Magneto-Inductive Underground Communications in a District Heating System

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Abstract—Feasibility of underground data communications is investigated by employing magnetic induction as the key technology at physical layer. Realizing an underground wireless sensor network for a district heating plant motivates this research problem. The main contribution of the paper is to find the optimal design for transmitter and receiver coil antennas, assuming predefined wire gauge and length at both sides. Analytical results show that the transmitter coil should have the smallest possible radius while the converse is true for the receiver coil.

I. INTRODUCTION

Wireless communications in underground and confined areas have always been a challenging problem due to the limited coverage range. We faced this problem on our way to realize a distributed control system for district heating networks. The goal is to create reliable communication links among pumps of an urban-wide hydraulic network. Pumps are located approximately 2 m underground in buildings’ basements and are physically connected together via welded steel pipes. A previous study has shown that a 40 m coverage range is fairly acceptable. It provides at least one reachable neighbor for each pump, while 60% of the pumps have five accessible neighbors in a typical piping layout.

Several technologies have been examined by researchers from various fields for underground and even in-pipe communications. A thorough survey is presented in [2] where the following technologies are investigated in details.

- Radio Frequency (RF) Electromagnetic (EM) waves through-the-soil
- RF EM waves for combined underground and above-ground scenarios
- Pipelines as EM waveguides
- Acoustic waves through the pipelines
- Acoustic waves through the district heating water
- Power Line Communications
- Employing electrical conductivity of the steel pipes
- Incorporating Cell Phone infrastructure
- Magnetic Induction (MI)

From the above choices, all but the last one, turned out to have serious shortcomings [2]. Therefore, the focus of this paper is on MI as the potential winning candidate.

MI, to be described in Section II, has already found several applications in practice. It is used in very low power wireless microphones and headsets that are used by security forces [3] in frequency ranges 11-15 MHz and distances up to 3 m providing a high level of security by involving quasi-static fields, to be described later. Lower frequencies of 300 Hz to 4 kHz are employed for distances up to a few kilometers in soil or water medium. They have been used in underground and underwater data communications for mining and military purposes, underwater navigation, remote triggering of underground and underwater mines and ammunition, seismic fence, and submarine communication systems [4], [5]. We will see in Section II, that in such cases the communication media are hardly penetrable by higher frequencies.

The main contribution of this paper is presented in Section III, where we derive the optimal design criteria for transmitter and receiver antennas of a magneto-inductive communication system by using computer simulations which are based on analytical formulas. Section IV concludes the paper by giving some calculations regarding applicability of the MI technology in our original district heating problem.

II. MAGNETIC INDUCTION

A. MI vs. RF EM waves

MI has been around since 1920’s being used as a means of communication between submarines [5]. There is no fundamental difference between MI and RF EM waves. MI also employs alternating Electric (E) and Magnetic (B) fields as the data carrier, but it uses low frequencies and low power such that the communication range falls off within the distance predominated by the static or quasi-static fields, as described in the sequel.

E and B fields which are created by an alternating electrical current in an antenna can be estimated by the summation of several terms that have different dependencies to the distance \( r \) from the antenna, e.g. \( 1/r^3 \), \( 1/r^2 \), and \( 1/r \). In the vicinity of the antenna, fields are very complex (to be shown in the simulations in Section III). Gradually, they are predominated by \( 1/r^3 \) terms. Up to this zone, is called the static field domain. Other terms predominate at larger distances and are called inductance field (quasi-static field) and radiation field, respectively [6]. The latter accounts for EM radiation.

To define which field component is the leading one, \( r \) should be compared with \( k = \omega/c \) where \( \omega \) is the angular frequency of E and B fields, \( v \) is the speed of EM waves in the medium, and \( k \) is called the wave number. \( k.r \gg 1 \) and \( k.r \ll 1 \) imply...
that the radiation and static fields are dominant, respectively. Otherwise, the field is quasi-static.

Static and quasi-static fields decay much faster than the radiation field with increasing distance from the current source. This property of non-radiating fields have made them intriguing for military applications where security is a concern, but it is of little value in our district heating problem. Actually, it is only the choice of the operating frequency which leaves us in the non-radiating field domain.

Why should we choose such low operating frequencies? The answer lies in the dielectric properties of lossy materials like wet soil. When materials are influenced by alternating E and B fields, their electric susceptibility ($\chi_e$) can be stated as a complex function of frequency:

$$\chi_e(\omega) = \chi_e'(\omega) - j\chi_e''(\omega)$$

The general behavior of the real part $\chi_e'(\omega)$ and the imaginary part $\chi_e''(\omega)$ for most of the materials is sketched in Fig. 1 [7]. $\chi_e'(\omega)$ increases most of the times with respect to $\omega$, except for short intervals where it drastically decreases. These intervals coincide with local maxima in $\chi_e''(\omega)$. $\chi_e'(\omega)$ defines the velocity of E and B fields in the matter, while $\chi_e''(\omega)$ accounts for attenuation due to molecular resonance phenomena.

![Fig. 1. Electrical susceptibility of matters versus frequency](image)

It provides an extra alternating field conductivity ($\sigma_a$) besides the common static field conductivity ($\sigma_s$) at resonance frequencies. The equivalent conductivity ($\sigma_e$) can be stated as:

$$\sigma_e = \sigma_s + \sigma_a = \sigma_s + \omega(\varepsilon_0\chi_e''(\omega))$$

Equation (2) and Fig. 1 imply that, E and B fields should have low frequencies to alleviate the total conductivity, and hence the attenuation. The lower frequency, the less attenuation.

A starting point to choose the appropriate operating frequency is to compare our problem with other MI applications. Knowing that the index of refraction ($n$) of EM waves in soil is $\sim 1.5$ [8], the speed of EM waves in soil is given as $v = c/n \approx 2 \times 10^8$ m/sec. Therefore, for distances of 10 m to 70 m and working frequencies of 2 kHz to 9 kHz, we have $k.r \in [0.006, 0.01]$ which falls well in the non-radiating field domains and does not interfere with any licensed frequencies.

The above discussion explains why MI is a potential solution to our problem. Furthermore, in non-radiating fields, there holds no relation such as $|E|/|B| = v$. Thus, B can be substantially larger than E. Therefore, in a MI system, multipath effects is minimized due to fairly similar values of magnetic permeability ($\mu$) for most of the materials.

### B. System Components

Fig. 2 shows the basic elements of an MI communication system. Coil antennas are used at the transmitter and receiver sides. Coils have low radiation resistance and transmit very little real power. Therefore, they have low performance for radiation field. Instead, a coil antenna has a high reactive power going back and forth to the antenna and the surrounding space in each cycle. This property is desirable for MI, in contrast with typical RF systems where reactive power is kept as low as possible.

![Fig. 2. MI transmitter and receiver circuits](image)

A coil antenna operating at low frequencies can be well modeled as an ideal inductance in series with a resistance which is equal to the resistance of the coil wire. We assume that the operating frequency is low enough, hence ignore the self capacitance across the coil in its lumped model.

At the transmitter, our objective is to produce the strongest possible B field at a considerable distance from the antenna. This will be achieved when the electrical current in the coil ($I_t$) is maximized. Therefore, a capacitor $C_t$ should be connected to the coil in series and bring it to resonance such that $I_t = v_t/R_t$. In practice, a voltage source with low output impedance should be used. For example, an audio amplifier in conjunction with a standard signal generator works well [2].

The receiver coil, on the other hand, should be sensitive to the slightest change of the passing magnetic flux and produce the highest voltage. The induced back-emf $v_r$ can be calculated in a multi-turn coil by:

$$v_r = \omega \sum_{i=1}^{N_r} \left[ \int_{A_i} \vec{B} \cdot \hat{n} \, ds \right]$$

where $\omega$ is the angular frequency of $I_t$, $A_i$ the area of individual turns of the Rx coil, $N_r$ the number of turns of the Rx coil, and $\vec{B}$ is the RMS magnetic field passing through it. If $\vec{B}$ has a constant amplitude at all points on the Rx coil surface and is parallel to $\hat{n}$, which has the unit length and is perpendicular to the coil surface, (3) will be simplified as follows:

$$v_r = \omega |\vec{B}| \sum_{i=1}^{N_r} A_i$$

which shows that $\sum_{i=1}^{N_r} A_i$ should be maximized in order to get the highest possible back-emf. Furthermore, if a capacitor $C_r$ is connected to both ends of the coil such that it brings the coil into resonance (see Fig. 2), the followings hold:

$$C_r = \frac{1}{\omega^2 L_r}$$

$$v_z \approx \frac{1}{jC_r \omega} I_r = -j \omega L_r \left( \frac{v_r}{R_r} \right) = -j Q v_r$$
provided that the input impedance of the receiver amplifier/filter is very large and the quality factor \( Q \) of the coil is defined as \( Q = \frac{\omega L_r}{R_r} \). Therefore, if the Rx coil has a high \( Q \), the passive band-pass receiver circuit acts as a preamplifier.

Combining (4) and (5), the design objective in the Rx antenna is to maximize the following expression:

\[
\frac{L_r}{R_r} \sum_{i=1}^{N_r} A_i
\]  

(6)

As another approach, coupling between the transmitter and receiver antennas can be formulated as mutual inductance between the two coils [9]. In district heating and water distribution systems, it is possible to amplify the mutual inductance by employing pipes as ferromagnetic cores of the transceiver antennas. This will increase the self and mutual inductance in expense of adding core losses. We have shown in [1] that the mutual inductance can be enhanced by several orders of magnitude when coils are wound around steel pipes. However, iron losses counteract mutual inductance enhancement and we have used air core coils throughout this paper.

Employing intermediate coils as magneto-inductive waveguides is another technique to increase the range in specific directions and are investigated in [10], and [11]. However, it is not practical in our problem to insert intermediate coils because of the lack of pipes accessibility at regular distance intervals.

III. COIL ANTENNA DESIGN

Given a specific wire gauge \( R \) and length \( l \), an interesting problem is to find how to wind the coils such that the communication range is extended as far as possible. Assumptions on wire gauge and length are reasonable since they define the cost and weight of the coils. They hold throughout this section.

We start by introducing the analytical expression for the magnetic vector potential first, and taking the curl of it to find \( \mathbf{B} \) [12].

\[
\mathbf{B}(r, z) = \frac{\mu I k}{4\pi\sqrt{ar}} \left[ -(z - h) \left( K(k) - \frac{2 - k^2}{2(1 - k^2)} E(k) \right) \hat{\mathbf{r}} + r \left( K(k) + \frac{k^2(r + a) - 2r}{2r(1 - k^2)} E(k) \right) \hat{\mathbf{z}} \right]
\]  

(7)

\( \mathbf{B} \) is expressed in terms of complete elliptic integral functions of first and second kind, i.e. \( K(k) \) and \( E(k) \), where their argument \( k \) is defined as:

\[
k = \sqrt{\frac{4ar}{(r + a)^2 + (z - h)^2}}
\]

All other parameters of (7) are expressed in Fig. 3. If \( \mathbf{B} \) is only sought on \( z \)-axis, (7) will be significantly simplified. In that case, it would be more convenient to directly use the Biot-Savart law to calculate on-axis \( \mathbf{B} \) field.

A. Relative positioning of Tx and Rx antennas

The graphical representation of \( |\mathbf{B}| \) for a multi-turn coil is given in Fig. 4 as a contour plot at two zoom levels. The coil is not shown in this plot, but it is laid in \( z = 0 \) plane and centered at the origin, with \( a = 0.1 \) m.

The field is calculated at each point in space by summing up (7) for individual wire turns, considering the exact radius and location for each turn of wire.

It is shown that \( |\mathbf{B}| \) is larger around the coil and on the coil’s axis. An interesting behavior is observed on \( z = 0 \) plane where \( |\mathbf{B}| \) has a local minimum at a radius of approximately 2.4 \( a \). This is due to the \( \mathbf{B} \) field generated by the currents flowing in farther parts of the coil which counteract the \( \mathbf{B} \) field generated by the currents flowing in the closer parts. At larger distances from the coil, \(|\mathbf{B}| \) starts to degrade monotonically.

Since the contours in Fig. 4 (right) do not follow an exact spherical pattern, there are variations in \(|\mathbf{B}| \) at fixed distances from the Tx antenna. It is clear that \(|\mathbf{B}| \) is larger around the \( z \)-axis. This behavior is quantified in Fig. 5, in which \(|\mathbf{B}| \) is drawn at three different fixed distances of 1 m, 10 m, and 100 m from the Tx antenna with \( a = 0.1 \) m. Each curve starts from vicinity of the \( z \)-axis, associated with \( \theta = 0.1^\circ \) to \( z = 0 \) plane, associated with \( \theta = 90^\circ \), where \( \theta \) is illustrated in Fig. 5. The results are shown in log-log scale in sake of highlighting the differences. Moreover, the curves are shifted vertically such that they all coincide at \( \theta = 90^\circ \). Fig. 5 shows that \(|\mathbf{B}| \) can vary up to approximately 20 dB at the Rx antenna based on positioning of the Tx antenna if \( \theta \in [1^\circ, 90^\circ] \). It also shows that \(|\mathbf{B}| \) is more sensitive for small values of \( \theta \).

Note that Fig. 5 does not say anything about the angles between Tx and Rx coils’ axes, which is important in calculating the magnetic flux passing through the Rx coil.
B. Cross section shapes

Given the mean radius of the coil $a$, and hence the number of its turns $N_r (N = l/(2\pi a))$, we have compared rectangular cross section shapes with different side proportions to see its effect on the Tx and Rx antennas. These are ranging from a single layered coil in which $b = 2R$ and $c = Rl/(\pi a)$ to a coil with the maximum number of winding layers which is made when $b \leq 2a$ and $c \geq R^2l/(\pi a^2)$. The latter holds, provided that each wire occupies a square surface of area $4R^2$ at the cross section of the coil. The most compact wiring, results in hexagonal surface assignment to each wire turn and $c \geq \sqrt{3}/2\pi a^2$. All of the above inequalities convert to equalities if the wire is very long such that no hole remains in the center of the multi-turn coil.

At the Tx coil, the difference between $|B|$ of a single layer and a square coil at $z = 0$ plane is in order of 0.01% of the $|B|$ created by either of them. In all other points in the space, this trifling difference is even more negligible. Therefore, the cross section shape of the Tx antenna is not tangibly influential on $|B|$.

At the Rx coil, on the other hand, the coil’s cross section shape is quite effective mostly due to its effect on the self inductance of the coil ($L_r$). This is explicitly presented in (6). Since the length and the mean radius of the coil are pre-defined, $\sum_{i=1}^{N_r} A_i$ is constant and $L_r$ must be maximized.

Maximization of self inductance of a coil was first posed by Maxwell in [13]. He has shown that the geometric mean distance between the conductors of an inductor should be minimized in order to get the maximum inductance. This is achieved when the cross section is circular. For rectangular shapes which are easier to manufacture, the best shape is a square, i.e. $b = c$ (See Fig. 6). This result is verified by simulations in Fig. 7.

The coil in this simulation is made up with 150 m of wire with gauge $R = 0.3$ mm. The cross section is rectangular and the mean radius is $a = 0.20$ m. Consequently, the number of turns $N_r$ turns out to be 119 which yields 11 winding layers for a square cross section. Fig. 7 confirms this fact by showing a peak at $N_r = 11$. It also shows that a single layer Rx coil has only 60% performance of the square shaped Rx coil.

In the above simulation, $L_r$ is calculated by injecting a current $I$ into the Rx coil and computing the induced magnetic flux $\Phi_i$ in each turn due to the current in all of the turns. Then, $L_i = \Phi_i/gives L_i and \ L_r = \sum_{i=1}^{N_r} L_i$. Note that, in calculating the total self inductance of the Rx coil, we only added up $L_i$ of individual turns which convey the self inductance of that turn plus its mutual inductance to all other turns.

C. Mean radius vs. Number of turns

Since the total wire length ($l = 2\pi aN$) is pre-defined, $aN$ is constant and we should make a trade-off between $a$ and $N$. At the Tx antenna, Fig. 8 exhibits the variations in $|B|$ at $z = 0$ as $r$ increases from 1 m up to 50 m, for different values of $a$ from 0.03 m to 0.25 m and square cross section. The lowest value is associated with the physical limitation of $a > b/2$ (see Fig. 6 for clarification).

Fig. 8 suggests that $a$ should be as small as possible to have the maximum $|B|$ at large distances. The same results were achieved when $z \neq 0$ and also on any straight line with slope $\alpha$ such that $z = \alpha r$. No matter how large a finite $\alpha$ is, the above hypothesis holds true and the smallest mean radius gives the maximum $|B|$ at distances of interest. The only exception is $\alpha = \infty$, i.e. $|B|$ on the $z$ axis in which the converse is true, but it does not have any practical importance.

For the Rx coil, (6) is depicted in Fig. 9 as a function of $a$. For all values of $a$, a square cross section is considered.
Fig. 9 admits that for the Rx coil, larger $a$ produces higher voltages, no matter if the number of turns or $L_r$ is decreased significantly. In practice, $a$ is usually specified by the space limitations. Moreover, $Q$ of the Rx coil should be kept greater than 1, not to attenuate the signal. The presence of $C_r$ capacitor is always necessary to adjust the phase of the received signal, such that it can be fed into amplifiers with a resistive input impedance.

**D. Summary of Coil Design Criteria**

- The Tx coil should have the smallest possible $a$ while the Rx coil should have the largest possible $a$.
- The cross section of the Rx coil should be circular or square (when a circular shape is difficult to manufacture), while the cross section shape of the Tx coil might be freely chosen based on practical winding issues. Combining the above two statements suggests that the Tx coil should be single layered with a very large length $(c)$ and $a = b/2$ which is impractical. Therefore, in practice, the maximum permissible length of the Tx antenna $(c_{\text{Max}})$ should specify the coil’s dimensions. As for the Rx coil, the maximum permissible radius $(a_{\text{Max}})$ specifies its size.
- The Rx coil center ought to be located on the Tx coil axis. Otherwise the attenuation in $|\vec{B}|$ could be approximated by the plot in Fig. 5.
- When the coils and their locations are known, the receiver should face the right direction such that $\vec{B}$ is perpendicular to the Rx coil surface. In worst case, when $\vec{B}$ and the coil’s surface are parallel, no signal will be received.

**IV. DISCUSSION**

Having found the best conditions for Tx and Rx antennas, it is time to investigate feasibility of communication in our district heating problem. We have taken the following assumptions into account:

For the Tx antenna, 150 m of wire with $R = 0.3$ mm and electrical resistance $0.06 \, \Omega/m$ is used to make the coil with $a = 1.30 \, \text{cm}$ and $c = 5.00 \, \text{cm}$. This arrangement results in 1836 turns of wire in 22 layers with $b = 1.32 \, \text{cm}$. For the Rx antenna, the same wire gauge and length is used, but with $a = 18.0 \, \text{cm}$ and a square cross section which result in 133 turns of wire with $b = c = 0.69 \, \text{cm}$. The Tx antenna is located at $z = 0$ plane, centered at the origin, while the Rx antenna is located on $r = z$ cone facing the perfect direction. Note that the location $r = z$ is far from the best position. Being closer to the $z$ axis is preferable in practice.

Electrical current through the Tx coil is assumed to be 100 mA RMS. Fig. 10 exhibits $\log_{10} |\vec{B}|$ as the inter-distance of antennas increases. When the Tx and Rx antennas are 40 m away from each other: $|\vec{B}| \approx 29.9 \, \text{nT}$.

At the Rx antenna, (6) turns out to be $2.50 \times 10^{-2} \, \text{Hm}^2/\Omega$. It should be multiplied by $\omega^2 |\vec{B}|$ to finally give $v_z$, which is equal to 265.5 mV at the operating frequency 3000 Hz. Assuming 1 mV as the threshold of a detectable signal, the calculated $v_z$ is easy to detect with a standard filter/amplifier combination at the receiver. This ample margin indicates that MI is a working method for making reliable communication links for underground wireless sensor networks.

**REFERENCES**