Channel Modeling for Metamaterial-Enhanced Underground Wireless Communications

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Abstract

Wireless communication is the prerequisite for the highly desired in-situ and real-time monitoring capability in underground environments, including oil reservoirs, groundwater aquifers, volcanos, among others. However, existing wireless communication techniques do not work in such environments due to the harsh transmission medium with very high material absorption and the inaccessible nature of underground environment that requires extremely small device size. Although Magnetic Induction (MI) communication has been shown to be a promising technique in underground environments, the existing MI system utilizes very large coil antennas, which are not suitable for deployment in underground. In this paper, we propose a metamaterial enhanced magnetic induction communication mechanism that can achieve over meter scale communication range by using millimeter scale coil antennas in the harsh underground environment. An analytical channel model for the new mechanism is developed to explore the fundamentals of metamaterial enhanced MI communication in various underground environments. The effects of important system and environmental factors are quantitatively captured, including the operating frequency, bandwidth, and parameters of metamaterial antennas, as well as permittivity, permeability, and conductivity of underground medium. The theoretical model is validated through the finite element simulation software, COMSOL Multiphysics.

Index Terms

Metamaterial antenna, Magnetic Induction, Underground Wireless Communications, Wireless Underground Sensor Networks.

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I. INTRODUCTION

The in-situ and real-time monitoring capability in underground environments can enable a wide variety of highly desired applications, such as enhanced oil recovery, accuracy groundwater quality management, smart underground power grid and pipeline, earthquake and volcano studies, among others. To establish such capability, wireless underground sensor networks (WUSNs) are proposed [1], [2] where sensor nodes are deployed underground and use wireless links to report the real-time measurements to administration center. Wireless communication among underground sensor nodes is the major challenge in WUSNs due to harsh transmission medium and inaccessible nature of underground environments [3], [4].

Most existing wireless communication techniques utilize the propagating electromagnetic (EM) waves radiated by dipole antennas, which have been proved to be effective when the underground device is larger than 10 cm and buried very close to the ground surface (less than 1 m) [5], [6], [7], [8]. In such case, the UHF EM wave-based signal can penetrate the short underground burial depth and utilize the air medium for effective propagation. However, the aforementioned underground applications require millimeter scale wireless sensor nodes that operate deep underground (up to 6000 ft beneath the ground surface in the case of oil reservoirs). On the one hand, only the underground medium with very high material absorption can be utilized deep underground. On the other hand, the millimeter scale dipole antenna can only effectively radiate SHF or even higher frequency EM waves, which have prohibitively high path loss in the underground medium. Hence, the EM wave-based wireless communication techniques do not work for the WUSNs with micro sensor nodes that are deployed deep underground.

The magnetic induction (MI) technique that utilizes magnetic coupling between coil antennas has been proposed as an alternative solution for wireless communications in underground [9], [10]. Compared with the EM wave-based solutions, MI communication does not rely on propagating EM waves but utilize magnetic field in the near field (the field within one wavelength from the transmitter). If low operating frequency with large wavelength is utilized (HF in this paper), MI communication encounters less absorption loss when penetrating the dense and lossy underground medium. In addition, MI communication does not suffer from the multipath fading problem in the EM wave-based solutions, since the randomly distributed soil, rock, and sand
components can dramatically affect the EM wave propagation but has almost no influence on the magnetic field as long as there is no magnetite (most cases in natural) [11], [9]. Moreover, the performance of coil antenna can be easily enhanced by adding a ferrite core or utilizing multiple turns without increasing the antenna size. Furthermore, if multiple coil antennas exist along the transmission medium between the two transceivers, the MI waveguide [11] is formed that can extend the wireless communication range [9], [12].

Despite the advantages, the overall path loss of the MI communication is still too high for practical deployment. Although the absorption loss of MI technique is low, the power of evanescent EM components in the near field attenuates much faster than propagating EM waves ($1/r^2$ vs $1/r^6$ where $r$ is the transmission distance). To address such problem, existing MI techniques utilize very large coil antennas (larger than 1 m in diameter) to mitigate the high attenuation rate problem [13]. However, the large antenna solution is infeasible for the micro WUSNs in the envisioned applications.

In this paper, we introduce metamaterials to enhance the underground MI communications, which can dramatically reduce the path loss and increase communication range in the harsh underground environments. The emergence of metamaterials has opened the door for manipulating EM fields and waves [14]. Metamaterials are artificial structures that are engineered to have properties that cannot be found in nature [15]. These artificial materials are made of carefully designed building blocks, or meta-atoms, which are much smaller than wavelength of incident EM wave. One of the important properties of metamaterials is the ability to enhance and manipulate evanescent components of EM waves. As discussed previously, the key problem of the existing underground MI communication technique is the fast attenuation rate of the evanescent EM component in the near field. In this paper, by equipping the underground sensor nodes with carefully designed metamaterial antennas (i.e. the MI coil antenna enclosed by a metamaterial sphere, as shown in Fig. 1), we show that efficient wireless signal transmission and reception can be realized in the harsh underground even the antenna size is in millimeter scale.

We for the first time analytically investigate the underground communication channel between the coil antennas enclosed by metamaterial spheres, as illustrated in Fig. 1. First, the EM field
that is generated by the MI coil and manipulated by the metamaterial sphere is analytically modeled in the underground environment. The magnetic field around the metamaterial enhanced coil antenna is shown to be increased by more than three orders of magnitude. Then based on the field analysis, the channel characteristics, including path loss, bandwidth, and channel capacity, are quantitatively derived and discussed for the MI point-to-point and MI waveguide communications. The results show that underground wireless communication range is signifi-
cantly increased by utilizing the metamaterial sphere. Moreover, much higher channel capacity is achieved in the same time. Finally, the effects of the practical metamaterial loss and dispersion are investigated. The theoretical model and discussions are validated through the widely used finite element simulation software, COMSOL Multiphysics [16].

The reminder of this paper is organized as follows. The related works are presented in Section II. Then, in Section III, the EM field intensity distribution around the proposed metamaterial enhanced antenna is derived. Base on the field analysis, the mutual induction and self induction for the MI communications are calculated. After that, in Section IV, the channel characteristics of metamaterial-enhanced MI communication including the point-to-point and MI waveguide communication are discussed. Finally, this paper is concluded in Section V.

II. RELATED WORK

The concept of the WUSNs is first proposed in [1], where the applications, system architecture, and research challenges are discussed. The channel models for EM wave-based wireless communication in soil medium are developed in [3], [5], [6]. To address the problem of extremely
high path loss of EM waves in pure underground environments, the hybrid WUSNs consisting of both aboveground data sink and shallowly buried underground sensor nodes are proposed in [7], [8], [17].

To reduce the high path loss due to the medium absorption, the magnetic induction (MI) technique has been widely used in many challenging environments, such as body area [18], [19], and underwater [20], [21]. The theoretical model of path loss and channel capacity for MI communication has been reported in [9], [22], which show the significant improvement over EM wave-based solution if only underground channel is available. However, MI communication system is highly directional and has limited channel capacity which limit its application. To address the challenges, a tri-directional coil antenna is proposed in [4], [19], [23] to increase the channel capacity and create an omnidirectional receiving antenna. In order to extend communication range, the MI waveguide presented in [11] is adopted. The theoretical system model for MI waveguide in underground environments is provided in [9] and the numerical simulation shows that the communication range can reach tens of meters when the diameter of the coil antenna is 30 cm. Another way to extend the MI communication range is to use very large coil antenna. In [13], a MI-based through-the-earth communication system is developed for mine disaster rescue, where the wireless device is bigger than a human. However, all the above MI techniques cannot realize the underground wireless communications among millimeter scale sensors, which are essential to the envisioned applications in this paper.

Metamaterials proposed in [24] can manipulate the EM fields by adjusting the value of the dielectric permittivity and magnetic permeability of the material to nearly any value, including positive, negative, or even near-zero values [14], [25]. It has been shown that metamaterials can greatly improve the efficiency of magnetic induction-based wireless energy transfer [26], [27]. The experiment in [26] shows that the efficiency of wireless power transmission can be as high as 47% by using metamaterial. While without using metamaterial the efficiency is only 17%. The corresponding theoretical model is developed in [27] where the loss of metamaterials is taken into consideration. In [28], by using a small metamaterial slab with $-1$ permeability at 63.87 MHz, the sensitivity of magnetic resonance imaging (MRI) is significantly increased. However, all above works consider the strong coupling mode [29], where two coils are very close to each
other. In underground wireless communications, to enlarge the coverage range, we consider the loose coupling to achieve longer transmission distance. Moreover, due to the different objectives, the system bandwidth for energy transfer and MRI is too small for wireless communications.

Besides wireless energy transfer and MRI, metamaterials have also been used in antenna design for the EM wave-based terrestrial communication systems. In [30], [31], metamaterials are introduced to miniaturize the antenna in portable devices. In [32], [33], [34], [35], the radiation of metamaterial-enhanced electrical dipole antenna are theoretically investigated. However, all the above antennas operate at UHF band that are not suitable in underground. Such antennas lose their efficiency if HF band signals are used.

As discussed in Section I, MI communication is a promising candidate solution in underground environments if the near field can be enhanced by the metamaterials. To our best knowledge, this is the first paper which analyzes the channel characteristics of metamaterial enhanced MI communications in underground environments.

III. METAMATERIAL-ENHANCED MUTUAL INDUCTION

In order to capture the channel characteristics of metamaterial enhanced MI communication, in this section, we first analyze the EM field (especially the magnetic field) intensity distribution around the transmitting antenna and the receiving antenna. We validate the theoretical result by matching it with the FDTD simulation results obtained by COMSOL Multiphysics. Then self induction of each metamaterial enhanced coil antenna and the mutual induction between the transceivers can be obtained, which are the key parameters in MI communications.

Generally, there are three kinds of structures utilizing metamaterials in antenna design: the rectangular structure utilizes a metamaterial slab in between the transmitter and receiver [27]; the cylindrical structure utilizes a metamaterial cylinder to surround the transceivers [32]; and the spherical structure uses a metamaterial sphere to enclose the antenna [33], [34]. For underground MI communications in this paper, we adopt the spherical structure to create an omnidirectional coverage. The spherical structure is also favorable for deployment. Hence, the metamaterial enhanced MI coil antenna considered in this paper is illustrated in Fig. I: the single-turn MI coil with radius of $a$ is enclosed by a metamaterial sphere with and internal radius of $r_1$ and
external radius of $r_2$. We define the space inside the metamaterial sphere as the first layer, the metamaterial layer as the second layer, and the space outside the sphere as the third layer. The MI coil is located at the center of first layer. In this section, we first derived the field distribution in each layer by using spherical wave equations. Then the mutual induction and self induction can be calculated.

A. Field Intensity Analysis

Since this paper focuses on the field enhancement due to the metamaterial sphere, we consider a basic single-turn coil without ferrite core. The system performance can be improved by simply increasing the number of turns and adding the ferrite core, which is easy to model but trivial to investigate here. The intensity distribution of the electromagnetic field generated by the coil in the space without the metamaterial sphere can be expressed in terms of spherical waves,

\[
\begin{align*}
\vec{H}_r &= -\frac{jk^2a^2NI_0 \cos \theta}{2d} h_1^{(2)}(kd) \vec{r}; \\
\vec{H}_\theta &= \frac{jk^2a^2NI_0 \sin \theta}{4d} \left[ h_1^{(2)}(kd) + kdh_1^{(2)'}(kd) \right] \vec{\theta}; \\
\vec{E}_\phi &= \frac{k^3a^2\eta NI_0 \sin \theta}{4} h_1^{(2)}(kd) \vec{\phi}; \\
\vec{H}_\phi &= 0; \vec{E}_r = 0; \vec{E}_\theta = 0,
\end{align*}
\]

where $k$ is the wavenumber, $N$ is the number of turns, $I_0$ is the current, $d$ is the distance from the origin, $\eta$ is the wave impedance, and $h_1^{(2)}(kd)$ is the spherical Hankel function of the second kind and order 1.

1) Field Analysis around transmitting antenna: We first investigate the EM field around a transmitting antenna with metamaterial sphere. As shown in Fig. [1] the coil is located at the center of the metamaterial sphere. According to the spherical vector wave functions [33], [36], the wave should be standing in the first layer and second layer. Since at the origin, the field strength cannot be infinity, we use the spherical Bessel function of the first kind. Moreover, the wave outside the sphere should be a traveling wave. Hence, we use the Hankel function of the second kind to model it. Since the transmission medium is underground soil, the conductivity should be taken into account. Therefore, in the third layer, we add the loss coefficient $e^{-\delta}$. Then the unknown magnetic fields in each layer of the metamaterial enhanced antenna can be
expressed by

1\(^{st}\) layer: \[\begin{align*}
\vec{H}_{r1} &= \frac{-2j\cos\theta}{\omega\mu_1 d} A_t j_1(k_1 d) \vec{r}, \\
\vec{H}_{\theta 1} &= \frac{j\sin\theta}{\omega\mu_1 d} A_t [j_1(k_1 d) + k_1dj_1'(k_1 d)] \vec{\theta};
\end{align*}\]

2\(^{nd}\) layer: \[\begin{align*}
\vec{H}_{r2} &= \frac{-2j\cos\theta}{\omega\mu_2 d} [B_t j_1(k_2 d) + C_t y_1(k_2 d)] \vec{r}, \\
\vec{H}_{\theta 2} &= \frac{j\sin\theta}{\omega\mu_2 d} \{B_t [j_1(k_2 d) + k_2dj_1'(k_2 d)] \\
&\quad + C_t [y_1(k_2 d) + k_2dy_1'(k_2 d)]\} \vec{\theta};
\end{align*}\]

3\(^{rd}\) layer: \[\begin{align*}
\vec{H}_{r3} &= \frac{-2j\cos\theta}{\omega\mu_3 d} D_t h_1^{(2)}(k_3 d)e^{-\frac{d-r_2}{\delta}} \vec{r}, \\
\vec{H}_{\theta 3} &= \frac{j\sin\theta}{\omega\mu_3 d} D_t \left[h_1^{(2)}(k_3 d) + k_3 dh_1^{(2)'}(k_3 d)\right] e^{-\frac{d-r_2}{\delta}} \vec{\theta};
\end{align*}\]

where \(A_t, B_t, C_t, \) and \(D_t\) are the unknown coefficients; \(j_1(kd)\) is the spherical Bessel function of the first kind and order 1, \(y_1(kd)\) is the spherical Neumann function of order 1, \(\delta\) is the skin depth of the underground medium which can be expressed by

\[\delta = \frac{1}{\omega \sqrt{\frac{\mu_\epsilon}{\mu \epsilon}} \left(\sqrt{1 + \frac{\sigma^2}{\omega^2 \epsilon^2}} - 1\right)},\]

where \(\sigma\) is the conductivity of soil. According to Maxwell equations, the normal component of the magnetic flux \((\vec{B})\) and the tangential component of the magnetic field \((\vec{H})\) should be continuous at the boundary. Thus by enforcing the boundary conditions, we let

\[\begin{align*}
\vec{\Phi}_{r 0} + \mu_1 \vec{H}_{r 1} &= \mu_2 \vec{H}_{r 2}; \\
\vec{\Phi}_{\theta 0} + \mu_1 \vec{H}_{\theta 1} &= \mu_2 \vec{H}_{\theta 2}; \\
\mu_2 \vec{H}_{r 2} &= \mu_3 \vec{H}_{r 3}; \\
\vec{H}_{\theta 2} &= \vec{H}_{\theta 3},
\end{align*}\]

where \(\vec{\Phi}_{r 0}\) and \(\vec{\Phi}_{\theta 0}\) are the excitation vectors. We can obtain the unknown coefficients by,

\[U_t = S_t^{-1}\Phi_t,\]
where \( U_t' = [A_t, B_t, C_t, D_t] \),

\[
S_t = \begin{pmatrix}
\frac{j_1(k_1 r_1)}{\mu_1} & -\frac{j_1(k_2 r_1)}{\mu_2} & -\frac{y_1(k_2 r_1)}{\mu_2} & 0 \\
\frac{j_1(k_1 r_2)}{\mu_2} & \frac{y_1(k_2 r_2)}{\mu_2} & -\frac{h_1^{(2)}(k_3 r_2)}{\mu_3} & 0 \\
0 & 0 & 0 & 0
\end{pmatrix},
\]

and

\[
\Phi_t = \begin{pmatrix}
-\frac{\omega^3 \mu_1^2 \epsilon_1 a^2 N I_0}{4} h_1^{(2)}(k_1 r_1) \\
-\frac{\omega^3 \mu_1^2 \epsilon_1 a^2 N I_0}{4} [h_1^{(2)}(k_1 r_1) + k_1 r_1 h_1^{(2)'}(k_1 r_1)] \\
0 \\
0
\end{pmatrix}.
\]

Then, by substituting the unknown coefficients into (2), the intensity distribution of magnetic field around the coil antenna with a metamaterial sphere can be obtained. Based on the field intensity, the self inductance can be updated by

\[
L = \frac{\Phi_1}{I_0} \approx -j \frac{2 A_t N \pi a^2 \sqrt{\mu_1 \epsilon_1}}{3 I_0},
\]

where \( \Phi_1 \) is the magnetic flux through the transmitting coil.

2) Field Analysis around receiving antenna: The outgoing wave from the transmitter is regarded as incoming wave for the receiver. To derive the concise formulation, we utilize another spherical coordinate whose origin is located at the center of the receiver. As shown in Fig. 2, \( O_1 A \) represents \( \vec{H}_{r_1} \) while \( AC \) represents \( \vec{H}_{\theta_1} \). Then, we decompose \( \vec{H}_{r_1} \) and \( \vec{H}_{\theta_1} \) into the directions of \( O_2 A \) and \( BA \), which represent the directions of \( \vec{H}_{r_2} \) and \( \vec{H}_{\theta_2} \) at the receiver. Since the transmission distance \( d \) is much larger than the size of the sphere, i.e. \( d >> r_2 \), and the
two coils are placed coaxially, $\vec{H}_{\theta 1}$ is much smaller than $\vec{H}_{r 1}$. Thus we only consider $\vec{H}_{r 1}$ here. Then,

$$
\begin{align*}
-\vec{H}_{r 2} & \simeq \vec{H}_{r 1} \cdot \cos(\pi - \theta_1 + \theta_2); \\
\vec{H}_{\theta 2} & \simeq \vec{H}_{r 1} \cdot \sin(\pi - \theta_1 + \theta_2).
\end{align*}
$$

Due to the relatively long distance between the transmitter and the receiver, $\theta_1 \simeq \pi$. Therefore, (9) can be simplified to $\vec{H}_{r 2} \simeq -\vec{H}_{r 1} \cdot \cos \theta_2$, and $\vec{H}_{\theta 2} \simeq \vec{H}_{r 1} \cdot \sin \theta_2$.

It should be noted that the transmitting coil and receiving coil are not necessarily coaxially placed in real underground deployment. The system performance may become unreliable due to the rotation of the coil antennas. This problem can be solved by the tri-directional coil antenna where three perpendicular coils are mounted together, as shown in the right figure in Fig. 1. The performance of the tri-directional coil antenna has been analyzed by our previous work [4] and other related work [19], [23]. Since this is out of the scope of this paper, we only consider the coaxial transmitter and receiver coils.

There are two kinds of waves outside the receiver: the incoming wave from the transmitter and the reflected wave from the boundary between Layer 2 of the receiver and the underground environment. Since the incoming wave is a function of the spherical Hankel function of the second kind and order 1, the reflected wave is a function of the spherical Hankel function of the first kind and order 1. The reflected magnetic field can be expressed by

$$
\begin{align*}
\vec{H}_{r 4} &= \frac{-2j\cos \theta}{\omega \mu_0 d'} D_r h_1^{(1)}(k_3 d') e^{-\frac{d-r_2}{\sigma}} \vec{r}, \\
\vec{H}_{\theta 4} &= \frac{j \sin \theta}{\omega \mu_0 d'} D_r \left[ h_1^{(1)}(k_3 d') + k_3 d' h_1^{(1)'}(k_3 d') \right] e^{-\frac{d-r_2}{\sigma}} \vec{\theta},
\end{align*}
$$

where $D_r$ the coefficient of the reflected wave, $d'$ is the distance from the origin of receiver.

The format of the EM field intensity distribution in the first and second layer around the receiver antenna is the same as that around the transmitter. However, the coefficients in the format are different and need to be identified. We denote the unknown coefficients are $A_r$, $B_r$, and $C_r$ for the first layer and second layer around the receiver. By substituting $A_t$, $B_t$, and $C_t$ by $A_r$, $B_r$, and $C_r$ in (2) respectively, we can obtain the magnetic field distribution around the receiver antenna. Then based on (9) and the boundary conditions, we can determine $A_r$, $B_r$, and
\[ U_r = S_r^{-1} \Phi_r, \]  
(11)

where \( U_r' = [A_r, B_r, C_r, D_r] \),

\[
S_r = \begin{pmatrix}
-j_1(k_1 r_1) & j_1(k_2 r_1) & y_1(k_2 r_1) & 0 \\
-j_1(k_1 r_1) + k_1 r_1 j_1'(k_1 r_1) & j_1(k_2 r_1) + k_2 r_1 j_1'(k_2 r_1) & y_1(k_2 r_1) + k_2 r_1 y_1'(k_2 r_1) & 0 \\
0 & -2j_1(k_2 r_2)_{\mu_2} r_2 & -2y_1(k_2 r_2)_{\mu_2} r_2 & 2h_1^{(1)}(k_3 r_2)_{\mu_3} r_2 \\
0 & -j_1(k_2 r_2)_{\mu_2} + k_2 r_2 j_1'(k_2 r_2)_{\mu_2} r_2 & y_1(k_2 r_2)_{\mu_2} + k_2 r_2 y_1'(k_2 r_2)_{\mu_2} r_2 & h_1^{(1)}(k_3 r_2)_{\mu_3} r_2 \\
\end{pmatrix},
\]
(12)

and

\[
\Phi_r = \begin{pmatrix}
0 \\
0 \\
\frac{2D_r}{d - r_2} h_1^{(2)}(k_3(d - r_2)) e^{-d-r_2} \\
\frac{2D_r}{\mu_3(d - r_2)} h_1^{(2)}(k_3(d - r_2)) e^{-d-r_2} \\
\end{pmatrix},
\]
(13)

where \( d \) is the distance from the origin of transmitter. Once \( A_r, B_r, \) and \( C_r \) are determined, the magnetic field intensity distribution around the coil antenna with the metamaterial sphere at the receiver side can be expressed in the same format given in (2).

Based on the magnetic field distribution around the receiving coil antenna, the mutual induction between the transmitting coil and the receiving coil can be derived. According to [38], the mutual inductance between the two coils can be found by utilizing the magnetic flux through receiving coil and the current in transmitting coil, which is

\[
M = \Phi_2 \frac{I_0}{I_0} \simeq -j \frac{2A_r N \pi a^2 \sqrt{\mu_1 \varepsilon_1}}{3 I_0},
\]
(14)

where \( \Phi_2 \) is the magnetic flux generated by transmitting coil goes through receiving coil. It should be noted that a term \( \cos \theta \) is neglected here since the distance between the two coils is long enough.

**B. Numerical Results and Simulation**

In order to validate the above theoretical modeling, we utilize the widely used Comsol Multiphysics to simulate the system. The AC/DC module and axis symmetry model are used here. The geometry of the simulation model is shown in Fig. 3. According to [33], in order to
make the transmitter resonant at dipolar mode to achieve the highest efficiency, we let \( r_1 = 0.01 \text{m} \) and \( r_2 = 0.025 \text{m} \). The diameter of the single-turn coil is \( 0.01 \text{m} \) and the input current is 1 A. The operating frequency is \( 10 \text{MHz} \). We use the typical environmental parameters of the air and underground soil medium. While the air has 0 conductivity, the conductivity of the soil is \( 0.005 \text{S/m} \). The relative permeability inside the sphere, in the metamaterial sphere and outside the sphere are 3, -1 and 1, respectively. While the corresponding relative permittivity are 3, -1 and 2. The right figure in Fig. 3 shows the 2D section plane of the whole simulation region, which includes the underground environment as well as the transmitter and the receiver. The left figure in Fig. 3 shows the zoom-in illustration of the coil enclosed by the metamaterial sphere. The distance between transmitter and receiver is 2.5 m which is 500 times larger than the coil’s radius. As shown in Fig. 3 the transmitter is located at upper side and the receiver is located at 2.5 m away. The infinite element domain is utilized to stretch the simulation domain to infinity.

Fig. 4 gives the magnetic fields around the transmitter and receiver with and without the metamaterial sphere, as a function of the distance between the transmitter and receiver. Both the theoretical results and the FDTD simulation results are provided. To guarantee the fairness, for the condition without metamaterial sphere, we use a sphere with the same size but has positive dielectric parameters. First of all, Fig. 4 shows that our theoretical model agrees very well with the simulation results at both transmitter side and receiver side, which validate the accuracy of the field analysis in this section. Second, as shown in Fig. 4 with the help of metamaterial
Fig. 4. Intensity distribution of magnetic field.

(a) Magnetic field at the transmitter.  
(b) Magnetic field at the receiver

Fig. 5. Magnetic field density at transmitting end.

sphere, the density of magnetic fields around both transmitter and receiver are greatly enhanced compared with the cases without metamaterial sphere. For example, at the receiving side, the magnetic field is around $0.12\, \text{A/m}$ with metamaterial sphere. While without metamaterial sphere,

Fig. 6. Magnetic field density at receiving end.
the field is around $3.69 \times 10^{-7} \text{A/m}$ which is about $3 \times 10^5$ times smaller than the case with metamaterial sphere, which proves the significant improvement by the metamaterial enhancement. Moreover, Fig. 5 and Fig. 6 show the intensity distribution of magnetic field at transmitting end and receiving end visually. The differences between the cases with metamaterial and without metamaterial are obvious. It is interesting to note that we observe a big jump of the field intensity at the boundary between the first layer and second layer. Mathematically, the reason is that the Neumann function approximates infinity when $d$ is very small. The field density gets very large since the second layer has the Neumann function and distance is small from origin.

The above results are based on the assumption that metamaterials are ideal and do not introduce any loss. However, due to the intrinsic property of the metallic atom, metamaterial has a certain level of loss. To investigate the impact of such loss, other than the ideal case, we also consider a practical case where the metamaterial has relatively high loss, i.e. $\mu = (-1 - 0.1j)\mu_0$ and $\varepsilon = (-1 - 0.1j)\varepsilon_0$. Fig. 7 shows the theoretical and simulated magnetic field around the transmitter and receiver when the metamaterial loss is taken into consideration. We can see that there is an obvious drop of the density of magnetic field at both transmitter and receiver compared with the ideal case. Around the receiver, the magnetic field is $6.8 \times 10^{-4}$, which is about 1860 times larger than the case without metamaterial sphere. Even it is much smaller than the ideal case, there is still a significant improvement by using the metamaterial enhanced coil antenna.
IV. Channel Characteristics of Metamaterial Enhanced MI Communications

In this section, based on the field analysis in Section III, the channel model of the metamaterial enhanced MI communication is developed for the underground environments. The channel characteristics for Point-to-Point and MI waveguide communication are presented and compared with the original MI communication system in [9]. The challenges for practical implementation are also discussed.

A. Channel Characteristics for Point-to-point Communications

For the point-to-point MI communication, the equivalent circuit model is shown in Fig. 8 which is similar to the model given in [9]. The only difference is that the coil antennas are enclosed by the metamaterial sphere here. In the circuit model, the resistance, and inductance of the coil are $R_C$ and $L$, respectively. The source generates an RMS voltage $V$ with source resistance $R_S$. Due to metamaterial sphere, the self inductance and mutual inductance are complex values. Since we use the same size coil and metamaterial sphere, all the transmitter, receiver and relay have the same self inductance. We use $L_r$ to denote the real part of self inductance and $L_i$ to denote the imaginary part. In the equivalent circuit, the capacitor is utilized to compensate the real part of the self inductance which is $C = 1/(\omega_0^2 L_r)$, where $\omega_0$ is the resonant frequency.

In this subsection, we compare three cases, the original MI system where the coil has 10
turns, the metamaterial enhanced MI system where the coil has single turn and 10 turns. Firstly, we derive the theoretical model for path loss, 3 dB bandwidth, and channel capacity. Then we numerically show the great improvement by utilizing metamaterial sphere.

1) Path loss: The path loss for point-to-point MI communication have been developed in [9] and [22]. However, the model is not applicable for metamaterial enhanced MI communication because the self inductance is a complex number here. Thus, we need to recalculate the path loss. We assume the coils are far enough from each other which means they are loosely coupled. Then, the received power and transmitted power can be expressed by

\[ P_r = \frac{1}{2} R_L |I_2|^2 = \frac{1}{2} \frac{R_L \omega^2 |M|^2 |V|^2}{|Z_1|^2 |Z_2|^2}, \]

\[ P_t = \frac{1}{2} \text{Re}(Z_1) \frac{|V|^2}{|Z_1|^2}, \]

where \( Z_1 = R_S + R_C + j\omega (L_r + jL_i) + 1/j\omega C \) and \( Z_2 = R_L + R_C + j\omega (L_r + jL_i) + 1/j\omega C \); \( L_r \) and \( L_i \) are the real part and the imaginary part of the metamaterial enhanced self inductance, which are developed in [8]; \( M \) is the metamaterial enhanced mutual coupling, which is developed in [14]. Then the path loss can be updated by

\[ L_{p2p}(d) = -10 \log \left( \frac{P_r}{P_t} \right) = -10 \log \left( \frac{R_L \omega^2 |M|^2}{\text{Re}(Z_1)|Z_2|^2} \right). \]

In the numerical evaluation, we use the same configuration for the coil antennas as previous discussions. The source output impedance is \( R_S = 10\Omega \) and the load resistance is \( R_L = 1\Omega \). We use AWG 26 wire for the coil, and the single-turn coil’s resistance is \( R_C = 4.2 \times 10^{-3}\Omega \). The numerical result is shown in Fig. [9] It is obvious that even the single turn coil with metamaterial
sphere can dramatically outperform the 10-turn coil without metamaterial sphere. Additionally, by utilizing multiple turns, the path loss can be further reduced. As the result shows, metamaterial can improve the mutual induction between two coils. Thus, power can be efficiently transmitted through magnetic coupling. It is worth noting that here we use relatively large source impedance and load impedance when compared with the coil’s resistance, because it is challenging to design a low output impedance circuit in practice. Meanwhile, at the receiver, it would be favorable to tolerant high load in real implementation. If we can reduce both of them, the path loss would be even lower.

2) Bandwidth: There exists tradeoffs between the low path loss and high bandwidth in MI communications. On the one hand, in order to achieve low path loss, we need high Q circuit that resonants at a single frequency. On the other hand, a broad bandwidth is desired for wireless communication. Thus, it’s challenging to achieve low path loss and broad bandwidth simultaneously for MI communication. The metamaterial sphere in fact further increases the circuit Q value, which further reduces the system bandwidth. Therefore, bandwidth analysis of the metamaterial enhanced MI communication is of great importance. In the Appendix, we prove that when the metamaterial sphere is small enough (the case in this paper), the mutual inductance and self inductance are independent of frequency. Thus, we consider that \( M \), \( L_r \) and \( L_i \) are not functions of frequency in the rest of this paper. Then, the 3 dB bandwidth can be derived from

\[
\begin{align*}
\frac{RL\omega^2|M|^2}{(R_S+R_C-\omega L_i)/(R_L+R_C-\omega L_i)} &= \frac{1}{2},
\end{align*}
\]

When \( \mu = -1 \) and \( \epsilon = -1 \), the imaginary part of \( L \) is very small (\(-1.36 \times 10^{-11}\) for single turn metamaterial enhanced coil) so that we can neglect it on the numerator. Then the bandwidth can be expressed by

\[
B_{1/2} \approx \frac{1}{\pi} \sqrt{\frac{2x_2 L_r/C - x_1 + \sqrt{x_1^2 - 4x_1x_2 L_r/C + 4x_2/C^2}}{2(x_2 L_r^2 - 1) - \frac{\omega_0}{\pi}},}
\]

where \( x_1 = \frac{\omega_0^2 (R_S+R_C)(R_L+R_C)^2}{2(R_S+R_C-\omega_0 L_i)(R_L+R_C-\omega_0 L_i)^2} \), and \( x_2 = \frac{\omega_0^2 (R_S+R_C)}{2(R_S+R_C-\omega_0 L_i)(R_L+R_C-\omega_0 L_i)^2} \).

By using the same configuration as before, the numerical result of the bandwidth in different MI communication systems are shown in Fig. 10 by using path loss in (17). The single-turn coil
Fig. 10. Bandwidth for point to point communication.

Fig. 11. Channel capacity of MI communication.

with metamaterial sphere has the broadest bandwidth. The 10-turn coil with metamaterial sphere has the narrowest bandwidth but the lowest path loss. The 10-turn coil without metamaterial sphere has highest path loss, whose bandwidth is in the middle of the previous two cases.

Moreover, by using (19), the calculated bandwidth for the above three cases are 223 kHz, 320 kHz and 3.7 kHz, respectively, which match the results in Fig. 10. Even using approximation, (19) is still accurate enough.

3) Channel capacity: According to Shannon-Hartley theorem, the channel capacity can be expressed by:

\[ C = B_{p2p} \log_2 \left( 1 + \frac{P_t L'_{p2p}(d)}{N} \right), \]

where \( P_t \) is the transmission power, \( L'_{p2p}(d) \) is the path loss derived in (17) in plain scale, and \( N \) is the noise level. According to [5], the noise power in underground environment is around -100 dBm and the transmission power is set as 1 dBm.

The numerical result is shown in Fig. 11. Due to the high path loss, it’s almost impossible for a
small coil to communicate over 1 m. While, by utilizing metamaterial sphere, the communication range can be around 10 meters while the data rate can be up to 100 kbps. It is indeed impressive since the diameter of the MI coil is only 0.01 m. The communication range over the radius of the coil can be $d/a = 2000$ which is much larger than the original MI system whose $d/a = 40$. Moreover, we notice that, in the near field when path loss is low enough, the bandwidth is more important. When the distance is relatively large, the path loss becomes very high. Under this situation, the path loss is more important than bandwidth.

**B. Channel Characteristics for MI Waveguide**

Since the communication range of the point-to-point MI system is limited, the MI waveguide can be adopted to extend the range [9]. The MI waveguide consists of relay coils which are simple devices without any energy source and signal processing. Due to the passively relay effect, the path loss can be dramatically reduced. In this subsection, we investigate the metamaterial enhancement in the MI waveguide communications. The equivalent circuit model of MI waveguide is shown in Fig. 12. Similar to the point-to-point case, the difference here is the metamaterial sphere that encloses each coil. We consider that there are $n$ coils, including one transmitter, one receiver and $n-2$ relay coils. We assume all the coils have the same inductance $L$ and resistance $R_C$. All the coils are loaded with a well-designed capacitor $C$ to tune the circuit. We first develop the theoretical model for the path loss, bandwidth and channel capacity of the MI waveguide. Then, its performance is compared with with the original MI waveguide developed in [9].
1) Path loss: We consider the mutual induction between two neighbor coils that is due to the loose coupling so that the reflected impedance can be neglected. Among the coaxially aligned $n$ coils, the first one is the transmitter; the last one is the receiver; and all other coils are relays. Similar to (17), the path loss can be expressed by

$$L_w(d) = -10 \log \left( \frac{P_r}{P_t} \right) = -10 \log \left( \frac{\omega^{2n-2} |M|^{2n-2} R_L}{|Z_n|^2 |Z|^{2n-4} \text{Re}(Z_1)} \right), \quad (21)$$

where $Z = R_C + j\omega(L_r + jL_i) + 1/j\omega C$, and $Z_n$ is the same as the $Z_2$ in (17). It should be noted that the path loss of MI waveguide is highly dependent on the term $\frac{\omega^2 |M|^2}{|Z|^2}$. In order to achieve low path loss, the waveguide need to be designed so that this factor is maximized.

The numerical results of the path loss of MI waveguide is shown in Fig. 13. For fair comparison, we let the different MI systems in Fig. 13 have similar $\frac{\omega^2 |M|^2}{|Z|^2}$. As a result, the interval between two adjacent relay coils are different. In particular, the adjacent coil interval for the three cases are 0.057 m (without metamaterial sphere), 2.25 m (single-turn with metamaterial sphere), and 3.25 m (10-turn with metamaterial sphere). As shown in Fig. 13 by using metamaterial, the path loss of the MI waveguide can be greatly reduced.

2) Bandwidth: Since all the relay coils are the same, they have the same self inductance and resistance. Then similar to (19), the 3 dB bandwidth can be expressed by using (21):

$$B_w \simeq \sqrt{\frac{2xL_r/C + \sqrt{(2xL_r/C - xR_C^2)^2 + 4x(1-xL_r^2/C^2)}}{2\pi^2(xL_r^2 - 1)} - \frac{\omega_0}{\pi}}, \quad (22)$$

where $x = \omega_0^2 n^{-1} \sqrt{\frac{1}{2(R_L + R_C)(R_C)^{2n-3}}}$. From (22) we can see the bandwidth of the MI waveguide is determined by the number of relay coils. The more relay coils, the narrower bandwidth.
The numerical result is shown in Fig. 14. The MI waveguide bandwidth after the metamaterial enhancement is generally smaller than the original MI waveguide, which is due to the higher Q value brought by the metamaterial sphere. More number of turns can further reduce the bandwidth.

3) Channel capacity: Similarly, by utilizing (20), (21), and (22), the channel capacity for MI waveguide can be derived. The numerical result is given in Fig. 15. As shown in the figure, it’s almost impossible for the 10-turn coil without metamaterial sphere to communicate over 10 m with any low data rate. By utilizing metamaterial sphere, the communication range for a MI waveguide with single-turn coils can be extended to almost 80 m. Considering that the diameter of each MI coil is only 0.01 m, this communication range is extremely large. The distance to size ratio \(d/a = 16000\), which is much larger than the original MI waveguide. If we use multiple turns, the range can be further extended.
C. Loss and Dispersion Effect

It should be noted that the above analysis is based on two ideal assumptions: the ideal metamaterial without any loss is used; and the metamaterial has the constant value of the negative permittivity and permeability. However, due to the intrinsic property of metamaterial, the loss and dispersion effect should be taken into account when evaluating the performance of the enhanced MI communication. In this subsection, the loss effect on path loss/bandwidth and dispersion effect on bandwidth are discussed.

1) Loss effect on channel characteristics: As shown in Fig. 7, the intensity of magnetic field drops dramatically due to the loss of metamaterial sphere. As a result, the mutual induction between coils becomes smaller. The loss effect on point-to-point MI communication’s path loss is shown in Fig. 16. We consider three cases, $\mu/\mu_0 = \epsilon/\epsilon_0 = -1 - 0.001j$, $\mu/\mu_0 = \epsilon/\epsilon_0 = -1 - 0.01j$, $\mu/\mu_0 = \epsilon/\epsilon_0 = -1 - 0.1j$. The number of turns is 10. As shown in Fig. 16, the communication range reduces to less than 5 m in the worst case. However, it is still much better than the coil without metamaterial sphere. The same situation will happen in the waveguide. The loss in the metamaterial sphere would cause high path loss between each pair of coils. Thus total path loss through the waveguide will be higher than before and the interval between two relay coils should decrease to maintain a satisfied performance.

If we consider the metamaterial loss, the resonant frequency will also shift. The self inductance $L = L_r + jL_i$ changes as the loss varies. We find that when loss is high, the absolute value of $L$ is small, and the ratio of $L_i/L_r$ is high. Thus, as loss increases, the quality factor becomes...
smaller and the resistance caused by $\omega L_i$ is much larger than $j\omega L_r$. Even we use a capacitor to compensate the real part of $L$, $\omega L_i$ can change with frequency which means the resistance now is a function of frequency. Thus, as we can see in Fig. 17, the resonant frequency shifts. The higher the loss, the more resonant frequency shifts. For very low loss $\mu/\mu_0 = \epsilon/\epsilon_0 = -1 - 0.001j$, the shift is not obvious. For higher loss, we can clearly see the shift, where the resonant frequency changes from 10 MHz to 8 MHz and 1.5 MHz. For MI waveguide, since all the coils have the same resonant frequency, the MI waveguide should resonant at the same frequency as its component. Hence the resonant frequency of MI waveguide has the same shift effects due to the metamaterial loss as the point-to-point case.

It should be noted that current fabrication technology can make loss lower than $0.1j$ [26]. The high loss case shown in this paper can be considered as the worst case.

2) Dispersion effect on channel characteristics: The metamaterial can achieve the desired negative permittivity and permeability at a certain frequency. It cannot provide a constant negative value at any frequency. Thus, in reality, the dispersion effect should be taken into consideration. In this paper, we use Drude model [15] to capture this effect. The Drude model takes the form:

$$
\epsilon(\omega) = \epsilon_0 \left( 1 - \frac{\omega_{pe}^2}{\omega (\omega - j\Gamma_e)} \right),
$$

$$
\mu(\omega) = \mu_0 \left( 1 - \frac{\omega_{pm}^2}{\omega (\omega - j\Gamma_m)} \right),
$$

where $\omega_{pe}$ and $\omega_{pm}$ are the electric and magnetic plasma frequencies, and $\Gamma_e$ and $\Gamma_m$ are the electric and magnetic collision frequencies. We consider two cases, one is $\Gamma_e = \Gamma_m = 10^{-2}\omega_0$ which is low loss, and the other one is $\Gamma_e = \Gamma_m = 10^{-1}\omega_0$ which is high loss. In order to let both
of permeability and permittivity be $-1$ at $\omega_0 = 10 MHz$, $\omega_{pe}$ and $\omega_{pm}$ should be $8.85 \times 10^7$ for low loss and $8.93 \times 10^7$ for high loss. The numerical result is shown in Fig. 18. Even consider this effect, the bandwidth is still good enough for the MI communication. Also we can clearly see the shift of resonant frequency. For MI waveguide, the dispersion effect is similar. Since all the coils have the same configuration and resonant frequency, the total dispersion effect is proportional to above result.

V. CONCLUSION

The in-situ and real-time information is highly desired in the inaccessible underground environments. Due to the harsh medium with high absorption and micro-sized antenna requirement, existing wireless communication techniques do not work. In this paper, we propose a metamaterial enhanced MI communication system for the underground wireless communication. A metamaterial sphere is utilized to enclose the coil antenna to enhance the near field that MI communication relies on. The magnetic fields around both the transmitter and receiver antennas are theoretically derived and validated by the FDTD simulations. Then rigorous channel model for the metamaterial enhanced MI communication is developed for underground environments. Moreover, the effects of metamaterial loss and dispersion are discussed. Our results prove that by using the metamaterial sphere, the magnetic field around the underground transmitter and receiver can be increased by more than three orders of magnitude. As a result, the magnetic coupling between two coils can be greatly increased. Finally, the communication range of both
the point-to-point MI communication and MI waveguide can be dramatically extended while the size of the antenna is kept in millimeter scale.

APPENDIX

When we analyze the bandwidth of the metamaterial enhanced MI communications in Section IV, we consider the mutual induction \( M \) and self induction \( L \) are not functions of frequency since the metamaterial sphere is small enough i.e. \( kr_1 << 1 \), and \( kr_2 << 1 \). Here we provide the proof.

Proof: The spherical functions mentioned in Section III can be simplified when the variables in those functions are much smaller than 1. If \( x << 1 \), \( j_1(x) \simeq \frac{x}{3} \), \( j_1'(x) \simeq \frac{1}{x} \), \( y_1(x) \simeq -\frac{1}{x^2} \), \( y_1'(x) \simeq \frac{2}{x^3} \), \( h_1^{(1)}(x) \simeq -\frac{j}{x^2} \), \( h_1^{(1)}'(x) \simeq \frac{2j}{x^3} \), \( h_1^{(2)}(x) \simeq \frac{j}{x^2} \), and \( h_1^{(2)}'(x) \simeq -\frac{2j}{x^3} \). Then (5) can be further developed as

\[
Q_t V_t = P_t, \tag{25}
\]

where

\[
Q_t = \begin{pmatrix}
\frac{r_1}{3} & -\frac{r_1}{3} & \frac{1}{r_1^2} & 0 \\
\frac{2r_1}{3\mu_1} & -\frac{2r_1}{3\mu_2} & -\frac{1}{r_1^2\mu_2} & 0 \\
0 & \frac{r_2}{3} & -\frac{1}{r_2^2} & -\frac{j}{r_2^3} \\
0 & \frac{2r_2}{3\mu_2} & \frac{1}{r_2^2\mu_2} & -\frac{j}{r_2^3}\mu_3
\end{pmatrix}, \tag{26}
\]

\[
V_t' = \begin{bmatrix}
A_t k_1 & B_t k_2 & C_t & D_t \\
k_2 & k_2 & D_t & 0
\end{bmatrix}, \tag{27}
\]

and

\[
P_t' = \begin{bmatrix}
-\frac{\omega \mu_1 a^2 I_0 j}{4r_1^2}, & \frac{\omega a^2 I_0 j}{4r_1^2}, & 0, & 0
\end{bmatrix}.
\tag{28}
\]

Since \( Q_t \) does not affected by frequency, \( Q_t^{-1} \) is can be regarded as a constant. It’s obvious that, \( A_t \) is independent of frequency. As a result, the self inductance is independent of frequency. Meanwhile, \( D_t \) is a function of \( \omega^3 \).

Similarly, (11) can be simplified by

\[
Q_r V_r = P_r, \tag{29}
\]
where

\[
Q_r = \begin{pmatrix}
-\frac{r_1}{3} & \frac{r_1}{3} & -\frac{1}{r_1} & 0 \\
-\frac{2r_1}{3\mu_1} & \frac{2r_1}{3\mu_2} & \frac{2}{r_1\mu_2} & 0 \\
0 & -\frac{2}{3} & \frac{2}{r_3} & -\frac{2j}{r_2^3} \\
0 & -\frac{2}{3\mu_2} & -\frac{r_3^2\mu_2}{r_2^2\mu_3} & \frac{j}{r_2^3\mu_3}
\end{pmatrix}, \tag{30}
\]

\[
V_r' = \begin{bmatrix}
A_r k_1, & B_r k_2, & C_r, & D_r
\end{bmatrix}, \tag{31}
\]

and

\[
P_r' = \begin{bmatrix}
0, & 0, & \frac{2jD_t}{k_3^2(d-r_2)^3}e^{-\frac{d-r_2}{3}}, & \frac{2jD_t}{k_3^2\mu_3(d-r_2)^3}e^{-\frac{d-r_2}{3}}
\end{bmatrix}. \tag{32}
\]

$Q_r$ is also independent of frequency. It should be noted that even $\delta$ is a function of $\omega$, the value of $\delta$ is almost a constant when the frequency falls into the bandwidth since the bandwidth of MI communication is very small. Since $D_t$ is a linear function of $\omega^3$, $P_r$ should be a linear function of $\omega$. Thus, $A_r$ does not contain $\omega$. As a result, the mutual inductance is independent of frequency.

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