Enhancement of Reflected Waves in Single-Hole Polarimetric Borehole Radar Measurement

Takashi Miwa, Motoyuki Sato, *Member, IEEE*, and Hiroaki Niitsuma, *Member, IEEE*

***Abstract*—A polarimetric approach is presented to extract information of reflected waves that is masked by the transmitter–receiver directly coupled wave in a single-hole borehole radar measurement. Radar polarimetry theory is expanded to an omnidirectional radar system with electric and magnetic dipoles arranged on the same axis. First, we formulate the transfer functions directly coupled between the antennas for cross-hole and single-hole arrangements in copolarized channels. We found that the theoretical scattering matrices of the direct-wave coupling is identical to the scattering matrix from a dihedral corner reflector. Second, we also consider signals in polarimetric channel of a wave reflected from a plane scatter in single-hole arrangements. As advanced reflection borehole radar measurement, we demonstrate a technique for both reduction of the directly coupled wave and enhancement of the reflected waves from a plane fracture with measured data in dipole–dipole and slot–slot antenna combinations. For quantitative determination of the scattering matrix, we use a technique to compensate the antenna transfer functions by the time derivative of the directly coupled signals in single-hole measurement. Also, we propose a technique to reduce the directly coupled component by adding vertical (VV) and horizontal (HH) signals and we showed that the directly coupled wave is effectively reduced and reflected waves are enhanced with experimental data. Finally, we show that this technique is more useful for near-range reflector detection than a conventional subtraction technique with moving average of the measured waveforms.**

***Index Terms*—Geologic measurements, radar polarimetry.**

I. INTRODUCTION

GROUND penetrating radars have good applicability to shallow subsurface sensing [1] such as geological survey, archaeological exploration, and detection of objects including man-made mines, buried pipes, and voids under pavement. A transmitter–receiver coupled wave, which is caused by a wave reflected from the ground surface, lateral wave along the surface, and directly coupled wave between antennas, often decrease the detectibility of shallow buried subsurface objects. Isolation of the transmitter and receiver is essential in radar system design. However, it is not easy to achieve this for most ground penetrating radar systems because the transmitter–receiver separation must be small for detection of radar targets located close to antennas. This is more serious in borehole radar because it uses antennas without any shielding structure in a thin borehole. To enhance the reflected wave from radar targets, many techniques have been proposed such as development of antennas matched to surface [2], background subtraction procedure, synthetic aperture technique [3], pulse compression technique, and polarimetric technique [4]. However, it is not easy to apply these techniques to borehole radar measurement because of its low flexibility of the structure of antennas, arrangement of the antenna, and the measurement system. The borehole diameter limits the system configuration of antennas and measurement systems. Generally, the frequency between 10–100 MHz is used in borehole radar systems, which is lower than that in the surface penetrating radar [5]. In this frequency range, we can avoid an effect of guided waves along a borehole which corresponds to the surface reflections for surface penetrating radars. However, the near-field component of a directly coupled wave is dominant in the borehole radar systems over these low frequencies. To reduce the directly coupled wave, commercial borehole radar systems use the antenna separation much larger than the antenna length. Accordingly, the accuracy of the information of radar targets near the borehole is decreased.

As a common signal processing technique, reduction of the directly coupled wave has been achieved by subtraction of the averaged waveforms of adjoining depths from the raw signal. If the medium around the borehole is independent of the depth, this method is effective and can enhance the reflection several meters away from the borehole. However, the resolution is not sufficient to determine the intersecting depth of plane reflectors to borehole and, also, this method has a possibility to suppress the amplitude of the reflection from a point reflector far away from the borehole. This method is suitable for the rough estimation of the depth and the dip of plane reflectors from a radar profile.

In order to reduce the directly coupled wave, we have studied polarimetric approaches. A cross-polarization measurement system with the combinations of dipole antennas and slot antennas on a conducting cylinder was developed and successfully applied to reduction of the directly coupled wave by using orthogonal polarizations [6]. But it reduces the amplitude of not only a direct coupling wave, but also reflected waves because flat fractures depolarize little. We have also developed a full polarimetric borehole radar system and proposed a compensation technique of the antenna transfer functions between the dipole antenna and the slot antenna by inverse filtering by the *in situ* antenna transfer function obtained in hole-to-hole (cross-hole) [7] and single-hole [8] measurement.

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The theoretical polarimetric characteristics have been analyzed in surface penetrating radar systems [9]. So, the reflected waves are successfully enhanced by the transformation of the polarization basis to the optimum polarization state. If we could have know the theoretical scattering matrices in the polarimetric borehole radar system, we could expect to effectively enhance the reflected wave and reduce the direct coupling wave.

Generally, the radar polarimetry has been discussed with both orthogonal electric field components on specific paths illuminated by directional antennas [10], [11]. However, we have to treat a polarimetric theory based on electric and magnetic field radiated by collinear omnidirectional antennas. So, the determination of a scattering matrix is different between these theories. In this paper, we model the directly coupled wave and reflected waves in polarimetric single-hole radar measurements with electric and magnetic dipoles arranged on the same axis. This model is used to extend the borehole radar polarimetry theory for an omnidirectional radar system. We start with formulation of polarimetric channel signals of the direct wave in cross-hole arrangements and directly coupled wave in single-hole arrangements. We also formulate polarimetric channel signals of a reflected wave from a plane scatter in single-hole arrangements. An equivalent scattering matrices of the direct coupling and the plane fracture are derived in same radar coordinate system for the polarimetric borehole radar. Method to determine the scattering matrix from the targets with single-hole measurements only is proposed and discussed. Finally, a simple polarimetric approach is applied to experimental data for reduction of the directly coupled wave and for enhancement of the reflected waves.

II. Formulation of Antenna Transfer Function

A. Direct Wave in Cross-Hole Arrangement

We successfully achieved polarimetric measurements over a wavelength 1–2 m in a borehole with the diameter of about 10 cm [7]. Dipole antennas and axial slot antennas on a conducting cylinder were used in order to measure both orthogonal polarizations. We had experimentally observed that the dipole and the slot antenna operate in the borehole as an electric and a magnetic dipole antenna with the different dipole length, respectively. Therefore, we can model the polarimetric borehole radar system as a radar system with a combination of electric and/or magnetic dipoles oriented to the same direction.

We start with deriving the relationship of effective length between transmitter and receiver in electric and magnetic dipoles. Fig. 1 shows cross-hole measurement model with electric dipoles. We define $\hat{\boldsymbol{z}}$ and $\hat{\boldsymbol{r}}$ as the unit vector oriented to the dipole axis and the unit propagation vector, respectively. A z-directed small dipole having the length of l and I current produces the electric field as

$$\boldsymbol{E}_\theta = \frac{j\omega\mu I l e^{-jkr}}{4\pi r}\left(1+\frac{1}{jkr}-\frac{1}{k^2r^2}\right)\sin\theta\hat{\boldsymbol{\theta}}. \qquad (1)$$

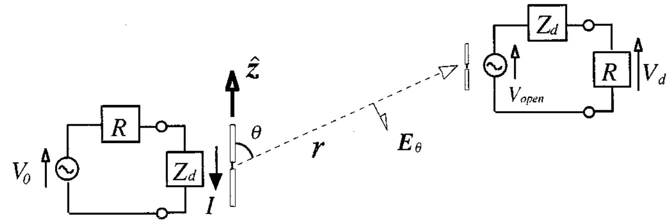

Fig. 1. Cross-hole arrangement model with electric dipoles and the equivalent circuits for a transmitter and a receiver.

Here, ω, μ, k, and θ are radian frequency, the permeability, the wave number, and the angle between $\hat{\boldsymbol{r}}$ and $\hat{\boldsymbol{z}}$, respectively. The effective length $\boldsymbol{l}_e(\hat{\boldsymbol{r}})$ of the small dipole is given as follows [12]:

$$\boldsymbol{l}_e(\hat{\boldsymbol{r}}) = l\sin\theta\hat{\boldsymbol{\theta}} = l\hat{\boldsymbol{r}}\times(\hat{\boldsymbol{z}}\times\hat{\boldsymbol{r}}). \qquad (2)$$

Here, we apply voltage V_0 to the transmitting electric dipole with the input impedance Z_d through the intrinsic impedance R, and define V_d to be the output voltage of a receiving electric dipole antenna at r away from the transmitter. The vector effective length $\boldsymbol{l}_{et}$ and $\boldsymbol{l}_{er}$ of the electric dipole defined for the transmitter and the receiver are given as follows:

$$\boldsymbol{l}_{et} = -\boldsymbol{l}_e(\hat{\boldsymbol{r}}) \quad \text{and} \quad \boldsymbol{l}_{er} = \boldsymbol{l}_e(-\hat{\boldsymbol{r}}). \qquad (3)$$

A ratio f_{dd} of the output voltage V_d to the exciting voltage V_0, which is the far-field antenna transfer function, is defined by using the vector effective length for both the transmitter and the receiver as follows [13]:

$$f_{dd} = \frac{V_d}{V_0} = j\omega\chi(r)\frac{\boldsymbol{l}_{et}}{R+Z_d}\cdot\frac{\boldsymbol{l}_{er}}{R+Z_d} = -j\omega\chi(r)\left(\frac{l\sin\theta}{R+Z_d}\right)^2 \qquad (4)$$

where $\chi(r)$ denotes the constant independent of the antenna characteristics and is defined as follows:

$$\chi(r) = \frac{\mu R e^{-jkr}}{4\pi r}. \qquad (5)$$

Next, we consider a magnetic dipole. Fig. 2 shows a cross-hole measurement model with the magnetic dipoles oriented to $\hat{z}$. When a small loop antenna having an area of S and a loop current I_t, the magnetic dipole produces a far-field magnetic component $\boldsymbol{H}_\theta$ as follows [12]:

$$\boldsymbol{H}_\theta = \frac{j\omega\varepsilon e^{-jkr}}{4\pi r}j\omega\mu S I_t\hat{\boldsymbol{r}}\times(\hat{z}\times\hat{\boldsymbol{r}}) = \frac{j\omega\varepsilon e^{-jkr}}{4\pi r}I_m\boldsymbol{l}_m(\hat{\boldsymbol{r}}) \qquad (6)$$

where ε denotes the permittivity of the medium. Here, we can arbitrarily select the effective length $\boldsymbol{l}_m(\hat{\boldsymbol{r}})$ of a magnetic dipole and its equivalent magnetic current I_m, which satisfy (6). Then, if we assume the schalar value of the magnetic current I_m to be the feeding voltage V_t at the transmitter, we obtain the effective length of the loop antenna with respect to the magnetic field as follows:

$$\boldsymbol{l}_m(\hat{\boldsymbol{r}}) = \frac{j\omega\mu S}{Z_s}\hat{\boldsymbol{r}}\times(\hat{z}\times\hat{\boldsymbol{r}}) \qquad (7)$$

where Z_s is the input impedance of the small loop antenna. The transmitting and receiving vector effective length of the mag-

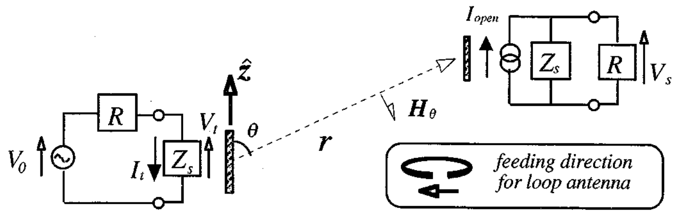

Fig. 2. Cross-hole arrangement model with magnetic dipoles generated by small loop antennas and the equivalent circuits for a transmitter and a receiver.

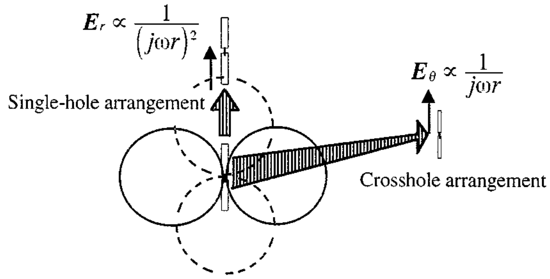

Fig. 3. Conceptual figure for a mechanism of direct-wave coupling in dipole–dipole combinations for single-hole and cross-hole arrangements. The solid and broken circles means the directivity of the far-field E_θ component and near field E_r component, respectively.

netic dipole are given in the feeding manner shown in Fig. 2 as follows:

$$\boldsymbol{l}_{mt} = \boldsymbol{l}_{mr} = \boldsymbol{l}_m(\hat{\boldsymbol{r}}) \tag{8}$$

respectively. From the equivalent circuit of the magnetic dipole model shown in Fig. 2, we obtain the far-field antenna transfer function f_{ss}, which is the ratio of the output voltage V_s to the exciting voltage V_0 as

$$\begin{aligned} f_{ss} = \frac{V_s}{V_0} &= j\omega \frac{\varepsilon e^{-jkr}}{4\pi r} \frac{Z_s \boldsymbol{l}_{mt}}{R+Z_s} \cdot \frac{RZ_s \boldsymbol{l}_{mr}}{R+Z_s} \\ &= j\omega\chi(r) \left(\frac{j\omega\mu S \sin\theta}{\eta(R+Z_s)}\right)^2 \end{aligned} \tag{9}$$

where η denotes the characteristic impedance of the medium. In the magnetic dipole case, the polarity of the vector effective length of the transmitter is the same as that of the receiver. It should be noted that the transmitting and the receiving electric dipole have an opposite polarity of the dipole moment to each other, however, the transmitting and receiving magnetic dipole have the same polarity of the magnetic moment as each other.

B. Directly Coupled Wave in Single-Hole Arrangement

Here, we discuss direct-wave coupling in a single-hole arrangement with small electric and magnetic dipoles. The mechanism of the directly coupled wave has already been analyzed [8]. In the dipole–dipole combinations, the radial component of the electric field $\boldsymbol{E}_r$ radiated from a small electric dipole is dominantly received at the collinear receiving dipole as shown in Fig. 3. The transmitting and receiving vector effective length $\boldsymbol{l}_{e0t}$ and $\boldsymbol{l}_{e0r}$ with respect to the axially oriented $\boldsymbol{E}_r$ component are given by substituting $\hat{\boldsymbol{r}} = \hat{z}$ as follows:

$$\boldsymbol{l}_{e0t} = -l(\hat{z}\cdot\hat{\boldsymbol{r}})\hat{\boldsymbol{r}} = -l\hat{z} \quad \text{and} \quad \boldsymbol{l}_{e0r} = l(-\hat{\boldsymbol{r}}\cdot\hat{z})(-\hat{\boldsymbol{r}}) = l\hat{z} \tag{10}$$

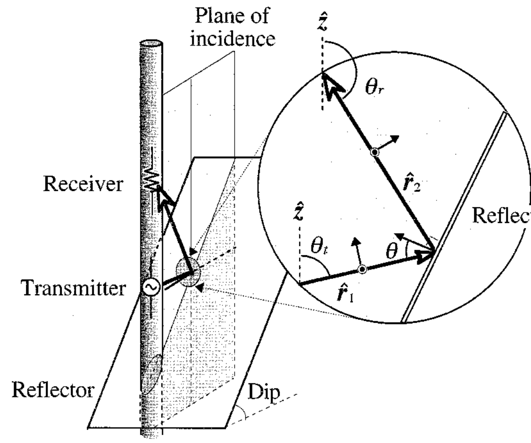

Fig. 4. Single-hole arrangement model in reflection from a plane reflector.

respectively. In the magnetic dipole case, we can similarly discuss with the duality of the electromagnetic field. Thus, the axially oriented H_r component is also dominant in the received signals. The transmitting and receiving vector effective length $\boldsymbol{l}_{m0t}$ and $\boldsymbol{l}_{m0r}$ with respect to the H_r component are given as follows:

$$\boldsymbol{l}_{m0t} = \boldsymbol{l}_{m0r} = \frac{j\omega\mu S}{Z_s}\hat{z}. \tag{11}$$

From (10) and (11), we obtain the transfer functions f_{dd0} and f_{ss0} with respect to the directly coupled wave in dipole–dipole and slot–slot antenna combinations as follows:

$$f_{dd0} = \frac{2\chi(h)}{h} \frac{\boldsymbol{l}_{e0t}}{R+Z_d} \cdot \frac{\boldsymbol{l}_{e0r}}{R+Z_d} = -\frac{2\chi(h)}{h}\left(\frac{l}{R+Z_d}\right)^2 \tag{12}$$

and

$$f_{ss0} = \frac{2\chi(h)}{h\eta^2} \frac{Z_s \boldsymbol{l}_{m0t}}{R+Z_s} \cdot \frac{RZ_s \boldsymbol{l}_{m0r}}{R+Z_s} = \frac{2\chi(h)}{h}\left(\frac{j\omega\mu S}{\eta(R+Z_s)}\right)^2. \tag{13}$$

These results imply that we can approximate far-field antenna transfer functions by taking derivative of directly coupled waves in single-hole measurements. We treat the direct coupling wave with the far-field approach in this formulation so this approach is not always applicable in the actual situation.

C. Scattering Matrix for Polarimetric Borehole Radar

Here, we consider the single-hole reflection model shown in Fig. 4. First, we define a vector effective length $\boldsymbol{l}_{el}(\hat{\boldsymbol{r}})$ of the loop antennas with respect to the electric field in order to discuss the scattering matrix repersentation with vector effective length of the electric field. The electric field $\boldsymbol{E}_\varphi$ radiated by the small loop in (6) is given by

$$\boldsymbol{E}_\varphi = -j\omega\chi(r)\frac{j\omega\mu S}{\eta} I_t(\hat{z}\times\hat{\boldsymbol{r}}) = j\omega\chi(r)\frac{V_0}{R+Z_s}\boldsymbol{l}_{el}(r). \tag{14}$$

The transmitting and receiving vector effective length $\boldsymbol{l}_{el\,t}$ and $\boldsymbol{l}_{el\,r}$ are defined in this feeding manner as

$$\boldsymbol{l}_{el\,t} = -\boldsymbol{l}_{el}(\hat{\boldsymbol{r}}) = \frac{j\omega\mu S}{\eta}(\hat{z}\times\hat{\boldsymbol{r}})$$

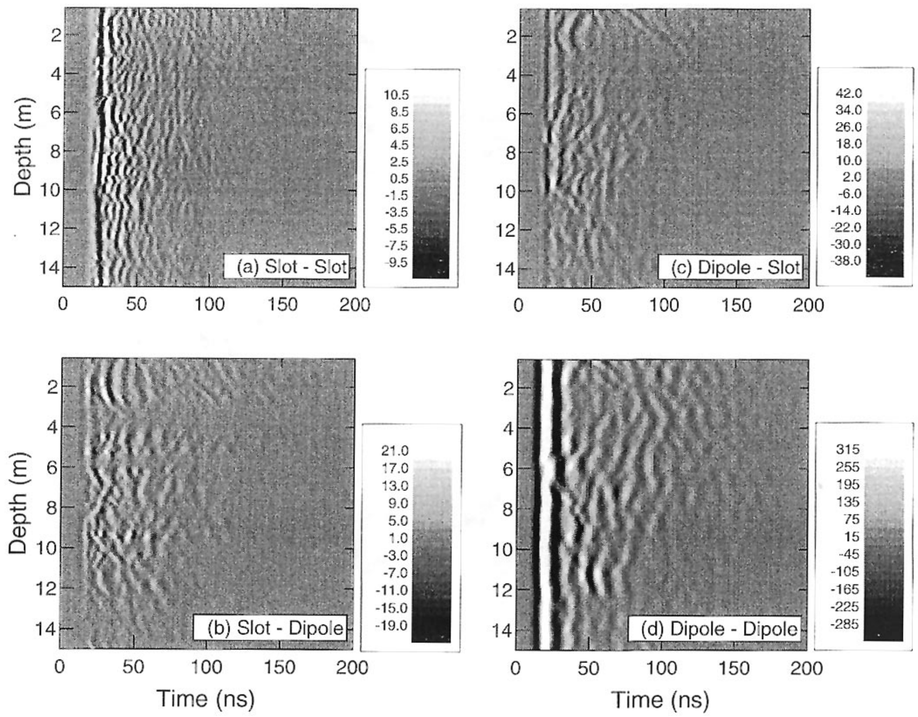

Fig. 5. Measured radar profiles in the polarimetric single-hole radar measurements with the antenna separation of 1.6 m, assuming a Gaussian pulse waveform with the bandwidth from 30 to 100 MHz.

and

$$\boldsymbol{l}_{el\,r} = \boldsymbol{l}_{el}(-\hat{\boldsymbol{r}}) = -\frac{j\omega\mu S}{\eta}\left(\hat{z}\times(-\hat{\boldsymbol{r}})\right) = \boldsymbol{l}_{el\,t} \tag{15}$$

respectively. If we assume a flat plane reflector far away from the borehole, the plane of incidence is uniquely defined as the plane including both the borehole axis $\hat{z}$ and the normal vector to the plane reflector. The reflected polarimetric signals $\boldsymbol{F}_p$ are given as a matrix representation with an antenna transfer matrix $\boldsymbol{G}$ and a target-dependent matrix $\boldsymbol{S}_p$ from a plane reflector by

$$\boldsymbol{F}_p = \chi(r_1 + r_2)\boldsymbol{G}\boldsymbol{S}_p\boldsymbol{G}. \tag{16}$$

Here,

$$\boldsymbol{G} = \pm\sqrt{j\omega}\begin{pmatrix} \dfrac{j\omega\mu S}{\eta(R+Z_s)} & 0 \\ 0 & \dfrac{l}{R+Z_d} \end{pmatrix} \tag{17}$$

and

$$\boldsymbol{S}_p = \sin\theta_r \sin\theta_t \begin{pmatrix} \Gamma_{\mathrm{TE}}(\theta) & 0 \\ 0 & -\Gamma_{\mathrm{TM}}(\theta) \end{pmatrix} \tag{18}$$

where θ, $\Gamma_{\mathrm{TE}}(\theta)$, $\Gamma_{\mathrm{TM}}(\theta)$, θ_t and θ_r denote the angle of incidence, the reflection coefficient of TE and TM wave from the plane reflectors, and the angle between $\hat{\boldsymbol{r}}_1$, $\hat{\boldsymbol{r}}_2$, and $\hat{z}$, shown in Fig. 4, respectively. This representation corresponds to a definition of a scattering matrix in the polarimetry theory [10]. The assumed feeding manner and $\boldsymbol{G}$ are important to determine the scattering matrix in the radar coordinate system.

A transmitting matrix $\boldsymbol{T}_0$ of the direct coupling between the transmitter and the receiver is similarly given by the matrix representation from (12) and (13) as follows:

$$\boldsymbol{F}_0 = \begin{pmatrix} f_{ss0} & 0 \\ 0 & f_{dd0} \end{pmatrix} = \frac{2}{j\omega h}\chi(h)\boldsymbol{G}\boldsymbol{T}_0\boldsymbol{G}. \tag{19}$$

Here,

$$\boldsymbol{T}_0 = \begin{pmatrix} 1 & 0 \\ 0 & -1 \end{pmatrix}. \tag{20}$$

The difference of the matrix components between $\boldsymbol{S}_p$ and $\boldsymbol{T}_0$ is caused by the change of the propagating direction between reflection and transmission.

III. Polarimetric Reflection Enhancement in Single-Hole Measurement

Using the formulation obtained in the previous section, we propose a method to enhance reflected waves against the directly coupled wave, which is important in practical measurements. The measured signal $\overline{\boldsymbol{F}}_m$ in polarimetric single-hole radar measurement is expressed by using the *in situ* antenna transfer matrix $\overline{\boldsymbol{G}}$ by

$$\begin{aligned}\overline{\boldsymbol{F}}_m &= \overline{\boldsymbol{F}}_0 + \overline{\boldsymbol{F}}_{p,m} \\ &= \overline{\boldsymbol{G}}\left(\frac{2\chi(h)}{j\omega h}\boldsymbol{T}_0 + \chi(r_{1,m} + r_{2,m})\boldsymbol{S}_{p,m}\right)\overline{\boldsymbol{G}}\end{aligned} \tag{21}$$

where the bar denotes the measured value and the subscript m denotes to the measuring depth. Here, the first term denotes the direct coupling and the second term denotes the reflected signal. Generally, the first term is much larger than the second term in (21). In addition, the second term is a function of the reflected

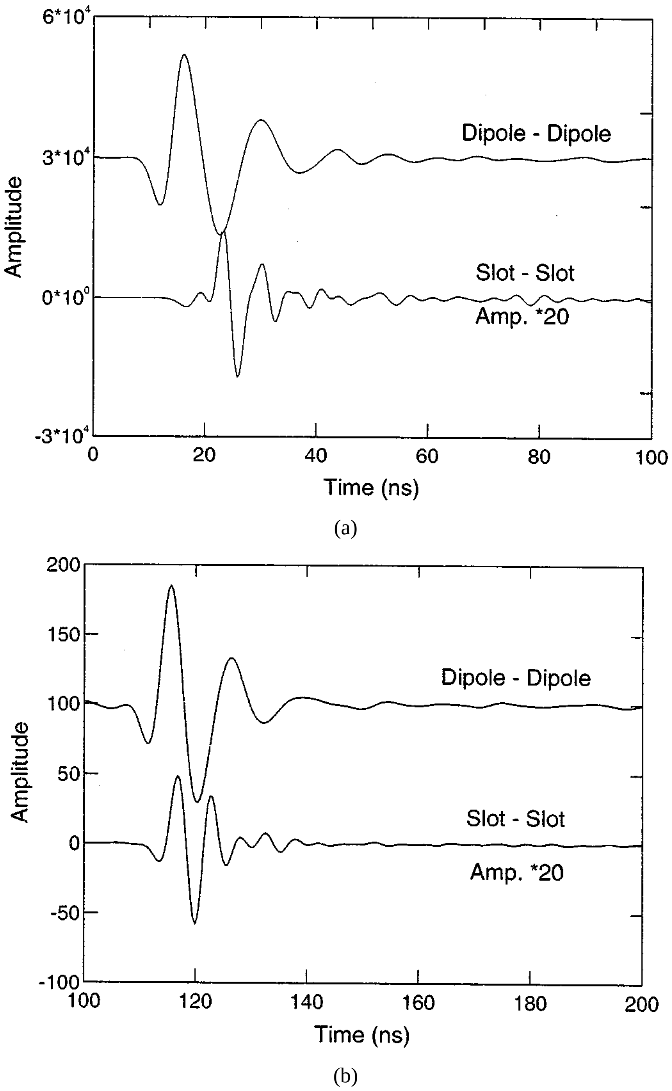

Fig. 6. Comparison of the time derivative of the averaged single-hole waveforms to cross-hole waveforms obtained with separation of 11 m in the same field for both polarizations. (a) Time-derivative waveform of the direct coupling. (b) Cross-hole waveform.

point, although the first term is a constant with respect to the measuring depth. Thus, averaging $\overline{\boldsymbol{F}}_m$ over the measured depths approximately yields the directly coupled wave in each antenna arrangement as

$$\overline{\boldsymbol{F}}_{ave} = \frac{1}{N}\sum_{m=1}^{N}\overline{\boldsymbol{F}}_m \cong \frac{2\chi(h)}{j\omega h}\overline{\boldsymbol{G}}\boldsymbol{T}_0\overline{\boldsymbol{G}}. \tag{22}$$

And $\overline{\boldsymbol{G}}$ can approximately be determined by using the averaged single-hole signals $\overline{F}_{ave,\,ss}$ and $\overline{F}_{ave,\,dd}$ in the slot–slot and dipole–dipole arrangement as

$$\overline{\boldsymbol{G}}^2 \cong \frac{j\omega h}{2\chi(h)}\begin{pmatrix}\overline{F}_{ave,\,ss} & 0\\ 0 & -\overline{F}_{ave,\,dd}\end{pmatrix}. \tag{23}$$

This approximation is theoretically valid only when the current distribution on the antennas is constant. However, it is justified by FDTD simulation and field experiments [8] when the ratio of the antenna separation to the dipole length is larger than 1.6 and when the frequency range is less than the resonant frequency.

Now, we can determine the scattering matrix $\overline{\boldsymbol{S}}_m$ by applying the $\overline{\boldsymbol{G}}^{-1}$ to the measured polarimetric single-hole signals as

$$\overline{\boldsymbol{S}}_m = \begin{pmatrix}S_{HH} & S_{HV}\\ S_{VH} & S_{VV}\end{pmatrix} = \overline{\boldsymbol{G}}^{-1}\overline{\boldsymbol{F}}_m\overline{\boldsymbol{G}}^{-1}. \tag{24}$$

Here, we consider the particular case to show the difference of the scattering matrix between the direct coupling wave and the reflected wave. When the target is a plane reflector far away from the borehole, the angle of incidence is almost 0°. The matrix component of the scattering matrix $\boldsymbol{S}_p$ should be an unit matrix. Then, we obtain a simple formula of the estimated scattering matrix by using (18) and (19)

$$\overline{\boldsymbol{S}}_m = \frac{2\chi(h)}{j\omega h}\begin{pmatrix}1 & 0\\ 0 & -1\end{pmatrix} + \chi(r_{1,\,m} + r_{2,\,m})\Gamma(0)\sin^2\theta_r\begin{pmatrix}1 & 0\\ 0 & 1\end{pmatrix}. \tag{25}$$

Equation (25) shows that adding S_{HH} and S_{VV} cancels the energy of the directly coupled wave and enhances the reflection from a plane reflector. The $S_{HH} + S_{VV}$ corresponds to the summation of the amplitude for the reflection coefficient of the TE and TM wave. Therefore, the polarization filtering of $S_{HH} + S_{VV}$ gives information of only the radar targets.

IV. Application to Field Data

A. Single-Hole Measurement

Field experiments were carried out in a field laboratory of Tohoku University, Kamaishi Mine, Iwate, Japan [6]. The host rock is granite and its relative dielectric constant and the conductivity are eight and less than 0.005 S/m, respectively. From the gallery, we can observe the fractures filled with water having aperture less than 0.1 cm and containing wet clay layer having aperture of about 3 cm. They can be main radar targets. A stepped frequency radar-system-based on a network analyzer (HP-8752A) was employed. The optical analog transmission link achieves the signal transmission between the network analyzer on the ground surface and the antennas in a borehole [7]. The antennas were set in a hollow fiberglass cylinder for insulation from water filling a borehole.

We conducted polarimetric single-hole measurement with the antenna separation of 1.6 m in the well KR-4. Fig. 5 shows the radar profiles obtaind with the inverse Fourie transformation of the measured frequency-domain data assuming Gaussian pulse waveform having a bandwidth from 30–100 MHz . We find that the detectable radar range is about 5 m in this particular field. Differences in amplitude and the waveform can be also observed for different antenna combinations. A conventional borehole radar can obtain only the radar profile of the dipole–dipole arrangement. The higher frequency is found to be dominant in the slot–slot antenna arrangement than that in the dipole–dipole arrangement. However, the directly coupled wave is relatively weak in the slot–slot arrangement. It is difficult to distinguish the reflections in the copolarization arrangements because a large directly coupled wave masks small reflected waves. While we can easily find the reflected waves in

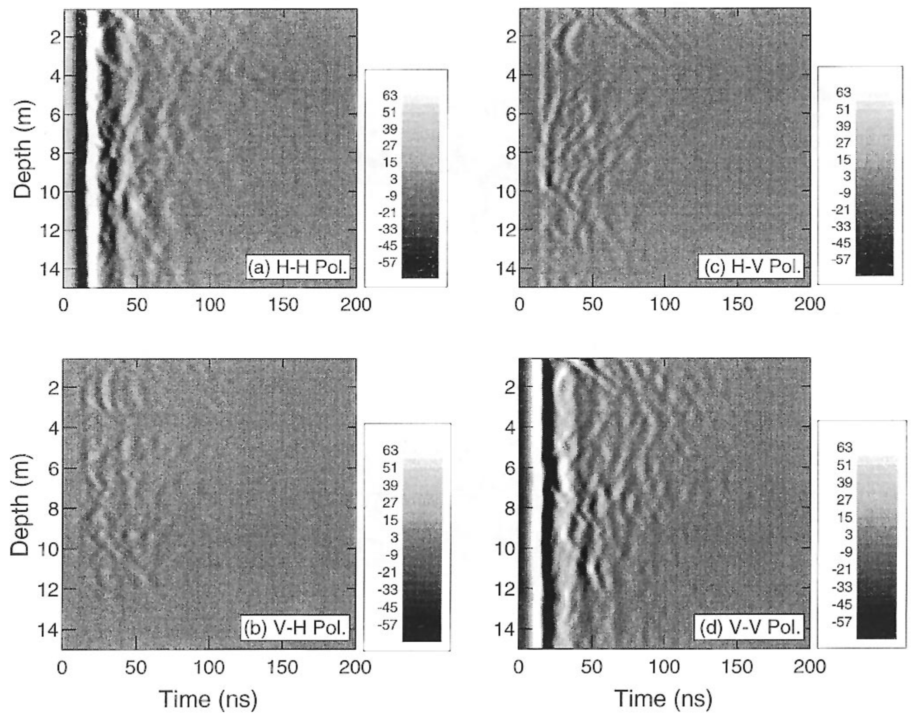

Fig. 7. Compensated radar profiles with the inverse filters $\overline{\boldsymbol{G}}^{-1}$ made with the time derivative of the averaged single-hole signals. The bandwidth of compensated signals is limited from 50 to 100 MHz.

the cross-polarization configurations due to better antenna isolation. However, the reflection from a reflector detected at 6 m in the dipole–dipole arrangement is relatively weak in the cross-polarization component because the cross-polarization components are sensitive to surface roughness of a fracture. If the fracture have smooth surface, depolarization cannot be observed theoretically.

B. Reflection Enhancement

The time derivative of the 150 averaged single-hole waveforms in both the antenna combinations are showed in Fig. 6. The cross-hole waveforms measured in the same field is also showed in the same figure. Both waveforms are similar in each antenna combination. Moreover, amplitude ratio of the cross-hole signal to the averaged single-hole signal is almost same between the dipole–dipole and slot–slot combinations. Fig. 7 shows radar profiles compensated with the inverse filters $\overline{\boldsymbol{G}}^{-1}$ of the time derivative of $\overline{F}_{ave,ss}$ and $\overline{F}_{ave,dd}$, which were shown in (23). We observe that these filters compensate the difference of the amplitude and frequency characteristics caused by the difference of the antenna transfer functions. However, a pulse compression effect is not so significant because the compensated signals are limited by a bandpass filter having bandwidth between 50 and 100 MHz, where the averaged single-hole signal can be identified as a signal directly coupled by the near field. The amplitude of the directly coupled wave in the cross-polarization components is much smaller than that in the copolarization components. The theoretical analysis of the measured transmission matrix $\boldsymbol{T}_0$ shown in (20) is experimentally validated.

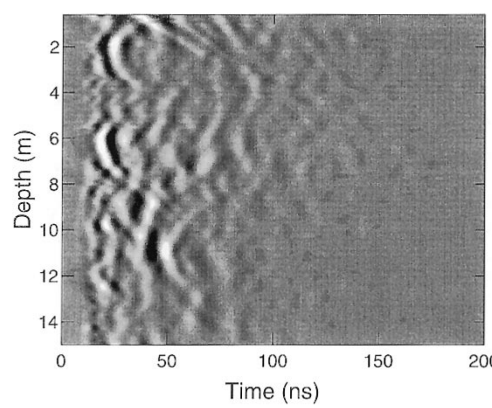

Fig. 8. Result of elimination of the directly coupled wave from the compensated radar profile by adding HH and VV components in Fig. 7.

Fig. 8 shows $S_{HH} + S_{VV}$ component in the compensated polarimetric radar profiles. We find that the directly coupled waves could be completely removed on the radar profile and can clearly recognize the reflected waves near the borehole, which we could not ever have obtained. Reflections from plane reflectors are observed across the borehole at 2 and 6 m. The significant large reflected waves less than 30 ns are influenced by the near-field radiation at those depths. The reflection by the near field is also enhanced because the relationship of the polarization characteristics between $\boldsymbol{T}_0$ and $\boldsymbol{S}_p$ is maintained even in the near-field radiation. We can observe the difference of the polarity of the reflection coefficient between the reflectors at 2 and 6 m. The existence of clutter in a lossy medium often makes it difficult to distinguish the phase information of reflected waves in conventional borehole radar measurements because the detectable distance shorten to less than several times of the used

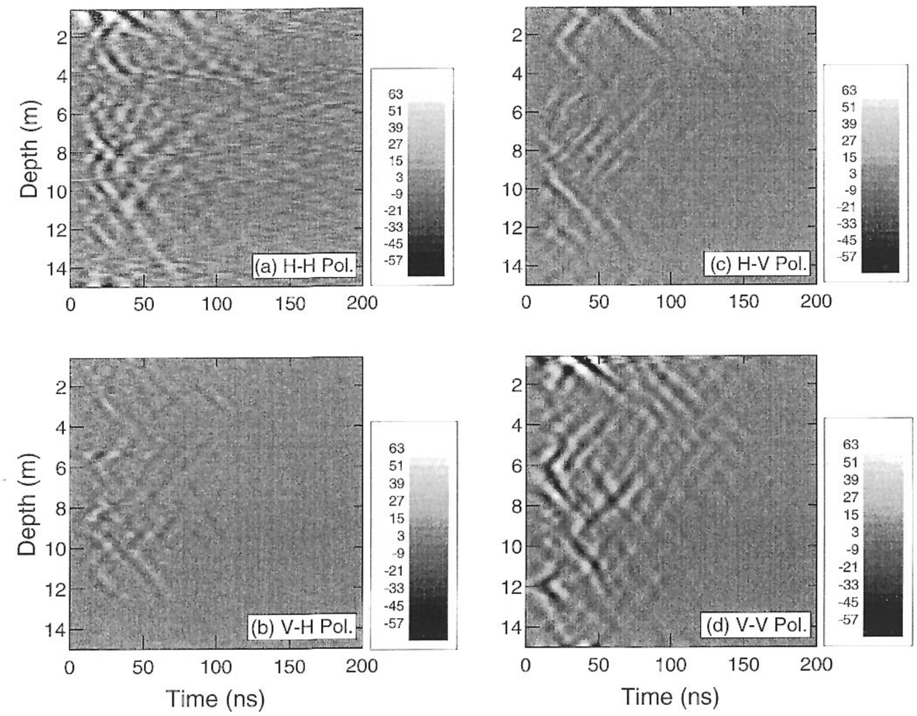

Fig. 9. Result of elimination of the directly coupled wave from the radar profile by subtraction of the averaged waveforms of the 11 adjacent traces from the waveform at each center depth as a conventional technique.

spatial pulse width in such the lossy medium. Fig. 9 show the radar profile subtracted the averaged waveforms of the 11 adjacent traces from the waveform at each center depth. This is a conventional signal processing commonly used for the enhancement of the reflections. These figures also give the information of the existence of the fractures. However, the phase information is found to be poor and the amplitude of the reflected wave is essentially reduced by this subtraction procedure. Thus, $S_{HH}+S_{VV}$ give more accurate information of reflected waves near the borehole. It is very important for further characterization techniques of reflectors to be able to obtain the amplitude and phase information.

In addition, a parabolic reflection profile from a point target can be observed at 11 m around 45 ns in Fig. 8. On the other hand, the procedure of the subtraction of the averaged waveforms reduces the parabolic reflection profile reflected from a point target far away from borehole as shown in Fig. 9(d). A reflection from a point target has the same scattering matrix as a plane reflector and is also enhanced by summation between the copolarization components. Therefore, this technique is effective for acquiring the target information not only near the borehole but also away from the borehole. In addition, the technique of the subtraction of the averaged waveform is effective in only relatively homogeneous medium around the borehole. However, the presented technique is useful independently of the homogeneity of the medium because the direct coupling wave is suppressed by use of the polarization characteristics.

V. Conclusion

In this paper, polarimetric approach has been used for reduction of a directly coupled wave and enhancement of a reflected wave in borehole radar measurement. The received signals in polarimetric borehole radar were analyzed, assuming usage of electric and magnetic dipoles as a transmitter and a receiver. The antenna transfer functions and the directly coupled wave are quantitatively represented with the effective length of electric dipoles and magnetic dipoles. The general representation of a scattering matrix from plane reflectors was derived in polarimetric borehole radar measurement. In single-hole measurements, it was theoretically shown that the transmission matrix for the directly coupled wave have the feature of the scattering matrix from a dihedral corner reflector as shown in (20) although the scattering matrix from a plane reflector is almost the unit matrix.

Theoretical analysis in single-hole measurement yields a technique to effectively remove the directly coupled wave and to enhance reflections, which was masked by the directly coupled wave, from plane reflectors. This technique is composed of two phases. In the first phase, to estimate the scattering matrix, the antenna transfer functions are compensated with the time derivative of the directly coupled signals determined by averaging all the received signals. In the second phase, taking the summation of the VV and HH components of the estimated scattering matrix gives the information of reflected waves.

We have demonstrated the validity of this method with polarimetric single-hole measurement data obtained in granite rock with dipole antennas and slot antennas on a conducting cylinder. The directly coupled signals in the both dipole–dipole and slot–slot antenna combinations were determined for a bandwidth between 50–100 MHz by averaging the 150 raw signals. The *in situ* antenna transfer functions $\overline{\boldsymbol{G}}$ were compensated with the time derivative of the estimated directly coupled signals so that a full-polarimetric borehole radar profile was successfully estimated with only single-hole measurement data. The $S_{HH}+S_{VV}$ procedure completely reduces the directly coupled waves and enhances the reflections from the plane reflectors. We conclude that the proposed antenna compensation technique and polarization filtering technique of $S_{HH}+S_{VV}$ are effective to obtain the information of the reflected waves without reduction of the amplitude of the reflection from the target near the borehole. We think that these techniques are also more useful for obtaining the phase information of the reflected wave in lossy or inhomogeneous medium than the conventional signal processing.

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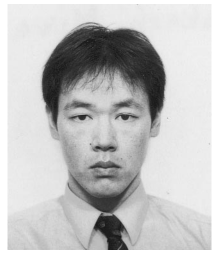

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