Electromagnetic Subsurface Remote Sensing†

S. Y. Chen¹, W. C. Chew¹, V. R. N. Santos², K. Sainath², and F. L. Teixeira²

¹University of Illinois at Urbana-Champaign
²The Ohio State University

March 15, 2016

Abstract—Fundamental principles and several applications of electromagnetic (EM) subsurface remote sensing methods are outlined and illustrated. Physical insight is emphasized rather than mathematical analysis. The various methods are classified according to the practical embodiment and their range of applications. These include borehole EM methods, ground penetrating radar, magnetotelluric methods, airborne methods, inductive EM methods, time-domain EM methods, and marine EM methods.

Keywords: Borehole, ground penetrating radar (GPR), magnetotelluric methods, inductive methods, time-domain electromagnetic (TDEM), controlled source electromagnetic methods (CSEM), well-logging.

1 Introduction

Subsurface electromagnetic (EM) remote sensing methods are applied to obtain underground information that is not available from direct surface observations. Since electrical parameters such as dielectric permittivity and conductivity of subsurface materials may vary dramatically, the response of electromagnetic signals can be used to map the underground structure [1]. Another major application of subsurface EM methods is to detect and locate underground anomalies such as mineral deposits or to detect, locate, and classify buried objects such as underground pipes, landmines, unexploded ordnances, etc.

Subsurface EM methods include a variety of techniques depending on the application, surveying method, hardware system available, and interpretation procedure; thus, a “best” method simply does not exist. Even though each system has its own characteristics, they still share some common features. In general, each system has a transmitter or emitter, which can be either natural (passive sensing) or artificial (active sensing), to send out the electromagnetic energy that serves as an input signal. In both cases, a receiver is needed to collect the response signal. The underground earth formation is characterized by the material parameters and geometry, which may be unknown or only partially known. The task of subsurface EM methods is to better characterize the underground properties from the response signal. For example, in a subsurface EM system of inductive type, the EM transmitter radiates a primary field into the subsurface. According to the conductivity of the subsurface earth formation in the vicinity of the system, this primary field induces currents in the subsurface, which in turn radiate a secondary field. The secondary field can be detected by the receiver and, after proper data processing and interpretation, information about the underground properties can be obtained.

† To appear in Wiley Encyclopedia of Electrical and Electronics Engineering.
One of the most challenging parts of subsurface EM methods is interpretation of the data. Since the incident field interacts with the subsurface in a very complex manner, it is never easy to subtract the information from the receiver signal. Many definitions, such as apparent conductivity, are introduced to facilitate this procedure. Data interpretation is also a critical factor in evaluating the effectiveness of the system. How good the system is always depends on how well the data can be utilized to provide the sought after information (detection, localization, classification). In the early development of subsurface EM systems, data interpretation largely depended on the personal experience of the operator, due to limitations on sensor hardware and signal processing, and to the general complexity of the problem. Only with the aid of powerful computers, improvements in signal processing and computational EM techniques, as well as on inversion (or imaging) algorithms it is possible to analyze such complicated problem in reasonable time. Computer-based interpretation and inversion methods are attracting more and more attention. Nevertheless, data interpretation remains “an artful balance of physical understanding, awareness of the geological constraints, and pure experience” [2].

In the following Sections of this article, we will illustrate several applications and outline basic principles of subsurface EM methods. Physical insight is emphasized rather than mathematical analysis. The various methods are classified here mainly according to the practical embodiment and their range of applications. As such, a same physical mechanism (for example, electromagnetic induction) may underlie different methods. The last Section provides an overview of numerical and analytical modeling techniques applied to subsurface methods. Further details of each method can be found in the references.

2 Borehole EM Methods

Borehole EM methods are an important part of well-logging methods. Since water is conductive and hydrocarbons (oil and gas) are insulators, resistivity measurements are good indicators of the presence of hydrocarbon-bearing zones. On the other hand, water has an unusually high dielectric constant, and permittivity measurement is therefore a good detector of moisture content.

![Figure 1: Basic configuration of a borehole induction tool, with one transmitter coil antenna (Tx) and two receiver coil antennas (Rx1 and Rx2) wrapped around a metallic mandrel.](image)
Early borehole EM methods consisted of mainly galvanic electrical measurements using very simple low-frequency or direct-current (DC) electrodes like the short and the long normal to execute well-logging for oil exploration. Basically, these logging tools are comprised of a series of electrodes. A transmitter electrode injects a current into the conductive formation. At another underground locations (along the same borehole or at another boreholes), receiver electrodes measure the relative voltages between them. These voltage differences provide an estimate of the effective resistivity of the surrounding formation. High resistivities (or, equivalently, lower conductivities) typically imply the possible presence of hydrocarbons. More sophisticated electrode tools operating on the same principle also exist and are routinely used. Some of these tools are mounted on a mandrel, which performs measurements centered in a borehole. These tools are called mandrel tools. Alternatively, the sensors can be mounted on a pad, and the corresponding tool is called a pad tool. We will examine galvanic or electrode methods in a bit more detail after we first consider induction tools next.

A very successful borehole EM method is induction logging. Compared to resistivity logging based on galvanic methods that utilize on electrodes, induction logging can be used at higher frequencies and in the presence of non-conductive borehole. The latter can occur, for example, due to the presence of fresh-water mud or oil-based mud inside the borehole used during the drilling process, or in air-filled boreholes. Since its introduction many decades ago [3], this technique has been used widely with confidence in the oil and gas industry. Extensive research work has been done in this area. The systems in use now are so sophisticated that many modern signal processing techniques are involved. Nevertheless, the principles still remain the same and can be understood by studying a simple case. Induction logging makes use of several coils wound on an isolating mandrel, also called a sonde. Some of the coils, referred to as transmitters, are powered with alternating current (AC). Figure 1 illustrates a basic induction tool with one single transmitter coil antenna and two receiver coil antennas. The typical distance between the transmitter and receiver antennas is on the order of a few meters or less. The transmitters radiate the field into the conductive formation and induce an eddy (secondary) current in the zone next to the transmitter. This eddy current is roughly proportional to the formation conductivity and radiates a secondary field, which can be detected by the receiver coils. The receiver signal (voltage) is normalized with respect to the transmitter current and represented as an apparent resistivity (or conductivity), which serves as an indication of the actual underground resistivity. Traditional induction tools employ horizontal coils as depicted in Fig. 1, but tilted coils can also be utilized [4], [5]. The latter generate fields with azimuth variation and hence can provide directional information useful for geosteering applications and horizontal drilling [6], [7]. Horizontal drilling occurs when a vertical borehole encounters a producing formation (payzone) with hydrocarbons and is then steered to remain with the payzone and achieve higher production rate. Since payzones are typically layers parallel to the ground (horizontal), the drilling proceeds horizontally from that point on. Horizontal drilling obviates the need for drilling many vertical boreholes over an extensive geographical area. Logging tools with tilted coil antennas can also discern both vertical and horizontal resistivities in anisotropic formations [8], [9]. The field produced by electrically small coil antennas is approximately equivalent to that of an equivalent magnetic dipole normal to the plane of the coil [10]. Coil antennas used in induction tools are considered electrically small because their diameters are much smaller than the wavelength of operation. The frequency of operation of induction tools is typically less than 2 MHz,
Electromagnetic subsurface remote sensing

which corresponds to a wavelength of approximately 150 m. On the other hand, typical coil diameters of induction tools are on the order of 20 cm. For induction tools with horizontal coils, the equivalent magnetic dipoles are aligned with the mandrel axis, whereas for induction tools with tilted coils, the equivalent magnetic dipoles have components both along the mandrel axis and perpendicular to it. With a proper arrangement, the tool can therefore possess equivalent magnetic dipoles with components along all three axes. Such tools are often referred to as triaxial tools.

To obtain information from the apparent conductivity measured by induction tools, we need to understand how apparent conductivity and true conductivity are related. According to Doll’s theory \[3\], the relation in cylindrical coordinates is given by

\[
\sigma_a = \int_{-\infty}^{\infty} dz' \int_{0}^{\infty} d\rho' g_D(\rho', z') \sigma(\rho', z')
\]

where \(\sigma(\rho, z)\) is the true formation conductivity and \(\sigma_a\) is the apparent conductivity. The kernel \(g_D(\rho, z)\) is the so-called Doll’s geometrical factor, which weights the contribution of the conductivity from various regions in the vicinity of sonde. We notice that \(g_D(\rho, z)\) is not a function of the true conductivity and hence is only determined by the tool configuration. The interpretation of the data would be simple if Doll’s theory were exact. Unfortunately, this is rarely the case. Further studies show that Eq. 1 is true only in some extreme cases. The significance of Doll’s theory, however, is that it relates the apparent conductivity and formation conductivity, even though the theory is not exact. In the early development of induction logging techniques, tool design and data interpretation were based on Doll’s theory, and in most cases it gives reasonable answers.

To establish a firm understanding of induction logging theory, we need to perform a rigorous analysis by using Maxwell’s equations as follows [11], [12]:

\[
\nabla \times \mathbf{H} = -i \omega \varepsilon \mathbf{E} + \mathbf{J}_s + \sigma \mathbf{E}
\]

\[
\nabla \times \mathbf{E} = i \omega \mu \mathbf{H}
\]

\[
\nabla \cdot \mathbf{H} = 0
\]

\[
\nabla \cdot \mathbf{D} = \rho
\]

where \(\nabla \cdot \mathbf{J}_s = i \omega \rho\).

In the preceding equations, the time-harmonic dependence \(\exp(-i\omega t)\) is assumed, and \(\mathbf{J}_s\) corresponds to the impressed current source. Parameters \(\mu\) and \(\varepsilon\) are the magnetic permeability and dielectric permittivity, respectively. To simplify the analysis, we assume that both the impressed source and geometry of the problem are axisymmetric; consequently, all the field components are independent of the azimuthal angle. Furthermore, it can be shown that there is no stored charge under the preceding assumption. A typical working frequency of induction logging is about 20 kHz; in this range of frequencies, the displacement current \(-i \omega \varepsilon \mathbf{E}\) is very small compared to the conduction current \(\sigma \mathbf{E}\) and hence is neglected in the following discussion. After these simplifications, we have

\[
\nabla \times \mathbf{H} - \sigma \mathbf{E} = \mathbf{J}_s
\]

\[
\nabla \times \mathbf{E} = i \omega \mu \mathbf{H}
\]

\[
\nabla \cdot \mathbf{H} = 0
\]

\[
\nabla \cdot \mathbf{E} = 0
\]

where we assume \(\nabla \cdot \mathbf{J}_s = i \omega \rho = 0\). For convenience, the magnetic vector potential \(\mathbf{A}\) is introduced next. Since \(\nabla \cdot \mathbf{H} = 0\) and \(\nabla \cdot (\nabla \times \mathbf{E}) = 0\), it is possible to define \(\mathbf{A}\) through \(\mathbf{H} = \nabla \times \mathbf{A}\). To specify the potential \(\mathbf{A}\)
uniquely, we choose \( \mathbf{E} = i \omega \mu \mathbf{A} \), which is only true when there is no charge accumulation. Substituting these expressions into Eq. 6, we have
\[
\nabla \times \nabla \times \mathbf{A} - i \omega \mu \sigma \mathbf{A} = \mathbf{J}_s.
\]
By using an appropriate vector identity, we have
\[
\nabla^2 \mathbf{A} + k^2 \mathbf{A} = -\mathbf{J}_s
\]
where
\[
k^2 = i \omega \mu \sigma.
\]
To show how the apparent conductivity and formation conductivity are related, we first write down the solution of Eq. 11 in a homogeneous medium as follows [13], [14]:
\[
A(\rho, z, \phi) = \frac{1}{4\pi} \int_{V'} \frac{J_s(\rho', z', \phi') \exp(ik\mathbf{r}_1)}{\mathbf{r}_1} dV'
\]
where
\[
\mathbf{r}_1 = \left[ (z - z')^2 + \rho^2 + \rho'^2 - 2 \rho \rho' \cos(\phi - \phi') \right]^{1/2}.
\]
The volume integration is evaluated over regions containing the impressed current sources and the coordinate system used in Eq. 13, as shown in Fig. 2. Usually, a small current loop is used as an excitation, which implies that only \( A_\phi \) exists. Hence, Eq. 13 can be furthermore simplified as
\[
A_\phi(\rho, z) = \frac{1}{4\pi} \int_{V'} J_\phi(\rho', z') \cos(\phi - \phi') \frac{\exp(ik\mathbf{r}_1)}{\mathbf{r}_1} dV'.
\]

Figure 2. Induction logging tool transmitter and receiver coil pair used to explain the geometric factor theory (redrawn from [14]).
When the radius of the current loop becomes infinitely small, it can be viewed as a magnetic dipole and thus the preceding integration can be approximated as

$$A_\phi (\rho, z) = \frac{m}{4\pi} \frac{\rho}{r_1^3} (1 - ikr_1) e^{ikr},$$  \hspace{1cm} (16)

where $m = N_T/\pi a^2$ is the magnetic dipole moment and $N_T$ is the number of turns wound on the mandrel. At the receiver point, the voltage induced on the receiver with $N_R$ turns can be represented as

$$V = 2\pi a N_T E_\phi = \frac{2 N_T N_R (\pi a^2)^2}{4\pi} i\omega \mu (1 - iKL) e^{iKL}/L^3$$ \hspace{1cm} (17)

where

$$E_\phi = i\omega \mu A_\phi (a, L)$$ \hspace{1cm} (18)

and $L$ is the distance between the transmitter and receiver. Since the voltage is a complex quantity, it can be separated into real and imaginary parts and expanded in powers of $kL$ as follows [13]:

$$V_R = -K\sigma \left(1 - \frac{2}{3} \frac{L}{\delta} + \cdots\right)$$ \hspace{1cm} (19)

$$V_X = K\sigma \left(\frac{\delta^2}{L^2} \left(1 - \frac{2}{3} \frac{L^2}{\delta^3} + \cdots\right)\right)$$ \hspace{1cm} (20)

where

$$K = \frac{(\omega \mu)^2 (\pi a^2)^2}{4\pi} \frac{N_T N_R I}{L}$$ \hspace{1cm} (21)

and

$$\delta = \frac{2}{\sqrt{\omega \mu \sigma}}.$$ \hspace{1cm} (22)

The quantity $K$ is known as the tool constant and is totally determined by the configuration of the tool, and $\delta$ is the so-called skin depth [15], which describes the attenuation of a conductor in terms of the field penetration distance, as illustrated in Figure 3. The quantity $V_R$ is called the $R$ signal. The apparent conductivity is defined as [13]

$$\sigma_a = \frac{V_R}{K} \simeq \sigma \left(1 - \frac{2}{3} \frac{L}{\delta} \right).$$ \hspace{1cm} (23)

In the preceding analysis, there are some important points that need to be discussed. In Eq. 19, we see that the apparent conductivity is a nonlinear function of the true conductivity, even in a homogeneous medium. The lower the working frequency or lower the true conductivity, the more linear it will be. The difference between true conductivity and apparent conductivity is defined as the skin effect signal,

$$\sigma_s = \sigma - \sigma_a.$$ \hspace{1cm} (24)

The leading term of the imaginary part, $V_X$, is not a function of true conductivity. In fact, it corresponds to the direct coupling field, which does not contain any formation information. What remains in $V_X$ is the
so-called \( X \) signal. Since the direct term is much larger than the residual part including \( V_R \), it is difficult to separate the \( X \) signal. The importance of the \( X \) signal is seen by comparing Eqs. 19 and 20, from which we find that the \( X \) signal is the first-order approximation of the nonlinear term in \( V_R \), the \( R \) signal. This fact can be used to compensate for the skin effect.

\[ V = \frac{i2 \pi \mu R \omega \mu}{4\pi} \int_{V'}^\infty \int_{V}^\infty dz'' \int_{\rho'}^\infty d\rho' \sigma(\rho'', z'') A_{\phi''}(\rho'', z'') \left( \frac{1}{r_2} \right) dV'. \]

The vector potential can also be separated into real and imaginary parts:

\[ A_\phi = A_{\phi R} + iA_{\phi I}. \]

Substituting Eq. 28 into Eq. 27 and separating out the real part of the receiver voltage, we have

\[ V_R = -\frac{(\omega \mu)^2 (2\pi \mu R)}{4\pi} \int_{-\infty}^\infty dz' \int_{0}^{\infty} d\rho' \sigma(\rho', z') A_{\phi R} \int_{0}^{2\pi} \cos(\phi - \phi') d\phi'. \]
Applying the same procedure, we obtain the apparent conductivity as

\[
\sigma_a = -\frac{V_R}{K} = \int_{0}^{\infty} d\rho' \int_{0}^{\infty} dz' \sigma(\rho', z') g_p(\rho', z'),
\]

where

\[
g_p = \frac{2\pi L \rho'}{(\pi a)^3} A_{\phi, r} \int_{0}^{2\pi} \frac{\cos(\phi - \phi')}{r_2^2} d\phi'.
\]

The function \(g_\phi\) is the exact definition of the geometrical factor. In comparison with Doll’s geometrical factor, \(g_\phi\) depends not only on the tool configuration, but also on the formation conductivity, since the vector potential depends on the formation conductivity. The integral-form solution above does not provide any computational advantage, since the differential equation for \(A_{\phi, r}\) must still be solved. But it is now clear from Eq. 30 that the apparent conductivity is the result of a nonlinear convolution. Equation 30 also represents the starting point of inverse filtering techniques, which make use of both the \(R\) and \(X\) signals to reconstruct the formation conductivity. Finding the vector potential \(A\) is still a challenge though. Although much faster to obtain, analytical or pseudo-analytical solutions are available only for a few simple geometries [10], [16], [17]. In most cases, we have to resort to brute-force numerical techniques such as the finite element method (FEM) [18], [19], [20], finite difference (FD) [9], [21], [22], [23], finite volume (FV) [24], [25], numerical mode matching (NMM) [26], [27], or the volume integral equation method (VIEM) [28],[29], [30], [31]. These methods are discussed in more detail later on in the last Section of this article.

Previously, we mentioned that Doll’s geometrical factor theory is only valid under some extreme conditions. In fact, it can be derived from the exact geometrical factor as a special case [14]. In a homogeneous medium, the vector potential \(A_{\phi, r}\) can be calculated as

\[
A_{\phi, r} \equiv \left(\frac{\pi a^2}{4\pi}\right) N_1 \frac{\rho'}{r_1^3} A \text{Re} \left\{1 - ik r_1 e^{ik r_1}\right\}.
\]

The integration in Eq. 31 can also be performed for \(r_2 \gg a\). The final result is

\[
\sigma_a = -\frac{V_R}{K} = \int_{0}^{\infty} \int_{0}^{\infty} g_D(\rho', z') \text{Re} \left\{1 - ik r_1 e^{ik r_1}\right\} \rho' d\rho' dz',
\]

where

\[
g_D(\rho', z') = \frac{L \rho'^3}{2 r_1^3 r_2^3}.
\]

It is now clear that Doll’s geometric factor and the exact geometric factor are the same when the medium is homogeneous and the wave number approaches zero.

So far we have discussed the basic theory of induction logging. We now use a simple example to show some practical concerns and briefly discuss the solutions. In Fig. 4, we show an apparent resistivity (the inverse of apparent conductivity) response of the commercial logging tool 6FF40TM (trademark of Schlumberger Ltd.) in the so-called Oklahoma benchmark. The piecewise-constant trace is the formation resistivity, and the dashed line is the raw (unprocessed) data of 6FF40. We notice that the apparent resistivity data roughly indicate the variation of the true resistivity, but around 4850 ft the apparent resistivity \(R_a\) is much higher than the true resistivity \(R_t\), which results from the “skin effect” [32]. From 4927 ft to 4955 ft, \(R_a\) is substantially lower than \(R_t\), which is caused by the so-called shoulder effect. The
shoulder effect arises when two adjacent low-resistance layers generate strong signals, even though the tool is not in these two regions. Around 5000 ft, there are a number of thin layers, but the tool's response fails to indicate them. This failure results from the tool's limited resolution, which is represented in terms of the smallest thickness that can be identified by the tool. The thin line in Fig. 4 represents the processed 6FF40 data after skin effect boosting (SB) and a three-point deconvolution (DC). Skin effect boosting is based on Eq.19, which is solved iteratively for the true conductivity from the apparent conductivity. The three-point deconvolution is performed under the assumption that the convolution in Eq. 30 is almost linear \[33\]. These two methods do improve the final results to some degree, but they also cause spurious artifacts observed near 4880 ft, since the two effects are considered separately. The thick line is the response of the HRI™ (high-resolution induction) tool (trademark of Halliburton Co.) \[34\]. For this tool, a more complex coil configuration is used to optimize the geometrical factor. After the raw data are obtained, a nonlinear deconvolution based on the X signal is performed. The improvement in the final results is significant. Another induction tools include the HDIL™ (hostile dual induction log) by Halliburton Co. and the AIT™ (array induction image tool) by Schlumberger Ltd., which uses eight induction-coil arrays operating at different frequencies \[35\]. The deconvolutions are performed in both radial and vertical directions, and a quantitative two-dimensional image of formation resistivity is possible after a large number of measurements \[36\].

Figure 4. Apparent resistivity responses of different tools in the Oklahoma benchmark. The improvement in resolution provided by the HRI tool is significant.

The aforementioned data processing techniques are based on the inverse deconvolution filter, which is computationally effective and easily run in real time on a logging truck computer. An alternative approach is to use inverse scattering theory, which is becoming increasingly practical and promising with the development of high-speed computers \[31\], \[37\].

In addition to induction tools, there are other borehole methods as well, such as electrode tools, which were briefly noted before, and propagation tools \[38\]. As mentioned, induction tools are suitable for non-conductive boreholes, since little or no conductivity in the borehole has a lesser effect on the
measurement. However, if the mud is very conductive it will generate a strong signal at the receiver and hence seriously degrade the tool’s ability to make a deep reading into the earth formation. In such a case, electrode tools are preferable, since the conductive mud places the electrodes into better electrical contact with the formation. Electrode tools employ very low frequencies (about 1 kHz or less) and the transmitted fields are governed by Laplace’s equation instead of Helmholtz equation \[39, 40, 41\]. Commercial electrode tools include DLL™ (dual laterolog) and SFL™ (spherical focusing log) from Schlumberger Ltd., and ADR™ (azimuthal deep resistivity) and AFR™ (azimuthal focus resistivity) from Halliburton Co. DLL and ADR enable deep measurements, while SFL and AFR are intended for shallower measurements \[42, 43, 44\]. The AFR tool also enables logging-while-drilling. There are also a number of tools mounted on pads to perform shallow measurements on the borehole wall. These may be just button electrodes mounted on a metallic pad. Due to their small size, they have high resolution but a shallow depth of investigation. Their high resolution capability can be used to map out fine stratifications on the borehole wall. When four pads are equipped with these button electrodes, the resistivity logs they measure can be correlated to obtain the dip of a geological bed. An example of this is the SHDT™ (stratigraphic high-resolution dip meter tool), also from Schlumberger Ltd. \[45\]. For oil-based mud the SHDT does not work well, and microinduction sensors have been mounted on a pad to dipping bed evaluation. Such a tool is known as the OBDT™ (oil-based mud dip meter tool) and is manufactured by Schlumberger Ltd. \[46, 47\]. When an array of buttons is mounted on a pad, it can be used to generate a resistivity image of the borehole wall for formation evaluation, such as dips, cracks, and stratigraphy. Sometimes these tools are called formation microscanners \[48\]. Examples of such tools are OMRI™ (oil-based mud imaging) and XRMI™ (water based mud imaging), both manufactured by Halliburton Co.

Sometimes information is needed not only relating to the conductivity but also to the dielectric permittivity. In such cases, the EPT™ (electromagnetic wave propagation tool) from Schlumberger Ltd. can be used. The working frequency of EPT can be as high as hundreds of megahertz to 1 GHz. At such high frequencies, the real part of the complex permittivity \(\varepsilon_e\), defined as

\[
\varepsilon_e = \varepsilon + i \frac{\sigma}{\omega},
\]

becomes dominant. EPT measurements provide information about dielectric permittivity and hence can better distinguish fresh water from oil. Water has a much higher dielectric constant (about 80) compared to oil (about 2). Phase delays at two receivers are used to infer the wave phase velocity and hence the permittivity. Interested readers can find materials on these methods in \[49, 50, 51\].

Other techniques in electrical well logging include the use of borehole radar. In such a case, a pulse is sent to a transmitting antenna in the borehole, and the pulse echo from the formation is measured at the receiver. Borehole radar finds application in salt domes where the electromagnetic loss is low. In addition, the nuclear magnetic resonance (NMR) technique can be used to detect the percentage of free water in a rock formation. The NMR signal in a rock formation is proportional to the spin echoes from free protons that abound in water. An example of such tools are the MRIL™ (magnetic resonance imaging logging) from Halliburton Co. and the PNMT™ (pulsed nuclear magnetic resonance tool), from Schlumberger Ltd. \[52\].
3 Ground Penetrating Radar

Another outgrowth of subsurface EM methods is ground penetrating radar (GPR). Because of its numerous advantages, GPR has been widely used in geological surveying, civil engineering, detection of manmade buried targets, and some other areas. As noted in [53], the GPR method was first developed in Germany in 1929. During the 1960's, GPR surveys were used to determine the thickness of ice sheets in the Arctic and Antarctica, and during the 1970s GPR systems became more popular in ice-free environments. Since 1980, there has been great increase in the use of GPR in a variety of different environments. Nowadays, GPR systems are routinely used for geophysical surveys in general, and especially to enable high-resolution near-surface imaging for archeology, geological, geotechnical, environmental (including demining), and urban planning applications. GPR is useful even for organic contaminant detection and nondestructive evaluation of civil structures such as bridges, overpasses, and buildings. The GPR design is largely application oriented. Even though various systems have different applications and considerations, their advantages can be summarized as follows: (i) Because the frequency used in GPR is much higher than that used in the induction method, GPR has a higher resolution; (ii) since the antennas do not need to touch the ground, rapid surveying can be achieved; and (iii) the data retrieved by some GPR systems can often be interpreted in real time [54], [55], [56], [57], [58]. On the other hand, GPR has some disadvantages, such as shallow investigation depth and site-specific applicability. The working frequency of GPR is much higher than that used in the induction method. At such high frequencies, the soil is usually very lossy. Even though there is always a tradeoff between the investigation depth and resolution, the maximum depth of GPR systems is usually 10 m or less and highly dependent on soil type and moisture content.

The basic working principle of GPR is illustrated in Fig. 5(a) [55]. The transmitter \( T \) generates transient or continuous EM waves that propagate into the underground. Whenever a change in the electrical properties of underground regions is encountered, the wave is partially reflected and refracted. A receiver \( R \) detects and records the reflected waves. From the recorded data, information pertaining to the depth, geometry, and material type can be obtained. As a simple example, we use Figs. 5(b) and 5(c) to illustrate how the data are recorded and interpreted. The underground contains one interface, one cavity, and one lens. At a single position, the receiver signals at different times are stacked along the time axis. After completing the measurement at one position, the procedure is iterated at all subsequent positions. The final results are presented in a two-dimensional map, which is called an echo sounder–type display or radargram [59]. When \( T \) and \( R \) move together, we obtain the so-called common-offset radargram. If only \( R \) moves while \( T \) remains stationary, we obtain the common-source radargram. To locate objects or interfaces, we need to know the phase velocity in the underground dielectric medium. The phase velocity \( v \) in a dielectric medium is

\[
v = \frac{c}{\sqrt{\varepsilon_r}}
\]

(36)

where \( c = 3 \times 10^8 \) m/s and \( \varepsilon_r \) is the relative permittivity. Usually, the transmitter and the receiver are close enough and thus the wave’s path of propagation is considered to be approximately vertical. The depth \( d \) of the interface can be therefore approximated as
Electromagnetic subsurface remote sensing

\[ d = \frac{v T_{d}}{2} \]  

(37)

where \( T_{d} \) is the total delay time. Figure 6 depicts an example of a common-offset radargram based on the above principle, for a subsurface region with three objects delineated in grey circles. This radargram was obtained using the commercial GPR system SIR3000™ from GSSI (Geophysical Survey Systems, Inc.), operating at 270 MHz (central frequency) and based on shielded antennas. The reflected traces visible from the radargram correspond to approximate hyperbolas, from which the object locations can be inferred.

Figure 5. Working principle of the GPR method (redrawn from [45]).

Figure 6. Example of common-offset radargram obtained through the GPR method, for a subsoil with three buried objects. The presence of the objects is inferred from the reflected traces in the shape of hyperbolas.
The block diagram of a typical baseband GPR hardware system is shown in Fig. 7. Generally, a successful system design should meet the following simultaneous requirements [54]: (i) efficient coupling of the EM energy between antenna and ground; (ii) adequate penetration relative to the target depth; (iii) sufficiently large return signal for detection; and (iv) adequate bandwidth for the desired resolution and noise control.

Figure 7. Block diagram showing operation of a typical baseband GPR system (redrawn from [44]).

<table>
<thead>
<tr>
<th>Material</th>
<th>Typical desired penetration depth</th>
<th>Approximate maximum frequency of operation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cold pure fresh-water ice</td>
<td>10 km</td>
<td>10 MHz</td>
</tr>
<tr>
<td>Temperate pure ice</td>
<td>1 km</td>
<td>2 MHz</td>
</tr>
<tr>
<td>Saline ice</td>
<td>10 m</td>
<td>50 MHz</td>
</tr>
<tr>
<td>Fresh water</td>
<td>100 m</td>
<td>100 MHz</td>
</tr>
<tr>
<td>Sand (desert)</td>
<td>5 m</td>
<td>1 GHz</td>
</tr>
<tr>
<td>Sandy soil</td>
<td>3 m</td>
<td>1 GHz</td>
</tr>
<tr>
<td>Loam soil</td>
<td>3 m</td>
<td>500 MHz</td>
</tr>
<tr>
<td>Clay (dry)</td>
<td>2 m</td>
<td>100 MHz</td>
</tr>
<tr>
<td>Salt (dry)</td>
<td>1 km</td>
<td>250 MHz</td>
</tr>
<tr>
<td>Coal</td>
<td>20 m</td>
<td>500 MHz</td>
</tr>
<tr>
<td>Rocks</td>
<td>20 m</td>
<td>50 MHz</td>
</tr>
<tr>
<td>Walls</td>
<td>0.3 m</td>
<td>10 GHz</td>
</tr>
</tbody>
</table>

Table 1. Typical Frequencies for Different GPR Applications  

The working frequency of typical GPR ranges from a few tens of megahertz to several gigahertz, depending on the application. The usual tradeoff holds: The wider the bandwidth, the higher the depth range (co-range) resolution but the shallower the penetration depth. A good choice is usually a tradeoff
Electromagnetic subsurface remote sensing

between resolution and depth. Soil properties are also critical in determining the penetration depth. It is observed experimentally that attenuation in different soils can vary substantially. For example, dry desert and nonporous rocks have very low attenuation (about 1 dBm−1 at 1 GHz) while the attenuation of sea water can be as high as 300 dBm−1 at 1 GHz. Some typical applications and preferred operating frequencies are listed in Table 1 [54].

The depth range of the GPR signal depends on the energy loss during the EM wave propagation. Two factors are associated to that loss: geometrical spreading caused wave propagation and attenuation by soil absorption. In GPR systems and for typical soils, the electrical conductivity of the soil is the primary factor affecting the wave field attenuation. The dielectric permittivity, on the other hand, primarily affects the phase velocity of propagation of the waves [60] and hence the two-way delay time of the return signal. The effect of the magnetic permeability is not considered for typical GPR measurements because the former does not vary significantly in relation to magnetic permeability of free space at typical GPR frequencies and soils [12], [61]. At GPR frequencies, the phase velocity \( v \) (Eq. 36) and attenuation constant \( \alpha \) can be written as a function of the permittivity and conductivity of the ground [62] as follows

\[
\alpha = \frac{\sigma}{2} \sqrt{\frac{\mu}{\varepsilon}}. \tag{38}
\]

The reflected signal is caused by discontinuities in the electrical properties between two layers in the subsoil or between a buried target and the surrounding subsoil. For a normal incidence signal (that is, with vertical direction of propagation or perpendicular of surface) and taking into account both displacement and conduction currents, the GPR reflection coefficient between two layers 1 and 2 in the subsoil can be expressed as [63], [64], [65]

\[
r_{GPR} = \frac{\sqrt{\sigma_1 + i\omega\varepsilon_1} - \sqrt{\sigma_2 + i\omega\varepsilon_2}}{\sqrt{\sigma_1 + i\omega\varepsilon_1} + \sqrt{\sigma_2 + i\omega\varepsilon_2}}. \tag{39}
\]

Further details on mathematical aspects of the GPR method, and information specific to each of its applications can be found in references [62], [63], [66].

The most popular GPR data acquisition techniques can be divided into three basic types (Fig. 8): (i) reflection profiles with constant spacing (common-offset) (Fig. 8a), (ii) velocity surveys such as common mid-point (CMP) in Fig. 8b or wide-angle reflection and refraction (WARR or common-source) in Fig. 8c, and (iii) tomography techniques. Velocity surveys are used to estimate the velocity of the GPR wave signal in the medium in order to convert the two-way delay time of reflection into depth profiles and check if the reflected signal is potentially from subsurface geological targets or from an interference source. Both CMP and WARR are used to obtain an estimate of the velocity of radar wave by varying the spacing between antennas and hence the two-way travel time at a given fixed location. In the CMP technique, the distance between transmitter and receiver antennas is increased in opposite directions, starting from a fixed center point. In WARR technique, one of the antennas is kept fixed (typically the transmitter or common-source setup) while the other is space successively apart. In addition to CMP and WARR, there are two other ways to determine the speed of propagation of electromagnetic wave in the medium: By knowing the dielectric constant of the medium a priori (possibly from independent measurements) and using equation (36), or else by using geological wells or trenches. In the latter case,
by knowing the depth $h$ of wells or trenches and the associated two-way travel time for the transmitted wave to be reflected back to the receiving antenna, the wave velocity can be simply estimated as:

$$v = \frac{2h}{\tau}.$$ 

(40)

**Figure 8.** Typical GPR survey types. a) Common-offset. b) Common mid-point (CMP). c) Common-source or WARR (wide-angle reflection and refraction).

Using the medium velocity and the depth of the targets it is possible to perform an interpretation of the radargram and understand anomalies in the subsurface. An example is shown in Fig. 9, where the radargram indicates sloping structures (comprising different geologic layers) and some discrete buried targets at the right end of the profile. In Fig. 9a, the radargram data was collected using a 100 MHz CU-II RAMAC™ GPR tool from MALA Geoscience AB. Others leading commercial GPR system providers include Ingegneria Dei Sistemi (IDS) SpA, Sensors & Software Inc., US Radar Inc., among others.

To meet the different application requirements, a variety of types of GPR signals are used. Broadly speaking, these can be classified in the following four categories: amplitude modulation (AM), frequency modulated continuous wave (FMCW), continuous wave (CW), and short-pulse or ultra-wideband (UWB). Next, we will briefly discuss the advantages and limitations of each of these types of GPR signals.
Electromagnetic subsurface remote sensing

Figure 9. Example of (a) common mid-point radargram (left) with respective interpretation (right), and (b) common-offset radargram (top) with respective interpretation (bottom).

There are two types of AM transmission used in GPR. For investigation of low-conductivity medium, such as ice and fresh water, a pulse modulated carrier is preferred \[67, 68\]. The carrier frequency can be chosen as low as tens of megahertz. Since the reflectors are well spaced, a relatively narrow transmission bandwidth is needed. The receiver signal is demodulated to extract the pulse envelope. For shallow and high-resolution applications, such as the detection of buried artifacts, a baseband pulse is preferred to avoid the problems caused by high soil attenuation, since most of the energy is in the low-frequency band. A pulse train with a duration of 1 to 2 ns, a peak amplitude of about 100 V, and a repetition rate of 100 kHz is applied to the broadband antenna. The received signal is downconverted by sampling circuits before being displayed. There are three primary advantages of the AM scheme: (i) It provides a real-time display without the need for subsequent signal processing; (ii) the measurement time is short; and (iii) it is implemented with small equipment but without synthesized sources and hence is cost effective. But for the AM scheme, it is difficult to control the transmission spectrum, and the signal-to-noise ratio (SNR) is not as good as that of the FMCW method.

For the FMCW scheme, the frequency of the transmitted signal is continuously swept, and the receiver signal is mixed with a sample of transmitted signals. The Fourier transform of the received signal results in a time domain pulse that represents the receiver signal if a time domain pulse were transmitted. The frequency sweep must be linear in time to minimize signal degradation, and a stable output is required to facilitate signal processing. The major advantage of the FMCW scheme is easier control of the signal spectrum; the filter technique can be applied to obtain better SNR. A shortcoming of the FMCW system is the use of a synthesized frequency source, which means that the system is expensive and bulky. Additional data processing is also needed before the display \[69, 70\].

A continuous wave scheme was used in the early development of GPR, but now it is mainly employed in synthetic aperture and subsurface holography techniques \[71, 72, 73, 74\]. In these techniques, measurements are performed at a single or a few well-spaced frequencies over an aperture at the ground surface. The wave front extrapolation technique is applied to reconstruct the underground region, with the resolution depending on the size of the aperture. Narrowband transmission is used and hence high-speed data capture is avoided. The difficulty of the CW scheme
Electromagnetic subsurface remote sensing comes from the requirement for accurate scanning of the two-dimensional aperture. The operation frequencies should be carefully chosen to minimize resolution degradation [54].

Impulse GPR systems (sometimes referred to as UWB or impulse GPR systems) are preferred in some applications because they allow time-gating of the surface signal. This is the signal reflected by ground surface into the receiver antenna. The surface signal can be much stronger than the reflected signal from buried targets and other structures in the subsoil and make the interpretation of the data difficult, especially for weak or deeply buried targets. In contrast to the direct signal, it is often difficult to achieve adequate system calibration by subtracting this signal a priori because the properties of the soil can vary and may not be known a priori. Alternatively, if short pulses are transmitted, the early time of arrival of the surface signal relative to the reflected signal from the subsoil may allow for these two signals to be well-resolved (separated) in time [59]. For that to occur of course, the difference in these two-way travel times needs to be larger than the temporal pulse width, thus short pulses are required.

Since the spectrum of short pulses is wideband, this type of GPR system employs UWB antennas and hardware capable of handling of broad spectrum of frequencies. As such, they tend to be more expensive than other GPR systems. In addition to time-gating, another advantage of short-pulse GPR systems is that it enables the use of additional target discrimination strategies such as complex frequency resonances [75]. Using appropriate processing techniques, UWB systems also enable more stable images of buried targets when only very limited knowledge about the soil conditions is available [76], [77]. In impulse GPR systems, dispersion effects on the soil affecting the transmitted pulse need to be properly taken into account [59].

Antennas play an important role in GPR system performance [78], [79]. An ideal antenna should introduce the least distortion on the signal spectrum or else one for which the modification can be easily compensated. Unlike the antennas used in the atmospheric radar, the antennas used in GPR should be considered as loaded. The radiation pattern of the GPR antenna can be quite different due to the strong interaction between the antenna and ground. Separate antennas for transmission and reception are commonly used, because it is difficult to make a switch that is fast enough to protect the receiver signal from the direct coupling signal. The direct breakthrough signals will seriously reduce the SNR and hence degrade the system performance. Moreover, in a separate-antenna system, the orientation of antennas can be carefully chosen to reduce further the cross-coupling level. Except for the CW scheme, other modulation types require wideband transmission, which greatly restricts the choice of antenna. Generally, antennas used in GPR systems require broad bandwidth and linear phase in the operating frequency range. Since the antennas work in close proximity to the ground surface, the interaction between them must be taken into account [79]. Four types of antennas, including element antennas, traveling wave antennas, frequency independent antennas, and aperture antennas, have been used in GPR designs.

Element antennas, such as monopoles, cylindrical dipoles, and biconical dipoles, are easy to fabricate and hence widely used in GPR system. Orthogonal arrangement is usually chosen to maintain a low level of cross coupling. To overcome the limitation of narrow transmission bandwidth of thin dipole or monopole antennas, the distributed loading technique is used to expand the bandwidth at the expense of reduced efficiency [80], [81], [82], [83].

Another commonly used antenna type is that comprised of traveling wave antennas such as long wire antennas, V-shaped antennas, and rhombic antennas. Traveling wave antennas distinguish
Electromagnetic subsurface remote sensing

themselves from standing wave antennas in the sense that the current pattern is a traveling wave rather than a standing wave. Standing wave antennas, such as half-wave dipoles, are also referred to as resonant antennas and are narrowband, while traveling wave antennas are broadband. The disadvantage of traveling wave antennas is that half of the power is wasted at the matching resistor [84], [85].

Frequency-independent antennas are often preferred in the impulse GPR system. It has been proved that if the antenna geometry is specified only by angles, its performance will be independent of frequency [79]. In practice, we have to truncate the antenna due to its limited outer size and inner feeding region, which determine the lower bound and upper bound of the frequency, respectively. In general, this type of antenna will introduce nonlinear phase distortion, which results in an extended pulse response in the time domain [54], [78], [79], [86]. A phase correction procedure is needed if the antenna is used in a high-resolution GPR system.

For some GPR systems, higher gain or a more directive radiation pattern is sometimes required. Aperture antennas, such as horn antennas, are preferred because of their large effective area [78]. A ridge design is used to improve the bandwidth and reduce the size. Ridged horns with gain better than 10 dBm over a range of 0.3 GHz to 2 GHz and VSWR lower than 1.5 over a range of 0.2 GHz to 1.8 GHz have been reported [87]. Since many aperture antennas are fed via waveguides, the phase distortion associated with the different modulation schemes needs to be considered.

Signal processing is one of the most important ingredients in GPR systems [88]. Some modulation schemes directly give the time domain data while the signals of other schemes need to be demodulated before the information is available. Signal processing can be performed in the time domain, frequency domain, or space domain. A successful signal processing scheme usually consists of a combination of several kinds of processing techniques that are applied at different stages. Here, we outline some basic signal processing techniques involved in the GPR system.

The first commonly used method is noise reduction by time integration (or time averaging). It is assumed that the noise is random, so that the noise can be reduced to by averaging $N$ identical measurements during some interval. This technique is commonly used in radar systems in general, and it works best for random (white) noise, being less effective to mitigate clutter or interference.

Clutter reduction can be achieved instead by subtracting the mean [89]. This technique is performed under the assumption that the statistics of the underground properties are independent of position. A number of measurements are performed at a set of locations over the same material type to obtain the mean, which can be considered as a measure of the system clutter.

Frequency filtering techniques are commonly used in FMCW systems. Signals that are not in the desired information bandwidth are filtered out (rejected). Thus the SNR of FMCW scheme is usually higher than that of the AM scheme. In some very lossy soils, the return signal is highly attenuated, which makes interpretation of the data difficult. If the attenuation information is available, the results can be improved by exponentially weighting the time traces to counter the decrease in signal level due to the loss [90]. In practice, this is done by using a specially designed amplifier. Caution is needed when using this method, since the noise can also increase in such a system [54], [90].
4 Magnetotelluric Methods

The basic principle of the magnetotelluric (MT) method is to use natural electromagnetic fields to investigate the electrical conductivity structure of the earth. This method was first proposed by Tikhonov in 1950 [91], who assumed the earth’s crust as a planar layer of finite conductivity lying upon an ideally conducting substrate. In this case, a simple relation between the horizontal components of the electric and magnetic fields at the surface can be found as [92]:

\[ i\mu_0 \omega H_x \approx E_y \gamma \cosh(\gamma l) \]  

where \[ \gamma = (i\sigma \mu_0 \omega)^{1/2} \].

The author used data observed at Tucson, Arizona and Zui, Russia to compute the value of conductivity and thickness of the crust that best fit the first four harmonics. For Tucson, the conductivity and thickness were about \( 4.0 \times 10^{-3} \) S/m and 1000 km, respectively. For Zui, the corresponding values are \( 3.0 \times 10^{-1} \) S/m and 100 km.

The MT method distinguishes itself from other subsurface EM methods because very low frequency natural sources are used. The actual mechanisms of natural sources have been under discussion for a long time, but it is now well accepted that the sources of frequency above 1 Hz are thunderstorms while the sources below 1 Hz are due to the current system in the magnetosphere caused by solar activity. In comparison with other EM methods, the use of a natural source is a major advantage. The frequencies used range from 0.001 Hz to \( 10^4 \) Hz, and thus very large investigation depths can be achieved, ranging from about 50 m to 100 m up to several kilometers. Since no transmitters are required, hardware installation is much simpler and has less impact on the environment. The MT method has also proved very useful in some extreme areas where conventional seismic methods are expensive or ineffective. The main shortcomings of the MT method are limited resolution and difficulty in achieving a high SNR, especially in electrically noisy areas [93].

In MT measurements, time-varying horizontal electrical and magnetic fields at the surface are recorded simultaneously. The data recorded in the time domain are first converted into frequency domain data by using a fast Fourier transform (FFT). An apparent conductivity is then defined as a function of frequency. To interpret the data, theoretical apparent conductivity curves are generated by the model studies. The model whose apparent conductivity curve best matches the measurement data is taken as an approximate model of the subsurface. Since it is more convenient and meaningful to represent the apparent conductivity in terms of skin depth, we first introduce the concept of skin depth by studying a simple case. The model we use is shown in Fig. 10, which consists of a homogeneous medium with conductivity \( \sigma \) and a uniform current sheet flowing along the \( x \) direction in the \( xy \) plane. If the density of current at the ground (\( z = 0 \)) is represented as [94]

\[ I_x = \cos \omega t, \quad I_y = I_z = 0 \]  

then the current density at depth \( z \) is

\[ I_x = e^{-z\sqrt{2\omega\mu_0\sigma}/2} \cos(\omega t - z\sqrt{2\omega\mu_0\sigma}), \quad I_y = I_z = 0. \]
When \( z \) increases, the amplitude of the current decreases exponentially with respect to \( z \); meanwhile, the phase delay progressively increases. To describe the amplitude attenuation, the skin depth \( \delta \) as defined in Eq. 22 is repeated below for convenience:

\[
\delta = \frac{2}{\sqrt{\omega \mu \sigma}}.
\]  

(45)

As noted earlier, the skin depth describes the distance after which the current amplitude decreases to \( e^{-1} \approx 0.37 \) of the current amplitude at the surface [94]. Since the typical units of meters in Eq. 45 is not convenient at typical MT frequencies, some prospectors prefer using the following formula:

\[
\delta = \frac{1}{2\pi} \sqrt{10\rho T}
\]  

(46)

where \( T \) is the period in seconds, \( \rho \) is the resistivity in \( \Omega/m \), where \( \delta \) is now expressed in km. For example, if the underground resistivity is 10 S/m and the period of the wave is 3 s, the skin depth is 2.76 km. Subsurface methods other than MT seldom have such large penetration depths.

The data interpretation of the MT method is based on model studies where the earth is modeled as a layered medium. For simplicity, we considered here a two- or three-layer medium. For a two-layer model as shown in Fig. 11, the general expression for the field can be written as [94]

\[
0 \leq z \leq h:
\]

\[
E_z = Ae^{i\sqrt{\sigma_1}z} + be^{-i\sqrt{\sigma_1}z}
\]

(47a)

\[
H_y = e^{i\pi/4} \sqrt{2\sigma_1 T} \left[ -Ae^{i\sqrt{\sigma_1}z} + Be^{-i\sqrt{\sigma_1}z} \right]
\]

(47b)

\[
h \leq z \leq \infty:
\]

\[
E_x = e^{-a\sqrt{\sigma_2}z}
\]

(48a)

\[
H_x = e^{i\pi/4} \sqrt{2\sigma_2 T} e^{-a\sqrt{\sigma_2}z}
\]

(48b)

where \( h \) is the thickness of upper layer, and \( \sigma_1, \sigma_2 \) are the conductivities of the upper and lower layers, respectively. Matching the boundary conditions at \( z = h \), we have

\[
A = \sqrt{\sigma_1} - \sqrt{\sigma_2} e^{-ah(\sqrt{\sigma_1}+\sqrt{\sigma_2})}
\]  

(49)
Electromagnetic subsurface remote sensing

\[ B = \frac{\sqrt{\sigma_1} + \sqrt{\sigma_2}}{2\sqrt{\sigma_1}} e^{ah(\sqrt{\sigma_1} - \sqrt{\sigma_2})}. \]  

(50)

\[ \begin{array}{c|cc}
Z = 0 & O \\
\hline
h & \text{First layer } \sigma_1 \\
Z = h & \text{Second layer } \sigma_2 \\
\end{array} \]

**Figure 11.** Two-layer model of the earth’s crust, used to demonstrate the responses of the magnetotelluric method.

Since we are interested in the ratio between the \( E \) and \( H \) field on the surface, using Eq. 47 at \( z = 0 \) we obtain

\[ \frac{E_y}{H_y} = \frac{1}{\sqrt{2\sigma_1T}} M N e^{-i(\pi/4 + \phi + \psi)} \]

where:

\[ M \cos \phi = \left( \frac{1}{\delta_1} \cosh \frac{h}{\delta_1} + \frac{1}{\delta_2} \sinh \frac{h}{\delta_1} \right) \cos \frac{h}{\delta_1} \]  

(52a)

\[ M \sin \phi = \left( \frac{1}{\delta_1} \sinh \frac{h}{\delta_1} + \frac{1}{\delta_2} \cosh \frac{h}{\delta_1} \right) \sin \frac{h}{\delta_1} \]  

(52b)

\[ M \cos \psi = \left( \frac{1}{\delta_1} \sinh \frac{h}{\delta_1} + \frac{1}{\delta_2} \cosh \frac{h}{\delta_1} \right) \cos \frac{h}{\delta_1} \]  

(52c)

\[ M \sin \psi = \left( \frac{1}{\delta_1} \cosh \frac{h}{\delta_1} + \frac{1}{\delta_2} \sinh \frac{h}{\delta_1} \right) \sin \frac{h}{\delta_1} \]  

(52d)

where \( \delta_1 \) and \( \delta_2 \) are the skin depths of upper and lower layers, respectively. In a multilayer medium, after applying the same procedure, we can obtain exactly the same relation between \( E_x \) and \( H_y \) as shown in Eq. 51 except that the expressions become more complicated. Because of this similarity, we have

\[ \left| \frac{E_x}{H_y} \right| = \frac{1}{\sqrt{2\sigma_a T}} = \frac{M}{N} \frac{1}{\sqrt{2\sigma_1 T}} \]  

(53)

where \( \sigma_a \) is defined as the apparent conductivity. If the medium is homogeneous, the apparent conductivity is equal to the true conductivity. In a multilayer medium the apparent conductivity is an average effect from all layers.
To obtain a better understanding of the preceding formulas, we first consider two two-layer models with an upper conductive segment and a lower resistive segment, and their corresponding apparent conductivity curves as shown in Fig. 12(a) [95]. At very low frequencies, the wave can easily penetrate through upper layer, and thus its conductivity has little effect on the apparent conductivity. Consequently, the apparent resistivity approaches the true resistivity of lower layer. As the frequency increases, less energy can penetrate the upper layer due to the skin effect, and thus the effect from the upper layer becomes dominant. As a result, the apparent resistivity is asymptotic to $\rho_1$. Comparing the two curves, we note that both of them change smoothly, and for the same frequency, case A has lower apparent resistivity than case B, since the conductive sediment of case B is thicker.

![Diagram](image.png)

**Figure 12.** Diagrammatic apparent resistivity curves for (a) two-layer and (b) three-layer models shown (redrawn from [84]).

The next example is a three-layer model as shown in Fig. 12(b) [95]. In this case, the center layer is more conductive than the two adjacent ones. As expected, the curve approaches $\rho_1$ and $\rho_3$ at each end. The existence of the center conductive bed is obvious from the curve, but the apparent resistivity never reaches the true resistivity of center layer $\rho_2$, since its effect is averaged by the effects from the other two layers.

So far we have only discussed the horizontally layered medium, which is a one-dimensional model. In practice, two-dimensional or even three-dimensional structures are often encountered. In a two-dimensional case, the conductivity changes not only along the z direction but also along one of the horizontal directions. The other horizontal direction is called the “strike” direction. If the strike direction is not in the x or y direction, we obtain a general relation between the horizontal field components as [95]

$$E_x = Z_{xx} H_x + Z_{xy} H_y$$  \hspace{1cm} (54a)

$$E_y = Z_{yx} H_x + Z_{yy} H_y.$$  \hspace{1cm} (54b)

Since $E_x$, $E_y$, $H_x$, and $H_y$ are generally out of phase, the $Z_{ij}$ factors are complex numbers. It can also be shown that $Z_{ij}$ have the following properties:
\[ Z_{xx} + Z_{yy} = 0 \]  \hfill (55)

\[ Z_{xy} - Z_{yx} = \text{constant} . \]  \hfill (56)

A simple vertical layer model and its corresponding curves are shown in Fig. 13 [95]. In Fig. 13(b), the apparent resistivity with respect to \( E_\parallel \) changes slowly from \( \rho_1 \) to \( \rho_2 \) due to the continuity of \( H_\perp \) and \( E_\parallel \) across the interface. On the other hand, the apparent resistivity corresponding to \( E_\perp \) has an abrupt change across the contact, since the \( E_\perp \) is discontinuous at the interface. The relative amplitude of \( H_\perp \) varies significantly around the interface and approaches a constant at a large distance, as shown in Fig. 13(d). This is caused by the change in current density near the interface, as shown in Fig. 13(f). We also observe that \( H_z \) appears near the interface, as shown in Fig. 13(c). The reason is that the partial derivative of \( E_\parallel \) with respect to \( \perp \) direction is nonzero.

**Figure 13.** Diagrammatic response curves for a simple vertical contact at frequency \( f \) (redrawn from [84]).

We have discussed the responses in some idealized models. For more complicated cases, their response curves can be obtained by forward modeling. Since the measurement data are in the time domain, we need to convert them into the frequency domain data by using a Fourier transform. In practice, five components are measured. There are four unknowns in Eqs. 56(a) and 56(b), but only two equations. This difficulty can be overcome by making use of the fact that \( Z_{ij} \) changes very slowly with frequency. In fact, \( Z_{ij} \) is computed as an average over a frequency band that contains several frequency sample points. A commonly used method is given in [96], according to which Eq. 56(a) is rewritten as
\[ \langle E_x A^* \rangle = Z_{xx} \langle H_x A^* \rangle + Z_{xy} \langle H_y A^* \rangle \] (57)

and

\[ \langle E_y B^* \rangle = Z_{xx} \langle H_x B^* \rangle + Z_{xy} \langle H_y B^* \rangle \] (58)

where \( A^* \) and \( B^* \) are the complex conjugates of any two of the horizontal field components. The cross powers are defined as

\[ \langle AB^* \rangle (\omega) = \frac{1}{\Delta \omega} \int_{\omega - (\Delta \omega/2)}^{\omega + (\Delta \omega/2)} AB^* d\omega. \] (59)

There are six possible combinations, and the pair \((H_x, H_y)\) is preferred in most cases due to its greater degree of independence. Solving Eqs. 58 and 59, we have

\[ Z_{xx} = \frac{\langle E_x A^* \rangle \langle H_y B^* \rangle - \langle E_y B^* \rangle \langle H_y A^* \rangle}{\langle H_x A^* \rangle \langle H_y B^* \rangle - \langle H_y B^* \rangle \langle H_x A^* \rangle} \] (60a)

and

\[ Z_{xy} = \frac{\langle E_y A^* \rangle \langle H_x B^* \rangle - \langle E_x B^* \rangle \langle H_x A^* \rangle}{\langle H_y A^* \rangle \langle H_x B^* \rangle - \langle H_x B^* \rangle \langle H_y A^* \rangle}. \] (60b)

By applying the same procedure to Eq. (55b), we have

\[ Z_{yx} = \frac{\langle E_x A^* \rangle \langle H_y B^* \rangle - \langle E_y B^* \rangle \langle H_y A^* \rangle}{\langle H_x A^* \rangle \langle H_y B^* \rangle - \langle H_y B^* \rangle \langle H_x A^* \rangle} \] (60c)

and

\[ Z_{yy} = \frac{\langle E_y A^* \rangle \langle H_x B^* \rangle - \langle E_x B^* \rangle \langle H_x A^* \rangle}{\langle H_y A^* \rangle \langle H_x B^* \rangle - \langle H_x B^* \rangle \langle H_y A^* \rangle}. \] (60d)

After obtaining \( Z_{ij} \), they can be substituted into Eqs. 54(a) and 54(b) to solve for the other pair \((E_x, E_y)\), which is then used to check the measurement data. The difference is due either to noise or to measurement error. This procedure is usually used to verify the quality of the measured data.

### 5 Airborne Electromagnetic Methods

Airborne EM methods (AEM) are widely used in geological surveys and prospecting for conductive ore bodies. These methods are suitable for large area surveys because of their speed and cost effectiveness. They are also preferred in some areas where access is difficult, such as swamps or ice-covered areas. In contrast to ground EM methods, airborne EM methods are usually used to outline large-scale structures. Airborne EM methods are preferred for more detailed investigations [97]. The difference between airborne and ground EM systems results from the technical limitations inherent in the use of aircraft. The limited separation between transmitter and receiver determines the shallow investigation depth, usually from 25 m to 75 m. Even though greater penetration depth can be achieved by placing the transmitter and receiver on different aircraft, the disadvantages are obvious. The transmitters and receivers are usually 200 ft to 500 ft above the surface. Consequently, the amplitude ratio of the primary field to the secondary field becomes very small and thus the resolution of airborne EM methods is not
Electromagnetic subsurface remote sensing

very high. The operating frequency is usually chosen from 300 Hz to 4000 Hz. The lower limit is set by the transmission effectiveness, and the upper limit is set by the skin depth. Based on different design principles and application requirements, many systems have been built and operated all over the world since 1940s. Despite the tremendous diversity, most AEM systems can be classified in one of the following categories according to the quantities measured: phase component measuring systems, quadrature systems, rotating field systems, and transient response systems [98].

In phase component measuring AEM systems, the in-phase and quadrature components are measured at a single frequency and recorded as parts per million (ppm) of the primary field. In the system design, vertical loop arrangements are preferred, since they are more sensitive to the steeply dipping conductor and less sensitive to the horizontally layered conductor [99]. Accurate maintenance of transmitter-receiver separation is essential and can be achieved by fixing the transmitter and receiver at the two wing tips. Once this requirement is satisfied, a sensitivity of a few part per million (ppm) can be achieved [98]. A diagram of the phase component measuring system is shown in Fig. 14 [99]. A balancing network associated with the reference loop is used to buck the primary field at the receiver. The receiver signal is then fed to two phase-sensitive demodulators to obtain the in-phase and quadrature components. Low-pass filters are used to reject very-high-frequency signals that do not originate from the earth. The data are interpreted by matching the curves obtained from the modeling. Some response curves of typical structures are provided in [100].

Quadrature AEM systems employ a horizontal coil placed on the airplane as a transmitter and a vertical coil towed behind the plane as a receiver. The vertical coil is referred to as a “towed bird.” Since only the quadrature component is measured, the separation distance is less critical. To reduce the noise further, an auxiliary horizontal coil, powered with a current 90 degrees out of phase with respect to the main transmitter current, is used to cancel the secondary field caused by the metal body of the aircraft.

Figure 14. Block diagram showing operation of a typical phase component measuring system (redrawn from [84]).
Since the response at a single frequency may have two interpretations, two frequencies are used to eliminate the ambiguity. The lower frequency is about 400 Hz and the higher one is chosen from 2 kHz to 2.5 kHz. The system responses in different environments can be obtained by model studies. Reference [101] gives a number of curves for thin sheets and shows the effects of variation in depth, dipping angle, and conductivity.

In AEM systems in general, it is hard to control the relative rotation of receiver and transmitter. The rotating field method has been introduced to overcome this difficulty. Two transmitter coils are placed perpendicular to each other on the plane, and a similar arrangement is used for the receiver. The two transmitters are powered with current of the same frequency shifted 90 degrees out of phase, so that the resultant field rotates about the axis, as shown in Fig. 15 [102]. The two receiver signals are phase shifted by 90 degrees with respect to each other, and then the in-phase and quadrature differences at the two receivers are amplified and recorded by two different channels. Over a barren area, the outputs are set to zero. When the system is within a conducting zone, anomalies in the conductivity are indicated by nonzero outputs in both the in-phase and quadrature channels. The noise introduced by the fluctuation of orientation can be reduced by this scheme, but it is relatively expensive and the data interpretation is complicated by the complex coil system [102].

The fundamental challenge of AEM systems is the difficulty in detecting the relatively small secondary field in the presence of a strong primary field. This difficulty can be alleviated by using the transient field method. A well-known system based on the transient field method is the so-called INPUT (INDuced PULsed Transient) method [103], which was designed by Barringer during the 1950s. In the INPUT system, a large horizontal transmitting coil is placed on the aircraft and a vertical receiving coil is towed in the bird with the axis aligned with the flight direction.

![Figure 15. Working principle of the rotary field AEM system (redrawn from [94]).](image)

The working principle of INPUT is shown in Fig. 16 [104]. A half sine wave with a duration of about 1.5 ms and quiet period of about 2.0 ms is generated as the primary field, as shown in Fig. 16(a). If there are no conducting zones, the current in the receiver is induced only by the primary field, as shown in Fig. 16(b). In the presence of conductive anomalies, the primary field will induce an eddy current. After the primary field is cut off, the eddy current decays exponentially. The duration of the eddy current is proportional to the conductivity anomalies, as shown in Fig. 16(c), that is, the duration times
Electromagnetic subsurface remote sensing becomes longer for higher conductivities. The decay curve in the quiet period is sampled successively in time by six channels and then displayed on a strip, as shown in Fig. 17. As we can see, the distortion caused by a good conductor appears in all the channels, while the distortion corresponding to a poor conductor only registers on the early channels.

Since the secondary field can be measured more accurately in the absence of the primary field, transient systems provide greater investigation depths, which may reach 100 m under favorable conditions. In addition, they can also provide a direct indication of the type of conductor encountered [102]. On the other hand, this system design gives rise to other problems inherent in the transient method. Since the eddy current in the quiet period becomes very small, a more intense source has to be used in order to obtain the same signal level as that in continuous wave method. The circuitry for the transient system is much more complicated, and it is more difficult to reject the noise due to the wideband property of the transient signal.
6 Inductive electromagnetic method

The inductive electromagnetic (IEM) method was developed in Sweden between 1919 and 1922, when Lundberg and Sundberg began working with electric exploration using variable fields and observing the magnetic field strength [99]. In its present form, the basic principles of the IEM method were established between 1925 and 1940 by Sundberg and Hedström. Among the attractive aspects of IEM methods are simple and fast operation, since induction process does not require direct contact with the ground, and very low-cost equipment. The IEM shares common aspects with both the borehole induction method and GPR seen before. Similarly to GPR, the IEM hardware can be deployed in mobile units above the surface to interrogate the earth properties just underneath it. Similarly to borehole induction method, it is based on the excitation of secondary (eddy) current in the subsurface.

The IEM method provides a measurement of the electrical conductivity and magnetic susceptibility of regions or objects in the subsurface near the transmitter. The IEM transmitted employs one or more coil antennas, wherein AC currents generate a primary (or inductor) electric and magnetic fields. As in the case of borehole induction methods, this primary field induces a flow of electric currents (eddy currents) in conductive media in the subsurface. This induced current flow creates a secondary magnetic field is created, which depends on the medium conductivity, as illustrated in Fig. 18. The resulting secondary field can be measured by receiving coil antenna(s).

![Figure 18. Schematic illustration of the physical mechanism behind IEM.](image)

In general, the secondary field is a complex function that depends on the spacing s between the transmitter and receiver coils, the operating frequency \( f \), and the conductivity of the medium \( \sigma \) [105], and details of the geometry and relative alignments between the coils. In certain operational conditions under low induction values defined below, the ratio between the secondary and primary magnetic fields is given by:

\[
\frac{H_S}{H_P} \approx \frac{\omega \mu_0 \sigma s^2}{4},
\]

(61)
Electromagnetic subsurface remote sensing

which is directly proportional to the conductivity of the medium. Under these conditions, it is possible to determine the conductivity, for low to moderate values, by solving for \( \sigma \) in the above equation. An important restriction for this procedure occurs in areas with high conductivity values (hundreds of mS/m or larger), wherein the magnetic field response is not linear with respect to \( \sigma \). Also, for very low conductivity values, the response becomes weak and the measurement may not be accurate due to low signal-to-noise ratio.

The magnetic susceptibility \( \chi_m \) is an intrinsic feature of each material and is associated with the ability of the material to acquire magnetization. In the IEM method, magnetic susceptibility values can be obtained from the knowledge of both the magnetic field ratio and the conductivity using the relation below [105]:

\[
\frac{H_S}{H_P} \approx \frac{2\sqrt{2}}{15} S^3 (\omega \mu_0 \sigma)^{3/2}
\]  

(62)

![Figure 19. Example of inductive electromagnetic data. (a) Apparent electrical conductivity (mS/m). (b) Magnetic susceptibility (ppt).](image)

Based on eq. (62) and by utilizing two successive measurements of the secondary magnetic field for an IEM transmitter in the vertical magnetic dipole (VMD) orientation (equivalent to the horizontal electrical current loops depicted in Fig. 18) at heights 0 m and \( d \) m above the surface, two respective values for \( \sigma \) from Eq. 62 can be obtained, denoted as \( \sigma_{0 \text{VMD}} \) and \( \sigma_{d \text{VMD}} \). Their difference is denoted as \( \Delta \sigma = \sigma_{0 \text{VMD}} - \sigma_{d \text{VMD}} \).

(63)

\[
\chi_m = 58 \cdot 10^{-6}(\sigma_{0 \text{VMD}} - \sigma_{d \text{VMD}})
\]

(64)

An example of a subsurface image obtained from IEM data can be seen in Fig. 19. This data was collected using the ground conductivity meter EM38B™ manufactured by Geonics Ltd. This is the top view of a 4 m by 6 m survey area, where the IEM measurements were taken every 0.2 m apart in both
directions. We can clearly identify anomalous regions with high values of conductivity in Fig. 19(a) and susceptibility in Fig. 19(b). This particular area has a sandy-clay soil with portions of sandstones, and the anomalies are linked with the formation properties.

The IEM method finds applications in many areas including detection of unexploded ordnances [106], archeology surveys [107], [108], soil mapping and precision agriculture [109], [110], soil salinity studies [111], [112], mapping of underground metal pipes [105], [113], among others.

7 Time-domain electromagnetic method

The time-domain electromagnetic (TDEM), also known as transient electromagnetic (TEM) method, was first developed in the Soviet Union during the 1960s. The development of TDEM arose from the need to investigate targets in deeper layers within very low resistivity subsoil. In these conditions, other subsurface EM methods in the frequency domain are unable to achieve a good resolution survey. TDEM is widely used in mining studies, due to its great capacity for penetration and sufficient resolution. A variant of TDEM was discussed before in connection with transient AEM systems.

A common problem with subsurface EM methods in frequency domain is the fact that the secondary magnetic field (produced by the subsoil media) is measured in the presence of the primary magnetic field (generated by the transmitter). Since the primary field is often orders of magnitude more intense than the secondary field, removal of the former during data processing and interpretation is always a challenging task. This process can generate additional noise and distortion in the secondary field data and therefore a loss of precision. One way to avoid interference from the primary field is to use a pulsed source instead of a continuous source and then measure the secondary field as the primary field is switched off. Note that this is a strategy akin to the time-gating seen before in connection to short-pulse GPR systems.

The TDEM method utilizes an electric current circulating through a coil positioned near the ground so that, according to Ampere's Law, it generates an associated magnetic field (primary field). When the current in the transmitter is switched off, the associated primary field also ceases to be excited. However, the shutdown the current is not instantaneous, taking also a brief period of time to the field to reach zero value. The strong variation of the current during this process generates a strong time-varying magnetic flux within the transmitter coil which, from Faraday's law, induces a time-varying magnetic flux to create an electromotive force (e.m.f.) in subsoil. This e.m.f. induces electric currents (eddy currents) in the conductive subsoil media according to Ohm's Law. This induction mechanism is basically the same as described before for borehole induction tools and the IEM method, except that the TDEM method exploits a transient effect due to an abrupt switch off of the current whereas in the other cases induction is produced by AC currents in steady-state. The induced eddy currents in the TDEM circulate initially through the subsoil region very close to the transmitting coil. At later times, the eddy currents diffuse wider and deeper into the subsoil. This effect is called “smoke rings” and illustrated in Fig. 20. As this diffusion process occurs, the eddy currents become progressively weaker.
Electromagnetic subsurface remote sensing

Figure 20. “Smoke rings” effect associated with eddy currents in the subsoil. At successive time instants, the eddy currents penetrate deeper and wider into the subsoil and at the same time becoming progressively weaker, due to the diffusion mechanism (redrawn from [114]).

As explained before, eddy currents also produce an associated (secondary) magnetic field. This secondary magnetic field can be measured by a receiver above ground. Due to Joule's effect, a portion of energy in the subsoil eddy currents is converted into heat, causing it to be attenuated in time domain. The amount of attenuation depends on the resistivity of subsoil medium. The less resistive the medium is, the smaller would be the energy loss in the eddy currents. As a result, it will take a longer time interval for the eddy currents (and hence the secondary field) to be fully attenuated. By measuring the rate of variation of the secondary magnetic field produced by the eddy currents, it is possible to obtain the information about the resistivity of the subsoil.

The electrical current generated by the TDEM transmitter equipment is typically in the form of alternating pulses to avoid static polarization of the medium. The measurement of the secondary magnetic field is made in discretized time intervals called “gates”. This operation principle of TDEM method is illustrated in Fig. 21.

Figure 21. Basic operational principle of time-gating used in the TDEM method. (a) Transmitter current and primary magnetic field. (b) Induced e.m.f. of the primary magnetic field. (c) Secondary magnetic field.
From Maxwell's equations, it is possible to relate the apparent resistivity to the output voltage measured at the receiver coil [115], [116] and use this information to determine the subsoil resistivity at different depths.

There are several different types of TDEM acquisition procedures. The size, position, and relative orientation of the transmitter loop and receiver coil antennas all depend on the particular application. The TDEM is a versatile method, and as such, it can be used on ground, aerial, and aquatic surveys, but it is more common applied for ground and aerial acquisitions. Among the ground (terrestrial) techniques, the most common TDEM configurations are summarized in Fig. 22 and listed below:

- **Single loop**: uses a single loop as transmitter and receiver, as depicted in Fig. 22(a);
- **Coincident loop**: uses different loops as transmitter and receiver but at the same location, as depicted in Fig. 22(b);
- **Tx and Rx loops**: the identical transmitter and receiver loops are separated at a fixed distance, as depicted in Fig. 22(c);
- **Central loop**: the receiver coil is positioned in the center of the transmitter loop. The receiver coil is much smaller than the transmitter loop, as depicted in Fig. 22(d);
- **Loop-loop**: the receiver coil is positioned in a fixed distance from of the transmitter loop. The receiver coil is much smaller than the transmitter loop, as depicted in Fig. 22(e);
- **Off-set**: the receiver coil is positioned in a fixed distance at the corner of the transmitter loop, as depicted in Fig. 22(f);
- **Dual loop**: uses two adjacent loops connected in parallel, as depicted in Fig. 22(g);
- **Moving loop**: used for two-dimensional or three-dimensional acquisitions, where a single loop or central loop is moved along a line, as depicted in Fig. 22(h);
- **Fixed loop**: the position of transmitter loop is fixed, and the receiver coil is moved along a line or area, as depicted in Fig. 22(i);
- **Drill-hole**: the transmitter loop is fixed at the ground level while the receiver coil is moved down inside a borehole;

Figure 22. Different deployments for the TDEM transmitter (Tx) and receiver (Rx) coils (redrawn from [117]).
Noise sources that may impact the interpretation of the TDEM method (Fig. 23) can be classified as galvanic and capacitive [118], [119]. In galvanic coupling, the interfering transmitter induces a current in a conductor that is in galvanic (direct) contact with the ground (such as transmission lines with grounded towers), forming an LR circuit which presents exponential decay with time. Such influence leads to erroneous interpretations of the low-resistivity subsoil layers that can only be identified by comparison with other subsurface EM methods. In capacitive coupling, the interfering transmitter induced currents in a buried cable, for example, thereby forming an LC circuit with oscillating decay easily seen in the data curve.

Figure 23. Noise sources in the TDEM method: (a) galvanic coupling and (b) capacitive coupling (adapted from [119]).

An example of TDEM data with apparent resistivity as a function of time can be seen in Figs. 24 and 25 [120]. The curves correspond to three-layer subsoil, with each layer having different values of resistivity, as shown in the plots on the right side of Figs. 24 and 25. For each TDEM field data provided, a respective geoelectric model can be obtained to furnish geological interpretation.
Electromagnetic subsurface remote sensing

Figure 24. TDEM example response corresponding to a high-low-high resistivity depth profile.

Figure 25. TDEM example response corresponding to a low-high-low resistivity depth profile.

8 Marine Electromagnetic Methods

All electromagnetic systems can be adapted to marine (seaborne) surveys, but the two most used are Magnetotelluric (MT) and marine Controlled-Source Electromagnetic (mCSEM). Like airborne EM acquisitions, seaborne methods are typically applied to large-scale investigations over wide geographical areas. These methods were first developed as research tools in the 1960’s (for MT) and 1970’s (for CSEM) to study the oceanic lithosphere and mantle. The mCSEM was introduced after the MT to help in delineating high resistivity zones of the lithosphere in deep water, since MT would provide only very low amplitude signals for these conditions at high frequencies [121], [122]. Both methods became later widely used also in commercially applications, especially oil and gas prospection. They are both used in tandem with seismic and gravity methods to first map reservoirs and subsequently to monitor oilfields under exploration.

The difference between airborne and seaborne systems is that the marine environment permits separations between transmitter and receiver on the order of tens of kilometers, which is less practical for planes or helicopters unless two or more aircrafts are used at the same time. A critical issue for seaborne surveys is the fact that water has a large electrical conductivity than the seafloor, thus
Electronic subsurface remote sensing screening somewhat the effects from conductive zones below the seafloor. The conductivity of seawater depends on its salinity and temperature. The effect of seawater conductivity on MT data is particularly strong, severely attenuating magnetic field at short periods below 1000 seconds. This limitation of MT has led to the development of mCSEM. Other particularity of seaborne EM methods is related to the position of the receivers since their typical deployment occurs by free fall onto the seafloor. This fact makes it necessary to record their position and orientation. For that purpose, acoustic transponders can be installed on the receivers [123].

The basic theory of the MT technique as discussed previously remains valid for seaborne methods; however, due the direct contact with seawater, the equipment used needs some adaptation to the marine environment. For example, land-based MT surveys use copper-copper sulfate (Cu/CuSO₄) electrodes which would be problematic in the seawater because high chemical gradients across the semi-permeable interface. Instead, marine MT surveys utilize silver/silver chloride (Ag/AgCl) electrodes. Marine MT surveys are traditionally carried with fluxgate or torsion fiber magnetometer sensors and DC coupled electric field sensors deployed directly on the seabed [124]. The sensors have sensitivities of the order of 1 nT, which can provide information about the electrical properties at depths greater than 30 km. For shallower targets, induction coil sensors are also used, since they can be more easily combined with electrical field amplifiers. There are two basic types of measurement deployments used in marine MT systems: (i) the long-wire system, which comprises insulated wires that can reach lengths on the order of 1 km and above and which uses Ag/AgCl electrodes connected to the ends of the wire and to a recording unit; and (ii) the short arm unit, a device with four arms composed by electrodes and salt bridges. Each salt bridge consists of a hollow tube attached to an Ag/AgCl electrode at one end and open to the sea at other [124], [125]. The depth of investigation of marine MT, as well as all electromagnetic methods, is controlled by the skin effect, as described below. Reference [123] presents more details on the typical equipment used for marine MT surveys.

In contrast to marine MT systems, CSEM systems utilize active sensors, making use of transmitter comprised of time-varying electric and magnetic dipole sources to induce currents in conductive materials. The electric or magnetic character of the induced currents can be determined and, as a result, an estimate of the vertical variation of electrical conductivity can be obtained. Electric and magnetic field recorders are deployed on the seafloor, weighed down by environmentally benign anchors, made from standard or degradable concrete. Electromagnetic fields are produced by antennas that can be 50 to 300 meters long and emitting currents of thousands of amps into the seawater. The antenna and transmitter system are towed close to the seafloor to maximize coupling with seafloor rocks and minimize coupling with the air (Fig. 26). Transmission currents are typically binary waveforms with 0.1 to 0.25 Hz fundamental and higher harmonics. Different combinations of source-receiver geometries can be used in mCSEM depending on the equivalent dipole arrangement obtained [125]: (i) vertical electrical dipole (VED); (ii) horizontal electrical dipole (HED); (iii) vertical magnetic dipole (VMD); and (iv) horizontal magnetic dipole (HMD). In general, mCSEM systems can be frequency- or time-domain.
The basic physical principle behind CSEM is the same that inductive electromagnetic method seen before. Summarily, the source sends a time-varying EM signal, the primary field. This time-varying signal induces eddy currents within conducting layers beneath the seafloor, whose behavior depend on electrical properties of the material (resistivity) and on the magnitude and frequency of the source signal. In its turn, the induced eddy currents produce a magnetic field (secondary field) that can be measured by the receivers deployed on the seafloor. The damped wave equation describes how the vector electric field interacts with the electric and magnetic properties of the material:

$$\nabla^2 E = \mu \sigma \frac{\partial E}{\partial t} + \mu \varepsilon \frac{\partial^2 E}{\partial t^2}$$

(65)

where, as before, $\sigma$ is conductivity, $\varepsilon$ is electric permittivity, $\mu$ is magnetic permeability. The vertical resolution is inversely proportional to the wavelength. At frequencies relevant to mCSEM, from 0.1 Hz to 10 Hz and to MT, from 0.0001 Hz to 1 Hz, the second derivative term above is negligible and Eq. 65 reduces to

$$\nabla^2 E = \mu \sigma \frac{\partial E}{\partial t}$$

(66)

which, for a time-harmonic excitation at angular frequency $\omega$ admits solutions of the form

$$E = E_0 e^{-i\beta z} e^{-\alpha z}$$

(67)

where $\alpha$ and $\beta$ are exponential attenuation and phase lag terms over distance $z$, related to the skin depth, see also Eq. 22. As the EM wave penetrates deeper below the seafloor, the eddy currents (and hence the secondary magnetic field) become progressively weaker due to the skin effect. As the skin depth is dependent of the frequency, the MT and CSEM methods exhibit different depths of effective penetration, thus their application will also differ. Table 2 summarizes the differences between marine MT and marine CSEM [126]. In shallow waters, one concern that exists for mCSEM systems is that energy propagating up to the air and back to the receivers (air waves) may become significant. If properly designed though, mCSEM systems can still be used to detect subsurface structures in such
cases. References [121] and [123] present a historical evolution of the mCSEM method and detailed description of equipment used.

Table 2. Comparison between marine magnetotelluric (MT) and controlled-source EM (CSEM) marine methods (adapted from [126]).

<table>
<thead>
<tr>
<th>MT</th>
<th>CSEM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Natural source</td>
<td>Manmade source</td>
</tr>
<tr>
<td>Plane wave, vertically polarization</td>
<td>Localized dipole source</td>
</tr>
<tr>
<td>Basin scale</td>
<td>Reservoir scale</td>
</tr>
<tr>
<td>Sensitive to high conductivity</td>
<td>Sensitive to high resistivity</td>
</tr>
<tr>
<td>Detection of structure</td>
<td>Detection of resistivity contrast</td>
</tr>
</tbody>
</table>

9 Modeling of Subsurface EM Sensors

In order to understand and interpret the response of electromagnetic sensors, particularly in complex scenarios, it is necessary to model electromagnetic fields in the subsurface environment. Such modeling is divided into two basic classes:

- **Forward models** seek to model diffusion and propagation of electromagnetic fields and the consequent sensor response in a given scenario (of assumed geophysical properties) by solving Maxwell’s equations in that scenario.
- **Inverse models** seek to estimate/determine/image the geophysical properties given for a set of known sensor responses.

Inverse modeling provides measurement interpretation and this is more directly linked to the end goal of a remote sensing activity. The forward modeling is an intermediate step often necessitated by the inverse modeling. Both these topics are vast. This Section will focus on briefly describing and discussing forward models only. Discussions on different aspects inverse modeling techniques can be found in other articles on this Encyclopedia such as *Image Reconstruction, Electric Impedance Imaging, Radar Signal Processing* and *Geophysical Signal and Image Processing*.

Analytical closed-form solutions to Maxwell’s equations, which link the geophysical environment’s inhomogeneous electrical properties with the set of observables, can possess distinct, powerful advantages over numerical results obtained via simulation. On the one hand, the analyst can gain insights into the key electromagnetic propagation, guidance, and scatter mechanisms that lead to the predicted observable values (sensor responses), and how material inhomogeneity, electrical conductivity anisotropy, material frequency dispersion, etc. in the subsurface influence the relative dominance of these mechanisms. An important consequence of the gained insights, which is directly relevant to subsurface remote sensing, is that the analyst better understands the (i) geophysical environment properties to which a given sensor’s data is sensitive or non-sensitive (and under what sensor/environment regimes), (ii) sources of ambiguity in sensor data, and (iii) required/desired characteristics of next-generation sensors to produce data sets more robust to ambiguities and more sensitive to the unknown, desired geophysical properties [127]. On the other hand, analytical models can be faster to evaluate numerically, as compared to running brute-force numerical algorithms, due to
their closed-form expressions; the speed advantage of many analytical models is not to be underestimated, especially for inversion (i.e., analytical inverse models), which can be very computationally intensive and long if done numerically [127], [128], [129], [130].

For complex geophysical environments with respect to material spatial inhomogeneity, frequency dispersion, anisotropy, and so on, analytical models often find themselves restricted when it comes to meaningful quantitative predictive capability across a broad range of problems. Owing largely to assumptions made from mathematical tractability considerations in solving Maxwell’s equations, this limitation often manifests, either explicitly or implicitly (e.g., in level of expected versus desired predictive accuracy), in the applicability of analytical models to limiting cases with respect to, for example, sensor geometry (transmitter and receiver locations and orientation), frequency, material anisotropy, and so on [64], [127]. To complement analytical models, particularly for modeling remote sensors in complex geophysical media where analytical models may fall short in terms of robustly predictive accuracy or outright unavailability, numerical modeling of the sensor, simulated using numerical algorithms, can provide the flexibility (i.e., with respect to the sensor and environment properties) and numerical precision needed for forward modeling, geophysical parameter inversion, and sensor evaluation.

Numerical electromagnetic modeling comprises the use of numerical algorithm(s) to solve Maxwell’s equations for a given transmitter (source) and environment geometry to generate numerical results for one or more vector or scalar field quantities, such as the electric field vector, magnetic field vector, or electric scalar potential. Many factors must be considered when deciding upon the numerical algorithm to use. Broadly speaking, one must balance fidelity in the approximation of the sensor’s physical geometry, transmission/reception scenario, and geophysical environment’s characteristics (inhomogeneity, frequency-dispersion, and anisotropy), desired solution precision, required computation speed, and available computation resources. A trade-off analysis must be made to decide on the relative importance of these factors, which will vary based upon the task. Tasks involving dynamic decision-making relying (either directly or indirectly) on the modeler’s output, and/or computationally heavy tasks (e.g., on-the-fly formation conductivity profile inversion and equipment geosteering), may lead to prioritization of faster throughput of formation structural information updates in order to aid timely characterization of the environment’s properties [130]. These dynamic updates can inform, for example, a response to steering subsurface exploration equipment to stay within the most hydrocarbon-productive regions of a formation [8], [131]. On the other hand, in scenarios where the sensor response is expected to exhibit low sensitivity to key formation parameters, higher numerical precision may be required to overcome ill-conditioning of the inverse problem [132].

When numerically modeling remote sensors, one could decide to include (perhaps approximately) or neglect the following, based on balancing the above-mentioned considerations: Exploration borehole, drilling fluid and invasion zone, finite antenna geometry, mandrel, the transmitter’s tugboat (in mCSEM), transient effects in a narrow-band transmission, material anisotropy, two or three-dimensional geophysical inhomogeneities (e.g., unconformities and faults), and so on [8], [133]. What (and how) sensor and environment features are modeled can have a significant impact on the expected reliability of results generated from the numerical modeler. Different ways of approximating the transmit antenna’s geometry and finite size, for example, can lead to numerical
discrepancies between different numerical modelers [8], [134], as well as between numerical sensor response results and laboratory-controlled experimental data [135]. Another important example draws from approximating fine, sub-wavelength features of a geophysical formation, such as alternating sand-shale micro-laminations and natural or drilling-induced formation fractures, which can effectively impart to the formation bulk-scale anisotropic electric conductivity [132], [136]. Neglecting Earth formation anisotropy (i.e., dictated by both the anisotropy ratio[s] and the tensor’s spatial orientation) in the numerical model, and using the model’s results in turn to interpret field data, can have serious consequences on the realized productivity of a reservoir. A classic example, concerning the consequences of neglecting formation conductivity anisotropy, is “low-resistivity pay” in formations composed of finely alternating layers of sand and shale that are perpendicular to the borehole: Namely, the electrically conductive shale micro-laminations dominate the borehole sensor’s “coaxial” (i.e., borehole-aligned transmitter and receiver) measurement response (suggesting low hydrocarbon infiltration), while in fact the formation possesses large quantities of hydrocarbons within the electrically resistive, hydrocarbon-infiltrated sand micro-laminations [132].

Once the computational geometry is decided, there are many numerical algorithms to choose from, whose computer implementations will actually execute the computations and produce the numerical results for subsequent analysis. Broadly speaking, the techniques can be differentiated between whether they solve Maxwell’s equations in the time or frequency domain, as well as what they discretize and how. These characteristics have consequences on a numerical algorithm’s capability to adequately account for propagation, scattering, and guidance effects, on the transmitted electromagnetic wave, arising from the complexity of sensor and environment characteristics stipulated in the numerical model. We name, in particular, algorithms based on Finite Element, Finite Difference, Finite Volume, and Integral Equation techniques, which involve direct volume or surface spatial discretizations of the geometry [19], [24], [137], as well as modal methods (i.e., numerical eigenfunction expansion), which approximate integral eigenfunction expansions as weighted summations and/or (depending on the particular eigenfunction basis employed, and geophysical structure) numerically evaluate discrete eigenfunction summations [10], [64], [138]. One can also cite methods hybridizing spatial and modal discretization and evaluation techniques together. One can cite, for example, combining finite difference and modal techniques to both model more complex conductivity variation along the formation’s depth and exploit the (assumed) formation’s geometry invariance along the transverse plane [139]. Another example worth mentioning involves using the results of modal evaluation, of a layered-medium’s tensor Green’s Functions, in a surface integral equation formulation to analyze responses of sensors in the proximity of objects buried in layered, subsurface anisotropic media [137].

Another important consideration, in selecting a numerical algorithm, is whether to use a time-domain or frequency-domain algorithm; this depends on the transmitted and recorded signals’ bandwidths, and how one wishes to approximate these bandwidths. Frequency-domain methods can be advantageous if the source is expected to be narrow-band, using only one or a few select transmission frequencies, and the sensor’s dwell time is long enough that transient effects can be ignored. Moreover these methods have the distinct advantage of allowing more straightforward simulation, of electromagnetic wave propagation in media of complex frequency-dispersive characteristics, by simply “plugging in” fixed values for each formation’s frequency-dependent material properties [64].
Moreover, if the time-domain solution is required, one can subsequently perform an inverse Fourier transform to recover the transient signal [75]. On the other hand, time-domain methods can be highly advantageous if one expects to use a broadband source such as a ground-penetrating radar, performing just one simulation (rather than one frequency-domain simulation at each sampled harmonic within the signal’s bandwidth, followed by inverse Fourier transform) to capture the transient signal’s propagation, guidance, and scattering behavior [75], [19], [59].

Another consideration is how to perform the discretization of space in Finite Element, Finite Difference, and Finite Volume methods to approximate the (continuously-varying) material and source current profile of the sensor/environment geometry with one represented by a spatially sampled (i.e., discretized) profile [19], [24]. A simple and popular choice of geometrical and material discretization is to assign to each mesh element a single set of constitutive material properties, leading to a piecewise-constant material profile. This is particularly suitable for modeling electromagnetic wave propagation in subsurface formations where sharp gradients in material properties are expected, such as at layer interfaces between rock and loose soil. On the other hand, the algorithm’s spatial meshing scheme can be made to possess enough flexibility, via refining the mesh (i.e., reducing the mesh element sizes), to capture fine-scale and/or more gradual material transitions if need-be. More robust algorithms based on these techniques can even assign to each mesh element very general material anisotropy, allowing the modeling of wave physics within formations characterized by bulk-scale uniaxial and biaxial anisotropy [19], [24]. One disadvantage of these methods is the large volumetric meshes which can result. For implicit, matrix-based methods this also results in large matrices that must inverted directly, or used in repeated rounds of multiplication with the unknown, to-be-solved solution vector to iteratively arrive at a well-converged solution vector; this would be done once per time step in time-domain methods. For explicit, matrix-free methods the fields would be explicitly calculated at the mesh nodes and/or faces (again, once per time step in time-domain methods) [21]. Another issue present in Finite Element, Finite Difference, and Finite Volume methods is the need for mesh truncation. This is because subsurface remote sensing is invariably associated to open-domain problems, whereas the computational mesh can only discretize a finite volume of space. In order to mimic an open-domain problem, absorbing boundary conditions [140], [141], [142] can be utilized to properly truncate the volumetric mesh. Among those, the Perfectly Matched Layer [143], [144], [145] is arguably the most utilized in practice.

Surface integral equation methods alleviate the large matrix sizes of volume-meshing techniques and the need for absorbing boundary conditions by discretizing 2-D surfaces, rather than 3-D volumes [146]. Their major drawbacks can be, however, the generation of dense (rather than sparse) system matrices, as well as the requirement of knowledge of the background formation’s four tensor (dyadic) Green’s functions, which correspond to the vector electric or magnetic fields produced by impulse electric or (equivalent) magnetic current sources [137]. Fast algorithms to accelerate the iterative solution of the dense matrices [147], as well as increasingly rapid, accurate, and robust techniques to calculate the tensor Green’s functions in layered anisotropic media (using modal methods, discussed below) [148], make surface integral equations yet another choice to consider among the space discretization-based numerical algorithms available for flexible modeling of subsurface remote sensors in complex geophysical media.

The flexibility afforded by these space-discretization methods does come at a price, however; increased modeled complexity in the domain can require longer solution times and/or access to
extensive compute resources, both of which may be impractical in scenarios requiring a steady stream of information to interpret and invert measured data, and/or make geophysical exploration decisions, in near real-time. As a trade-off, in scenarios where layered inhomogeneity and material anisotropy are expected to be (at least locally, within the sensor’s proximity) the primary contributors to the relevant wave physics governing the sensor responses, one can resort to modal techniques that involve expressing the measured field as an integral and/or discrete superposition of eigenfunctions supported in the structure [64], [138], [149]. Modal expansion algorithms (particularly, when involving infinite or semi-infinite spectral integrals) encounter problems in computing the electromagnetic near-field, namely slow convergence and poor numerical precision. The poor numerical precision and slow convergence arise when the spectral-domain integrand exhibits highly oscillatory and/or slowly decaying behavior (i.e., versus the wavenumber integration variable[s]), both of whose severities vary significantly versus the sensor and layering geometry [64], [150]. These difficulties encourage the development of “Discrete Complex Image”-based indirect eigenfunction expansion evaluation techniques to obviate the slow, laborious computation of said spectral integrals [151], as well as techniques to accelerate convergence of the near-field solution obtained via direct numerical integration [150]. However, the solution processes of these indirect evaluation methods may not necessarily produce results amenable to rigorous error-control and error-checking, which is a drawback as compared to direct numerical evaluation of the eigenfunction expansions themselves, where one can perform adaptive hp refinement to achieve convergence within a desired threshold [152]. Recent progress however has resulted in algorithms which define sensor and environment material/layering-robust integration paths that eliminate or greatly reduce integrand oscillation and infinite-integral truncation error while imparting rapid exponential and/or algebraic decay, making direct numerical evaluation of the eigenfunction expansions feasible across a broad range of sensor and environment layering/material scenarios [17], [148]. Along with their observed strong low-frequency stability properties, which are important when considering for example CSEM transmitters and borehole sensors operating in the wide frequency range of 0.01Hz-2MHz to achieve deep wave penetration through conductive media [16], [153], [154], the high, robustly-achievable numerical precision, computational speed, and rigorous error-checking makes direct numerical evaluation a serious algorithm candidate to consider along with direct space-discretization ones.

For more complex layered structures, such as those comprising two or more vertically-offset, laterally-adjoined planar-layered structures, or two or more radially offset, vertically-adjoined cylindrically-layered structures, numerical mode matching techniques represent another class of viable modal algorithms to effect prediction of sensor responses [4], [26], [129], [149], [155], [156].

References


45 | Electromagnetic subsurface remote sensing


