Accurate Antenna Models in Ground Penetrating Radar Diffraction Tomography

Meincke, Peter; Kim, Oleksiy S.

Published in:
Proceedings of IEEE Antennas and Propagation Society International Symposium

Link to article, DOI:
10.1109/APS.2002.1016984

Publication date:
2002

Document Version
Publisher's PDF, also known as Version of record

Link back to DTU Orbit

Citation (APA):
Accurate Antenna Models in Ground Penetrating Radar Diffraction Tomography

Peter Meincke and Oleksiy Kim*

Ørsted-DTU, Section for Electromagnetic Systems, Building 348, Technical University of Denmark, DK-2800 Kgs. Lyngby, Denmark, pme@oersted.dtu.dk.

1 Introduction

Linear inversion schemes based upon the concept of diffraction tomography have proven successful for ground penetrating radar (GPR) imaging [1], [2], [3]. In many GPR surveys, the antennas of the GPR are located close to the air-soil interface and therefore, it is important, as illustrated in [2], to incorporate in the inversion scheme the presence of this interface. In [2] the antennas are modeled as ideal (Hertzian) electric dipoles. Since practical GPR antennas are not ideal, it is of interest to investigate the validity of this model.

In this paper we extend the formulation of [2] to hold for arbitrary antennas. For simplicity the 2.5-D case is considered, that is, it is assumed that the scattering object in the soil is invariant in one direction, which for instance is the case for a pipe. The arbitrary antennas are modeled by their plane-wave receiving- and transmitting spectra. We find these spectra numerically for a resistively loaded dipole using the method of moments. Also, we illustrate through a numerical example the importance of taking into account the correct antenna pattern in GPR diffraction tomography.

2 The 2.5-D Forward Model

An example of the GPR configuration involving the planar air-soil interface is shown in Figure 1. A Cartesian $xyz$ coordinate system is introduced such that the $xy$ plane coincides with the interface and such that $z > 0$ is air. An object, which is assumed infinitely long in the $x$ direction, is buried in the soil. The propagation constant of air is $k_0(\omega) = \omega/\sqrt{\varepsilon_0}$ and that of soil is $k_1(\omega) = \omega/\sqrt{\varepsilon_1}$, assuming that the soil is lossless.

The position of the receiving antenna is described by $(z_r, y_r, z_r)$ and that of the transmitting antenna is $(x_t, y_t, z_t) = (x_t + x, y_t + y, z_t)$ with the offset $(x, y, z, 0)$ being fixed. It is assumed that the conductivity $\sigma(y, z)$ is much less than the contrast in permittivity $\Delta \varepsilon(y, z) = \varepsilon(y, z) - \varepsilon_1$, i.e., $\sigma(y, z) \ll \omega \Delta \varepsilon(y, z)$ over the frequency band of interest. Assuming an $x$-directed antenna described by the current density $J_0(x, y, z, \omega)$, the background electric field $E_b$ in the soil can be expressed as

$$
E_b(x, y, z, \omega) = \frac{1}{(2\pi)^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} S_b(k_x, k_y, \omega) \exp(i[k_x(x-x_t) + k_y(y-y_t)])
$$

$$
+ \pi \omega k_0 \delta(\omega) \exp(-i\omega t))
$$

(1)

where the plane-wave transmitting spectrum $S_b$ is [3, 4]

$$
S_b(k_x, k_y, \omega) = \frac{\omega \mu_0}{4} \mathbf{J}_b(k_x, k_y, -\gamma_0, \omega) \cdot \mathbf{P}(-k_x, -k_y, \omega).
$$

(2)

*The Danish Technical Research Council is acknowledged for supporting this work.
Herein, \( \gamma_i = \gamma_i(k_x, k_y, \omega) = \sqrt{k_i^2(\omega) - k_0^2 - k_p^2} \), \( i = 0, 1 \), the expression for the dyadic \( \hat{F}(k_x, k_y, \omega) \) describing the interface can be found in [2, (6)], and \( \hat{J}_0 \) is the Fourier transform of the current density defined as

\[
\hat{J}_0(k_x, k_y, k_z, \omega) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} J_0(x, y, z, \xi, \eta, \omega) \exp(-i[k_x x + k_y y + k_z z]) \, dx \, dy \, dz. \tag{3}
\]

Assuming that the buried object is a weak scatterer, such that it satisfies the Born approximation, the output of the receiving antenna is

\[
s(x, y, z, \omega) = \frac{i\omega^2 \mu_0}{8\pi^2} \int_{-\infty}^{\infty} R(k_x, k_y, \omega) \cdot \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \hat{E}_0(x', y', z', \omega) \Delta \epsilon(x', x) \exp(i(k_x x' + k_y y' + \gamma_0 z' - \gamma_1 z')) \, dx' \, dy' \, dz' \, dk_x \, dk_y \, dk_z, \tag{4}
\]

where \( R \) is the receiving plane-wave spectrum of the antenna. In case of a reciprocal antenna, the receiving spectrum is related to the normalized transmitting spectrum \( S \) (resulting from an input signal of 1 volt) as

\[
R(k_x, k_y, \omega) = \frac{\gamma_1(k_x, k_y, \omega)}{\omega \mu_0 Y_0} S(-k_x, -k_y, \omega) \tag{5}
\]

where \( Y_0 \) is the characteristic admittance of the transmission line connected to the receiving antenna. Replacing \((k_x, k_y)\) with \((k_x', k_y')\) in the expression (1) for the background field and inserting into (4), performing the integration over \( z' \) and \( k_x', k_y' \), carrying out the substitution \( k_y' = k_y + k_0 \), and Fourier transforming with respect to \( y_r \), we arrive at

\[
\hat{s}(k_y, y_r, \omega) = \frac{i\omega^2 \mu_0}{8\pi^2} \int_{-\infty}^{\infty} \frac{R(k_y', k_y + k_0, \omega) \cdot S(k_y', k_y', \omega)}{\gamma_1(k_y', k_y + k_0, \omega)} \exp(i[(\gamma_0(k_y', k_y + k_0, \omega) + \gamma_1(k_y', k_y + k_0, \omega))y_r - \gamma_0(k_y', k_y + k_0, \omega)]y_r) \, dy_r' \, dk_y' \, dk_y' \, dk_y'. \tag{6}
\]

Now we can use the result of [2, Appendix] to calculate the integration over \( k_y' \) and \( k_y'' \) asymptotically as \( z' \to -\infty \). The result is

\[
\hat{s}(k_y, y_r, \omega) \sim D(k_y, \omega) \Delta \epsilon_1(k_y, \sqrt{4k_1^2(\omega) - k_p^2}) \tag{7}
\]

where

\[
D(k_y, \omega) = \frac{-\omega^2 \mu_0}{16\pi k_1(\omega)} R(0, \frac{1}{2} k_y, \omega) \cdot S_0(0, -\frac{1}{2} k_y, \omega) \sqrt{4k_1^2(\omega) - k_p^2} \exp(i\sqrt{4k_1^2(\omega) - k_p^2}y_r + \frac{1}{2} k_y y_r) \tag{8}
\]

and \( \Delta \epsilon_1(y, z) = \Delta \epsilon(y, z)/z \). Using (5), (2), and [2, (6)], it can be shown that

\[
R(0, \frac{1}{2} k_y, \omega) \cdot S_0(0, -\frac{1}{2} k_y, \omega) = \frac{1}{8\pi k_1^2(\omega)} J(0, -\frac{1}{2} k_y, \omega) \cdot J(0, \frac{1}{2} k_y, \omega) \]

\[
\cdot [4k_1^2(\omega) - k_p^2]/[\sqrt{4k_1^2(\omega) - k_p^2} + \sqrt{4k_1^2(\omega) - k_p^2}]^2 \]

where \( J \) is the Fourier transform of the current density describing the normalized transmitting spectrum \( S \) in (5).
3 Inversion

Using the fact that $\Delta \epsilon$ is real, this function can be found from its Fourier transform $\tilde{\Delta \epsilon}$ through

$$\Delta \epsilon(y, z) = \frac{1}{2\pi^2} \text{Re} \left[ \int_{k_z \geq 0} \int \tilde{\Delta \epsilon}(k_y, k_z) \exp(i(k_y y + k_z z)) \, dk_y \, dk_z \right].$$

To use the relation between the measured data $s$ and $\tilde{\Delta \epsilon}$ in (7) we must carry out the substitution $k_z = \sqrt{4k_0^2(\omega) - k_y^2}$ and restrict $k_y$ to the range $|k_y| < 2k_1(\omega)$. Hence,

$$\Delta \epsilon(y, z) = \frac{1}{2\pi^2} \text{Re} \left[ \int_{-2k_1(\omega)}^{2k_1(\omega)} \frac{2k_1(\omega)\sqrt{\mu_0\epsilon_0} \tilde{\delta}(k_y, \omega)}{D(k_y, \omega)k_z} \exp(i(k_y y + k_z z)) \, dk_y \, d\omega \right].$$

4 Numerical Example

The inversion scheme of Section 3 is now tested on synthetic GPR data. Figure 1 shows a two-layer dielectric pipe with outer diameter 24 cm and inner diameter 10 cm located 1 m below the interface. The permittivities of the soil, the inner and outer regions of the pipe are 10\(\varepsilon_0\), 10.26\(\varepsilon_0\), and 10.4\(\varepsilon_0\), respectively. The synthetic GPR data are calculated using the method described in [4]. It is assumed that the radar uses 60 frequencies equally spaced in the range 20 MHz < $f$ < 1.3 GHz. The antennas are assumed to be dipoles of length 2 m, resistively loaded according to the Wu-King profile [5] to increase the bandwidth.

Figures 2 and 3 show two images of $\Delta \epsilon(y, z)/\epsilon_0$ obtained from the inversion formula (10).

Figure 1: The fixed-offset GPR configuration involving a two-layer circular pipe. The antennas are resistively loaded $\hat{x}$-directed dipoles.

In Figure 2 it is assumed that the antennas are Hertzian dipoles. It is seen that the image is highly distorted. In Figure 3 the correct antenna model is used. The current distribution on the antennas in the presence of the air-soil interface is determined using the method of moments. It's observed that the image is very accurate. This example shows that it's an inaccurate approximation to use simple Hertzian dipole models to describe realistic GPR antennas in diffraction tomography.
Figure 2: The image of $\Delta\epsilon(y,z)/\varepsilon_0$. The antennas are modeled as Hertzian dipoles.

Figure 3: The antennas are modeled as resistively loaded dipoles.

References


