Case 4 was −76.6 dBm, marking the
.. figure:: figure15.png
   :alt: FIGURE 15. Induced voltage of (a) Case 1, (b) Case 3, (c) Case 3 through the Rx-RM, (d) Case 2 through the Rx-RM, (e) Case 4 with no interval through the Rx-RM, (f) Case 4 with an interval through the Rx-RM at a distance of 25 m in a real underground corridor.

.. figure:: figure16.png
   :alt: FIGURE 16. (a) Measured and modeled performances of $M_{ch}$ for Case 4 as a function of distance and comparison of modeled (b) $V_{ch}$ for Case 1 and Case 4 in an underground corridor as a function of frequency.

Improvement of 10.4 dB. From the circuit, $V_{ch}$ was modeled up to the range of 100 m, as shown in Fig. 17(a). When the noise baseline in the 3.2 kHz span was 2 $\mu$Vrms, the maximum distance in Case 4 within an SNR of 9.5 dB was 44.6 % longer than Case 1. Accordingly, the magnetic flux density $B_{out}$ was modeled because $B_{out}$ was considered as a performance index for a sensor node in MI-based communication. Fig. 17(b) shows the results. Case 4 at 34.5 dBm is capable of forming $B_{out}$ of 3.8 pT at a distance of 100 m. Moreover, according to the 10 pT requirement, Case 4 can cover a range of up to 72.3 m. The characteristics of a magnetic field channel can be evaluated by $PL$. Fig. 18 presents $P_r$ and path loss $PL$ of Case 1 and Case 4 as a function of distance. At 100 m, $P_r$ of Case 4 was $-102.1$ dBm and $P_r$ of Case 1 was $-112.7$ dBm. The $PL$ is described by the proposed model as

.. math::
   P_r(d_{ch}) = \frac{|V_{ch}|^2}{2 |Z_t+Z_r+R_{pre}|^2} \left| \frac{R_{pre}}{R_{pre}+Z_{tot}+Z_{rt}} \right|^2.

For Case 4, $PL$s at 25 m and 100 m distance were 100.7 dB and 136.6 dB, respectively, whereas for Case 1, $PL$s at 25 m and 100 m distance were 108.1 dB and 144.2 dB, respectively. Here, the decay trend for the two systems was similar even as the distance increases.

**B. INTER-LAYERS IN A BUILDING**

We evaluate whether a MI-based link system could be used in buildings with concrete walls due to practical necessity. Fig. 19(a) describes a diagram for evaluating the MI channel effect in a real building environment. $V_{ch}$ was investigated according to the floor spacing. There was no significant reduction in $V_{ch}$ in the single-floor spacing compared to the line-of-sight distance, but $V_{ch}$ reduced considerably in the two-layer interval. The measured data in the single-floor spacing is presented in Fig. 19(b). The induced voltage $V_1$ of ETS-6509 at 34.5 dBm had 327 uVrms. For Case 4, $V_{ch}$ was saturated to 1.75 Vrms due to the high gain of the Rx-RM, showing that the single-story concrete floor had little effect on the channel medium. However, it was confirmed...
that the two floors of concrete deteriorated $V_{ch}$ tremendously. Fig. 19(c) shows $V_{ch}$ between the second and basement floors of the building. At 34.5 dBm, the measured voltage $V_2$ was 19.1 mV_{rms} leading to SNR of 8.7 dB which was considered to a magnetic channel influence in the underground corridor of 70 m. Also, $V_1$ of ETS-6509 was to be 2.24 $\mu$V_{rms} with the noise baseline of 1 $\mu$V_{rms} in the span of 3.2 kHz. The characteristics and performances for Case 4 was listed in Table 1, where $f_{req}$ was 20.84 kHz. As a result, these data indicated that the MI-based transceiver could be implemented through interlays on the second floor or higher of the building only under high Tx power conditions of 34.5 dBm or higher. Table 2 shows the comparison results and characteristics of the existing technology for the MI-based sensor node. In the MI communication [10], [15], the connection technology of the sensor node is explained based on the technology using COTS products in the MHz band. On the other hand, in the

![Figure 17](image1.png)

**FIGURE 17.** Comparison of modeled (a) $V_{ch}$ and (b) $B_{out}$ for Case 1 and Case 4 under the condition of an underground corridor as a function of distance.

![Figure 18](image2.png)

**FIGURE 18.** Comparisons of modeled $P_r$ and $PL$ for Case 1 and Case 4 under the condition of an underground corridor as a function of distance.

![Figure 19](image3.png)

**FIGURE 19.** (a) Diagram for evaluating the MI channel effect in a real building environment and measured $V_{ch}$ for Case 4 (b) in the single-floor spacing and (c) between the second and basement floors.

**TABLE 1.** Characteristics and performances for case 4.

| Meaning | Value |
|---------|-------|
| $V_{ch}$ in a 25 m corridor at 34.5 dBm Tx | 121.3 $\mu$V_{rms} |
| SNR through the Rx-RM on the $V_{ch}$ | 46 dB |
| $P_r$ in a 25 m corridor at 34.5 dBm Tx ($R_{v_{ch}}= 50 \Omega$) | 239 pW |
| $PL$ in a 25 m corridor for the $P_r$ | 100.7 dB |
| Distance for 9.5 dB SNR at 34.5 dBm Tx (modeled) | 68.0 m |
| Distance for 10 pT at 34.5 dBm Tx (modeled) | 72.3 m |
| $B_{out}$ in a 100 m corridor at 34.5 dBm Tx (modeled) | 3.8 pT |
| $PL$ in a 100 m corridor at 34.5 dBm Tx (modeled) | 136.6 dB |
| SNR in two-story spacing at 34.5 dBm Tx | 8.7 dB |
application field of MI localization [2], [8], the detection method through the WUSNs based on the low-frequency 3-D antenna structure is described. Our technology (Case 4) confirmed the signal connection (SNR: 8.7 dB) with the low Tx power (34.5 dBm) at the interval of the second floor in the building. The summary of the contents developed through this paper is as follows.

1. A low-power, low-volume MI hardware platform, which maximized SNR for two-way connectivity, has been developed as the physical layer of WUSNs.
2. Since an empirical circuit model completely consisting of variables from the signal source to the output load has been developed, the characteristics of the real environment reflecting a magnetic channel were evaluated.
3. The MI-based transceiver has been deployed in a building with concrete walls (two-story space) in order to test and analyze the feasibility of the proposed hardware.

VII. CONCLUSION
This paper introduced a low-power, low-frequency MI-based transceiver that could be used in underground environments blocked by stairs or floors. In addition, we performed electrical modeling using empirical-based variables to reflect the boundary conditions of a real environment. Obviously, we found that signal attenuation occurs sharply when we measured between two layers rather than between one layer. Likewise, when there were barrier walls in the basement rather than the open corridor space, the signal also rapidly decreased. The results show that MI-based data link over certain intervals (e.g., 25 m or more) can be performed only when the Tx power level is generally above 30 dBm. Hence, our study confirmed that the proposed mobile low-frequency MI hardware system is suitable for extreme or emergency two-way communications in blocked underground environments.

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### TABLE 2. Comparison results and characteristics of the existing technology for the MI-based sensor nodes.

| Tx - Rx Structure | 1-1 circular [15] | multi-multi [10] | 3-3 cubic [8] | 3-3 sphere [2] | This work (Case 4) |
|------------------|------------------|------------------|----------------|----------------|------------------|
| Coil size (max)  | 66 mm (L)        | 40 mm (r)        | 300 mm (L)     | 104 mm (r)     | 250 mm (r)       |
| (L: length/c: radius) |                  |                  | 80 turns $f_0$ = 2.025 kHz |                  |                  |
| T/Rx part        | COTS             | 42 turns         | COTS, 29 turns  |                  |                  |
|                  | 13.56 MHz, 18mA  | $f_0 = 62$ MHz, 1 A | 125 kHz, 0.8 A |                  |                  |
| Range            | 12 m (indoor)    | 20 m             | 35 m           | 38 m           | 25 m (two floors) |
| Model            | FEM (theoretical) | FEM (theoretical) | Measurement     | Circuit (theoretical) | Circuit (empirical) |
| Application      | Test-bed         | Complex          | Underground     | Complex         | Underground       |
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