0.05–3 GHz VNA characterization of soil dielectric properties based on the multiline TRL calibration

Arkadiusz Lewandowski¹, Agnieszka Szypłowska², Marcin Kafarski², Andrzej Wilczek², Paweł Barmuta¹ and Wojciech Skierucha²

¹ Institute of Electronic Systems, Warsaw University of Technology, Nowowiejska 15/19, 00-665 Warsaw, Poland
² Institute of Agrophysics, Polish Academy of Sciences, Doświadczalna 4, 20-290 Lublin, Poland

E-mail: a.lewandowski@elka.pw.edu.pl

Received 20 July 2016, revised 12 October 2016
Accepted for publication 26 October 2016
Published 12 January 2017

Abstract
We present a methodology for characterization of soil relative dielectric permittivity in the frequency range 0.05–3 GHz. Soil samples are placed in a measurement cell constructed out of a EIA 1-5/8” coaxial transmission line, and then measured with a calibrated vector-network-analyzer. From these measurements the relative dielectric permittivity is obtained by use of a modified Boughriet algorithm. In order to calibrate the vector-network-analyzer directly at the EIA 1-5/8” coaxial-transmission-line measurement planes, we use the multiline through-reflect-line method. This method, while providing superior vector-network-analyzer calibration accuracy, is also easy to implement since it uses only transmission lines with known lengths and a single unknown highly-reflective termination. The implemented calibration method was compared to a simplified approach that uses the standard SOLT calibration in Type-N reference planes, and then accounts for the Type-N/EIA 1-5/8” adapters by removing their electrical delay. Experimental results for teflon and soil samples with different moisture content and salinity confirmed the validity of our approach.

Keywords: soil relative dielectric permittivity, vector network analyzer, multiline through-reflect-line calibration

(Some figures may appear in colour only in the online journal)
Open-ended probes can be used to measure dielectric properties of fine-grained soil [10]. For the purpose of soil complex dielectric permittivity determination in a broad frequency range, coaxial transmission line cells are used [11–14] which allow to measure larger sample-volumes than the open-ended probes. Reference [14] introduces a frequency-domain approach based on wideband scattering-parameter measurements of soil samples inserted into a measurement cell constructed out of a EIA 1-5/8” coaxial transmission line. This cell is connected to a vector-network-analyzer (VNA) and from the measurement of its scattering parameters, the dielectric spectrum of the soil sample is determined. The use of a coaxial transmission lines with a large diameter allows to measure large volumes of soil which improves the measurement accuracy. However, this also introduces significant calibration problems, as there are no commercially available calibration standards in the EIA 1-5/8” coaxial transmission line. Even when using such standards, it would be difficult to shift the measurement plane from the connection plane down to the actual plane at which the soil sample begins. Thus, in [14] an approximate approach is used in which the calibration is performed with Type-N calibration standards and the cell is connected through Type-N/EIA 1-5/8” adapters. The impact of these adapters is then approximately accounted for by removing only their electrical delay.

In this work, we present a wideband scattering-parameter-based approach for the measurement of soil dielectric spectrum in which the scattering-parameters of the soil sample are determined at the reference planes at which the sample is actually inserted. To this end, a custom-designed set of calibration standards is used to implement the multiline through-reflect-line (TRL) calibration [15]. This set consists of five EIA 1-5/8” coaxial transmission lines with different lengths designed based on the methodology [16]. By defining the calibration standards in the EIA 1-5/8” coaxial-transmission-line plane, we can accurately calibrate out the Type-N/EIA 1-5/8” adapter, as opposed to the approximate approach of calibrating at the Type-N connector planes, and thus shift the measurement plane after the calibration down to the very soil sample.

2. Measurement setup

Schematic of the measurement setup is shown in figure 1(a). We performed scattering parameter measurements with the Anritsu MS4642A VNA in the frequency range 0.05 to 3 GHz. The measurement planes were set at the EIA 1-5/8” connectors which were attached to each VNA port through a cascade of a Type-N/EIA 1-5/8” adapter, a Type-N precision microwave cable, and a Type-N/3.5 mm adapter. Soil samples were inserted into a measurement cell constructed out of a 60 mm long coaxial transmission line made out of copper and then connected to measurement system through additional EIA 1-5/8” flanges. The cross-sectional dimensions of the line are specified by the EIA RS-225 standard: the inner diameter of the outer conductor is 38.8 mm \pm 0.075 mm while the outer diameter of the inner conductor is 16.9 mm \pm 0.050 mm. The diameters of the inner and outer conductors were specifically chosen in order to measure large sample volume, which is important for inhomogeneous materials, such as soil. These dimensions were analogous to the setup used for soil samples measurements in [14].

Prior to the characterization of the measurement cell with soil samples, the VNA was calibrated with the multiline TRL method [15] which is commonly considered as the most accurate technique for the vector-network-analyzer (VNA) calibration. This method uses a set of transmission lines with different lengths and the same but otherwise unknown propagation constant, a reflect standard which is assumed to be identical on both VNA ports but otherwise unknown, and a thru connection. All of the unknown parameters of the calibration standards, that is, the propagation constant of the lines and the reflection coefficient of the reflect standard are then determined along with the VNA calibration coefficients [15].

In order to implement the multiline TRL method in our measurement environment, we designed a set of five EIA 1-5/8” coaxial transmission lines by use of the method [16]. Figure 2 shows a picture of this set of five EIA coaxial transmission lines by use of the method [16]. Each line has the length of 500 mm while the shortest one is 60 mm. The method [16] allows to choose the line lengths so as to obtain an accurate calibration in a given measurement bandwidth (which is 0.05 to 3 GHz in our case). In order to obtain an accurate calibration at lower frequencies, the longest line should be in general made as long as possible. We limited the length of this line to 500 mm long for practical reasons, since connecting a longer line would be very difficult.

The calibration lines were manufactured out of copper and have the same diameters as the measurement cell. They were connected to the VNA measurement ports through additional EIA 1-5/8” flanges. As a reflect standard [15], we used an open circuit constructed by connecting to the EIA 1-5/8” flange a shielded section of the outer conductor. We implemented the multiline TRL calibration procedure [15] in the MATLAB environment [17]. In the VNA calibration and error correction we used the model proposed in [18].

It is important to note that the use of calibration standards implemented at the EIA 1-5/8” measurement plane has two significant advantages. First, by defining the calibration standards in this plane, we can accurately calibrate out the Type-N/EIA 1-5/8” adapter. Secondly, the multiline TRL calibration allows us to calibrate out also the discontinuity occurring at the connection plane (due to the center-conductor support element). Indeed, since the transmission lines are defined as reflection-less (with respect to their characteristic impedance) and the connection-plane discontinuity is identical for all connected elements, its electrical parameters are lumped into the VNA calibration coefficients. Thus, we can obtain scattering-parameter measurement of the measurement cell itself.

3. Modeling and extraction

We model the scattering-parameter measurement of the cell based on figure 2. The measurement cell can be described as a cascade connection of transmission lines with different dielectrics: sections A1 and A2 correspond to the short air
gaps, sections \( T_1 \) and \( T_2 \) describe the teflon beads supporting
the sample and the center conductor, while section \( M \) corre-
sponds to the line filled with the material under test (MUT).

In order to model the scattering parameters of the cell, we
use the transmission-matrix description. For a two-port net-
work, the scattering matrix \( S \) and transmission matrix \( T \) are
defined as \cite{19}:

\[
\begin{bmatrix}
 b_1 \\
 b_2 \\
 a_1 \\
 a_2 
\end{bmatrix} = \begin{bmatrix}
 S_{11} & S_{12} \\
 S_{21} & S_{22} \\
\end{bmatrix}
\begin{bmatrix}
 a_1 \\
 a_2 
\end{bmatrix}, \quad \text{and} \quad
\begin{bmatrix}
 b_1 \\
 b_2 
\end{bmatrix} = \begin{bmatrix}
 T_{11} & T_{12} \\
 T_{21} & T_{22} \\
\end{bmatrix}
\begin{bmatrix}
 a_1 \\
 a_2 
\end{bmatrix}.
\]

(1)

where

\[
S = \begin{bmatrix}
 S_{11} & S_{12} \\
 S_{21} & S_{22} 
\end{bmatrix}, \quad \text{and} \quad T = \begin{bmatrix}
 1 & \det S & S_{11} \\
 S_{21} & -S_{22} & 1
\end{bmatrix}.
\]

(2)

It can be easily shown that for a cascade connection of two-
port networks with transmission matrices \( T_i \), for \( i = 1, \ldots, N \),
referred to the same impedance \( Z_{\text{ref}} \), the overall transmis-
sion matrix is given by a product \( T_N \cdots T_1 \) \cite{19}. Thus, for
the model given in figure 2, we can write the measured trans-
mission matrix as:

\[
T_{\text{cell}} = T_A T_T T_M T_T T_A^{-1}.
\]

(3)

Since we know the lengths of the the teflon and air sec-
tions, we can determine their transmission matrices \( T_A, T_A^{-1},
T_T, T_T, T_T, \) and \( T_T, \) and from the measured transmission matrix
of the cell determine the transmission parameters of the sample
itself as:

\[
T_M = (T_A T_T T_T T_T T_A^{-1})^{-1} T_{\text{cell}} (T_T T_T T_T T_A^{-1})^{-1}.
\]

(4)

In order to evaluate (4), we need to write transmission
parameters of each transmission-line section in figure 2 with
respect to the same reference impedance \( Z_{\text{ref}} \). Expressing first
these parameters with respect to the line characteristic imped-
ance \( Z_0 \) (which is different for each section) we obtain \cite{20}:

\[
T_0 = \begin{bmatrix}
 e^{-j\gamma l} & 0 \\
 0 & e^{j\gamma l}
\end{bmatrix}
\]

(5)

where \( \gamma \) is the complex propagation constant and \( l \) is the
length of the line section. Assuming a non-magnetic dielectric with
a relative dielectric permittivity \( \varepsilon_r \) and neglecting conductor
loss \cite{3}, we can write the propagation constant \( \gamma = j\omega \sqrt{\varepsilon_r} / c \).

\footnote{We assume that the conductor loss in the coaxial-line walls is much smaller
than the dielectric loss of the soil sample.}
and the characteristic impedance $Z_0 = \eta/\sqrt{\varepsilon_f}$, where $c$ is the speed of light in vacuum and $\eta$ is the characteristic impedance of the coaxial transmission line with vacuum-dielectric\(^4\). By comparing (5) with (2), we note that $S_{11} = S_{22} = 0$, which is a direct consequence of the fact that a line is impedance-matched to its characteristic impedance $Z_0$. Rewriting (5) with respect to an arbitrary reference impedance $Z_{\text{ref}}$ we obtain [20]:

$$T_j = \frac{e^{i\theta_j}}{1 - \Gamma_j^2} \begin{pmatrix} e^{-2i\theta_j} - \Gamma_j^2 & (1 - e^{-2i\theta_j})\Gamma_j \\
1 - (1 - e^{-2i\theta_j})\Gamma_j & 1 - e^{-2i\theta_j}\Gamma_j^2 \end{pmatrix}$$

(6)

where $\Gamma_j = Z_0 - Z_{\text{ref}}/Z_0 + Z_{\text{ref}}$.

Reference impedance $Z_{\text{ref}}$ can be chosen arbitrarily. However, it is convenient to set $Z_{\text{ref}}$ to the characteristic impedance $Z_{\text{ch}} = \eta/\sqrt{\varepsilon_{\text{ch}}}$ of the air-filled sections, where $\varepsilon_{\text{ch}}$ is the dielectric permittivity of air. This is due to the fact that a VNA calibrated with the multiline TRL calibration measures scattering-parameters with respect to the characteristic impedance of the transmission lines used in the calibration [15, 20]. In this work we use air-filled coaxial transmission lines so the reference impedance of the multiline TRL calibration is precisely $Z_{\text{ch}}$.

Having determined the transmission matrix of the MUT from (4), we convert it to the scattering-parameter representation by inverting (2). We then apply the extraction algorithm described in appendix to obtain the relative dielectric permittivity of the MUT.

4. Materials

In order to verify our system, we measured a 20.6 mm thick teflon sample inserted into the measurement cell. After that, several soil samples of various water content and salinity were examined. All of the soil samples were prepared based on a sandy soil material (texture: sand 89.9%, silt 9.6%, clay 0.5%) by adding a predefined amount of moistening liquid and mixing. The particle-size distribution was measured by the laser diffraction method according to the procedure described in [21] and was presented in Table 1. The values of water content $\theta_M$ calculated on a dry mass basis, mass $m$, volume $V$, volumetric water content $\theta_V$, and bulk density $d$ were listed in Table 2.

| Sample | $\sigma_e$ (mS m\(^{-1}\)) | $\theta_M$ (g g\(^{-1}\)) | $m$ (g) | $V$ (cm\(^3\)) | $\theta_V$ (%) | $d$ (g cm\(^{-3}\)) |
|--------|-----------------|----------------|------|-------------|-------------|----------|
| A0     | 3.21            | 80.98         | 44.26| air-dry     | 1.83        |          |
| A1     | 3.21            | 77.60         | 45.62| 8.1         | 1.70        |          |
| A2     | 3.21            | 90.86         | 45.34| 18.2        | 2.00        |          |
| A3     | 3.21            | 96.16         | 45.31| 26.1        | 2.12        |          |
| B1     | 18.72           | 95.06         | 47.22| 18.3        | 2.01        |          |
| B2     | 36.38           | 93.70         | 47.37| 18.0        | 1.98        |          |
| B3     | 36.38           | 94.06         | 47.04| 18.2        | 2.00        |          |

Table 1. Particle-size distribution (PSD) of the soil material.

| Sample | $\sigma_e$ (mS m\(^{-1}\)) | $\theta_M$ (g g\(^{-1}\)) | $m$ (g) | $V$ (cm\(^3\)) | $\theta_V$ (%) | $d$ (g cm\(^{-3}\)) |
|--------|-----------------|----------------|------|-------------|-------------|----------|
| A0     | 3.21            | 80.98         | 44.26| air-dry     | 1.83        |          |
| A1     | 3.21            | 77.60         | 45.62| 8.1         | 1.70        |          |
| A2     | 3.21            | 90.86         | 45.34| 18.2        | 2.00        |          |
| A3     | 3.21            | 96.16         | 45.31| 26.1        | 2.12        |          |
| B1     | 18.72           | 95.06         | 47.22| 18.3        | 2.01        |          |
| B2     | 36.38           | 93.70         | 47.37| 18.0        | 1.98        |          |
| B3     | 36.38           | 94.06         | 47.04| 18.2        | 2.00        |          |

Table 2. Parameters of the soil samples.

Material with distilled water to achieve three levels of water content up to saturation. Samples with the designator B were obtained from the more saline soil material as A samples. They were wetted up to the same target moisture of $\theta_M = 0.10$ g g\(^{-1}\) (on a dry mass basis) with either distilled water (samples B1 and B2) or a KCl solution (sample B3) of electrical conductivity 2 S m\(^{-1}\).

After preparation, each soil sample was placed between the inner and outer conductor of the 60 mm long measurement cell, supported by two 4 mm thick teflon beads. The precise positions of the supporting beads for each tested sample were measured and accounted for in the permittivity extraction algorithm. The dielectric measurements of the samples were performed at room temperature $22 \pm 1 \, ^{\circ}\text{C}$.

5. Results and discussion

In figure 3 we present results for the teflon sample used as a reference material. These results were obtained with the use of the VNA error-correction procedure based on the multiline TRL calibration (see section 2) and compared with a simplified error correction method proposed in [14]. This simplified procedure uses a classical two-port short-open-load-thru (SOLT) calibration procedure to calibrate the VNA in the Type-N reference planes and then corrects for the impact of Type-N/EIA 1-5/8” adapters by removing only their electrical length.

We see that the real part of the dielectric permittivity of teflon sample obtained with our procedure agrees very well with the literature: in the frequency range 0.05–3 GHz we obtained flat spectrum with $\varepsilon' = 2.07 \pm 0.02$ which differs from $\varepsilon' = 2.035 \pm 0.005$ reported in [9] by around 1.7%. We also see that the imaginary part $\varepsilon''$ is close the noise floor of our measurement system.

We further see the the spectrum obtained with the simplified error-correction procedure gives for small frequencies values of $\varepsilon'$ by almost 10% smaller than expected for teflon and that it also overestimates the loss. This can be easily justified with the fact that by correcting only for the electrical length of the adapters, we lump their loss into the consecutive MUT measurements.

\(^4\)These expressions are valid only for TEM lines such as the coaxial transmission line discussed in this paper; for waveguides with non-zero cutoff frequency one would have to additionally include the frequency dependence of the group velocity.
In figure 4 we present dielectric spectra for non-saline soil samples. We see that the real part $\varepsilon'_{r}$ of the dielectric permittivity (see figure 4(a)) increased with the moisture content of the samples and varied from about 3 for the air-dry sample to a little over 16 for the saturated one. As expected for soil with a small clay content, the permittivity values did not decrease much with the increase of frequency.

Although all of the presented saline samples were wetted to achieve the same target moisture $\theta_{M} = 0.10 \text{ g g}^{-1}$, the real part of dielectric permittivity varied among the samples (see figure 5(a)). It was caused by slight differences in bulk density (see table 2). Also, evaporation during sample preparation could have influenced the final water content.

The imaginary part $\varepsilon''_{r}$ of dielectric permittivity of the air-dry sample was under 0.1 in the entire frequency range with
relatively small scatter (see figure 4(b)). The impact of bulk electrical conductivity on the imaginary part of dielectric permittivity was observed even for non-saline wet samples at frequencies below 0.5 GHz. For the saline-soil samples (see figure 5(b)), the imaginary part \( \varepsilon'' \) of the dielectric permittivity rose significantly at low frequencies. For the most saline B3 sample, electrical conductivity dominated the \( \varepsilon'' \) spectrum in the entire applied frequency range.

The dielectric permittivity spectra of the samples of various moisture content obtained from the same non-saline soil as in the present study was examined in the 0.1–1.2 GHz frequency range in [8]. The results obtained in the present work agreed with the previous findings.

6. Conclusions

In this work, we presented a technique for wideband 0.05–3 GHz characterization of soil dielectric spectra by use of vector-network-analyzer (VNA) scattering-parameter measurements. Our technique is based on VNA measurement of soil samples inserted into a EIA 1-5/8” coaxial-transmission-line cell. Before the measurements, the VNA is calibrated with the multinline through-reflect-line (TRL) method [15] with a custom-designed set of calibration standards based on the EIA 1-5/8” coaxial-transmission-line. The use of calibration standards defined in the same waveguiding structure as the measurement cell makes it possible to remove the systematic errors due to the cell connectors, and thus set the measurement planes at the soil sample itself. Experimental results for teflon and soil samples with different water content and salinity verified the validity of our approach.

Acknowledgments

Arkadiusz Lewandowski has been supported by the Polish National Science Center in the framework of the project no. 2011/03/D/ST7/01731.

Appendix

Below we briefly summarize the extraction algorithm. From transmission parameters of the MUT we obtain scattering parameters by inverting (2) which yields the following model of the MUT measurement:

\[
S_M = \frac{1}{1 - \Gamma_M e^{-2\gamma_M}} \left( (1 - e^{-2\gamma_M}) \Gamma_M - e^{-\gamma_M}(1 - \Gamma_M^2) ight)
\]  

(A.1)

where for \( Z_{\text{ref}} \) set to \( Z_{\text{0A}} = \eta / \sqrt{\varepsilon_{\text{A}}} \) we further have:

\[
\Gamma_M = \left( \frac{1 - \frac{\varepsilon_{\text{M}}}{\varepsilon_{\text{A}}} \sqrt{\varepsilon_{\text{M}}} \varepsilon_{\text{A}}}{1 + \frac{\varepsilon_{\text{M}}}{\varepsilon_{\text{A}}} \sqrt{\varepsilon_{\text{M}}} \varepsilon_{\text{A}}} \right), \quad \gamma_M = \frac{\omega}{c} \sqrt{\varepsilon_{\text{M}}}. \]

(A.2)

The problem now is to determine \( \varepsilon_{\text{M}} \) from the actual scattering parameters of the MUT obtained by applying the correction equation (4) to the measured transmission matrix \( T_{\text{cell}} \) of the cell. From the classical Nicholson–Ross–Weir equations [9, 23] we obtain the estimate of \( \Gamma_M \) as:

\[
\hat{\Gamma}_M = K \pm \sqrt{K^2 - 1}, \quad \text{where} \quad K = \frac{S_{11}^2 - S_{21}^2 + 1}{2S_{11}}. \]

(A.3)

In order to choose the proper sign in (A.3), we enforce \( |\hat{\Gamma}_M| \leq 1 \). From this estimate we can then determine the estimate of the term \( T = e^{-\gamma_M} \) as:

\[
\hat{T} = \frac{S_{11} + S_{21} - \hat{\Gamma}_M}{1 - (S_{11} + S_{21})\hat{\Gamma}_M}. \]

(A.4)

Consecutively, we obtain the estimate of \( \gamma_M \) as:

\[
\hat{\gamma}_M = -\ln \frac{\hat{T}}{T}. \]

(A.5)

In order to select the proper root of the natural logarithm, we enforce the continuous change of \( \hat{\gamma}_M \) as a function of frequency. Eventually, we determine the relative dielectric permittivity of the MUT as:

\[
\hat{\varepsilon}_M = -\frac{\hat{\gamma}_M^2}{\omega}. \]

(A.6)

As we can see, the choice of \( Z_{\text{ref}} \) does not affect the extraction algorithm. Indeed, it only changes the reflection coefficient \( \Gamma_M \), however, this term is not further decomposed in order to solve for \( \varepsilon_{\text{M}} \).

References

[1] Robinson D A, Campbell C S, Hopmans J W, Hornbuckle B K, Jones S B, Knight R, Ogden F, Selker J and Wendroth O 2008 Soil moisture measurement for ecological and hydrological watershed-scale observatories: a review Vadose Zone J. 7 358–89
[2] Bittelli M 2011 Measuring soil water content: a review HortTechnology 21 293–300
[3] Kaatze U and Huebner C 2010 Electromagnetic techniques for moisture content determination of materials Meas. Sci. Technol. 21 082001
[4] Topp G C, Davis J L and Annan A P 1980 Electromagnetic determination of soil water content: measurement in coaxial transmission lines Water Resour. Res. 16 574–82
[5] Noborio K 2001 Measurement of soil water content and electrical conductivity by time domain reflectometry: a review Comput. Electron. Agric. 31 213–37
[6] Skierucha W, Wilczek A, Szyplowska A, Slawiński C and Lamorski K 2012 A trd-based soil moisture monitoring system with simultaneous measurement of soil temperature and electrical conductivity Sensors 12 13545–66
[7] Kaatze U and Feldman Y 2006 Broadband dielectric spectrometry of liquids and biosystems Meas. Sci. Technol. 17 R17–35
[8] Szyplowska A, Wilczek A, Kafarski M and Skierucha W 2016 Soil complex dielectric permittivity spectra determination using electrical signal reflections in probes of various lengths Vadose Zone J. 15
[9] Baker-Jarvis J 1990 TN 1341: transmission/reflection and short-circuit line permittivity measurements Technical Report NIST
[10] Wagner N, Schwing M and Scheuermann A 2014 Numerical 3-d fem and experimental analysis of the open-ended coaxial line technique for microwave dielectric spectroscopy on soil IEEE Trans. Geosci. Remote Sens. 52 880–93

[11] Wagner N, Emmerich K, Bonitz F and Kupfer K 2011 Experimental investigations on the frequency- and temperature-dependent dielectric material properties of soil IEEE Trans. Geosci. Remote Sens. 49 2518–30

[12] Bobrov P P, Repin A V and Rodionova O V 2015 Wideband frequency domain method of soil dielectric property measurements IEEE Trans. Geosci. Remote Sens. 53 2366–72

[13] Ba D and Sabouroux P 2010 Epsimu, a toolkit for permittivity and permeability measurement in microwave domain at real time of all materials: applications to solid and semisolid materials Microw. Opt. Technol. Lett. 52 2643–8

[14] Lauer K, Wagner N and Felix-Henningsen P 2012 A new technique for measuring broadband dielectric spectra of undisturbed soil samples Eur. J. Soil Sci. 63 224–38

[15] Marks R B 1991 A multiline method of network analyzer calibration IEEE Trans. Microw. Theory Technol. 39 1205–15

[16] Lewandowski A, Wiatr W, Opalski L J and Biedrzycki R 2015 Accuracy and bandwidth optimization of the over-determined offset-short reflectometer calibration IEEE Trans. Microw. Theory Technol. 63 1076–1089

[17] MathWorks 2014 MATLAB R2014a www.mathworks.com/products/matlab/

[18] Marks R B 1997 Formulations of the basic vector network analyzer error model including switch-terms 50th ARFTG Conf. Digest (Portland, OR, December 1997) vol 32 pp 115–26

[19] Kerns D M and Beatty R W 1967 Basic Theory of Waveguide Junctions and Introductory Microwave Network Analysis (Oxford: Pergamon)

[20] Marks R B and Williams D F 1992 A general waveguide circuit theory Natl. Inst. Stand. Technol. J. Res. 97 533–62

[21] Bieganowski A, Chojecki T, Ryżak M, Sochan A and Lamorski K 2013 Methodological aspects of fractal dimension estimation on the basis of psd Vadose Zone J. 12

[22] Rhoades J D, Chanduvi F and Lesch S 1999 Soil Salinity Assessment, Interpretation of Electrical Conductivity Measurements (Food and Agriculture Organization of the United Nations)

[23] Boughriet A-H, Legrand C and Chapoton A 1997 Noniterative stable transmission/reflection method for low-loss material complex permittivity determination IEEE Trans. Microw. Theory Technol. 45 52–7