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To cite this article: Bektaş Çolak & Selçuk Helhel (2019) A new error reduction technique for reflection coefficient measurements for use in quick laboratory tests, International Journal of Electronics, 106:2, 237-249, DOI: 10.1080/00207217.2018.1523473

To link to this article: https://doi.org/10.1080/00207217.2018.1523473

Accepted author version posted online: 14 Sep 2018.
Published online: 12 Oct 2018.

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A new error reduction technique for reflection coefficient measurements for use in quick laboratory tests

Bektaş Çolak\textsuperscript{a} and Selçuk Helhel \textsuperscript{a,b}

\textsuperscript{a}Engineering Faculty, Department of Electrical and Electronics, Alanya Alaaddin Keykubat University, Alanya, Antalya, Turkey; \textsuperscript{b}Engineering Faculty, Department of Electrical and Electronics Engineering, Akdeniz University, Antalya, Turkey

\textbf{ABSTRACT}
Reflection coefficient measurement is one of those fundamental methods for determining the performance of radar absorber materials, and reflections due to these measurements are originated from the discontinuities in the measurement system and they result in error. In this paper, a new technique for the calibration of the scalar reflection coefficient measurements in the frequency range of 3–18 GHz in a portable two-sectioned metallic chamber is being proposed for use in quick laboratory tests with reasonable error (redacted error). Calibration measurements are performed by two calibration standards in frequency domain and transformed to time domain for further calculation. Proposed technique exhibits a good agreement with the theoretical values for especially in compact chambers where plane-wave conditions are not fully satisfied, and it is not in need of any complex time gating process. Furthermore, it does not require considering the phase differences which occurs at the measurement plane, providing a cheap and faster solution, reducing time and complexity in characterization of radar absorber materials. This technique with the mentioned test set-up could supply scalar reflection coefficient measurements with overall 0.55 dB error level and useful for practical applications where high level of accuracy is not required.

\textbf{1. Introduction}
Radar-absorbing materials (RAM) can be evaluated by one port reflection measurement system. Measurements are available either in a free space [Rocha, L. S et al. 2013, Kemptner et al. (2012), Smith, Chambers, and Bennett (1992), Orlob, Member, Reinecke, and Denicke (2013), and Bartley and Begley (2012)] or in guided-wave systems (Zhao, Jiang, & Jing, 2011). The guided wave measurement set-ups are more robust than free space measurement system, but they are so sensitive to the sample sizes. Free space set-ups include more error terms than the guided wave measurement setups. The source of errors and the available elimination techniques of them are given by Smith et al. (1992). One of the key errors in free-space measurements is the source match error. The gating technique in time domain is the most famous one to remove the source match errors (Bartley & Begley, 2012; Shirai & Ishikawa, 2010; Dunsmore, 2008). However, time domain gating is a sensitive process and introduces errors if not used carefully (Dunsmore, 2008). For
example, the gating windows and phase synchronization of the sample with calibration standards are crucial for obtaining accurate results. Majority of the reflection coefficient measurements require at least three calibration standards along with various error correction techniques. The number of calibration standards could be reduced, for example, by shifting the calibration plane Smith et al. (1992). In this technique, the position of the conductive calibration standard, resembling the short part of the ‘open, short, load’ calibration cycle is shifted to two other predefined positions. Usage of calibration plane technique requires managing the positioning of the sample and calibration standards as well as additional signal processing techniques (Gu, Houtz, Randa, & Walker, 2011). Therefore, if the shifted calibration positions at measurement set-up are not fixed accurately for each measurement, the phase errors cause erroneous results. In order to prevent these errors, the measurement systems require a very accurate and sensitive positioning mechanism, increasing the cost significantly. Since maxima and minima points can be observed due to sample thickness (half-wave transformer and quarter wave transformer like phenomena may occur at some frequencies), a compensation algorithm for the deterministic phase differences originated from the thickness of the sample plate is needed (Min et al. 2006). Different calibration techniques using shifting calibration plane is summarized in Akay, Kharkovsky, and Hasar (2001) and Stumper, Member, and Schrader (2014). Meanwhile, there are still no more works for the calibration of low-cost set-ups for the compact screened chambers.

This paper presents a new technique (Colak, 2016) for the scalar one port reflection measurements in a portable two sectioned metallic chamber for use in quick laboratory tests with an acceptable error. The proposed technique is not in need of any complex time gating process and does not require considering the phase differences occurring at the measurement plane, and it is providing a fast and simple characterization of RAM. During this study, the calibration measurements are performed in frequency domain and transformed to time domain for further calculation using inverse Fourier transforms (IFT).

This paper is organized as follows: first, the measurement set-up for the proposed technique is described. Subsequently, the details of the technique are discussed. Afterwards, the results of proposed error reduction technique are compared with three standards method introduced by Smith et al. (1992) to validate the proposed technique. Finally, the paper is concluded with a brief summary and the Appendix.

2. Test set-up

A portable two-sectioned metallic chamber shown in Figure 1 is used for the measurements. The chamber’s dimensions are 2 m x 1 m x 1m, and it is built to isolate measurement environment from outside radiation to achieve more reliable test results. The inner walls of the chamber are covered by 20 cm long pyramidal absorbers. The chamber is designed for performing both reflection and transmission measurements. In this work, only the reflection part is used. There is a removable metallic thin plate between two sections to perform as a reference measurement plane. There is a 50 x 50cm square opening at this conductive reference plate in which the sample of the test object (or RAM) and calibration standards are placed. The thickness of the conductive reference plate is 1.5 mm. The circumference of the open window is fixed by a conductive surface and covered by the absorbers as shown in Figure 1(b). The absorbers on the reference plane are available only in the reflection antenna side of the chamber. The antenna used in the test set-up is a double-ridged horn antenna and the measured sample is within the main lobe of this antenna. The Vector Network Analyzer, (VNA) used in the measurements is an HP E8363B.
3. New proposed error reduction technique

3.1. Geometry

The reflection geometry is illustrated in Figure 2. Thickness of the dielectric plate, $d$, introduces a phase difference in the time domain response of both conductive and dielectric plate measurements. We used a double-ridged horn antenna having an average gain of 10 dB in the frequency range of 218 GHz.

The reference point is starting at the face of the metallic plate looking at the antenna. So this point is assigned as $z=0$, where EM waves propagate in the $z$ direction. As a comparison, the theoretical reflection coefficient, $\Gamma_{dB}$, is calculated by Equation (A.1) given in the Appendix. This equation includes from both edges of a dielectric plate with finite thickness, $d$. At some
frequencies, maximum and minimum points occur because of the cancellation or addition of both edges reflections. It is assumed that the sample dielectric constant, \( \varepsilon_r \), is greater than the unity, therefore it can be derived from well-known equations on maximum and minimum points of reflection coefficient for normal incidence of TEM wave at multiple dielectric layer that the maximum value of reflection coefficient occurs at frequency points, \( f_{\text{max}} \), is defined as in Equation (1):
\[
f_{\text{max}} \mid_n = \frac{c}{4d\sqrt{\varepsilon_r}}(2n + 1) \quad n = 0, 1, 2,
\]
where \( c, n \) and \( \varepsilon_r \) are the speed of the propagation in free space, the index of the repetitive frequencies relating maximum value of reflection coefficient and relative dielectric constant of the medium. As a simplicity, two sequential maximum points can be obtained by using the frequency values at \( n=0 \) and \( n=1 \). Then, a period of the maximum points in the frequency domain, \( T \), can easily be obtained by Equation (2).
\[
T = (f_{\text{max}} \mid_{n=1}) - (f_{\text{max}} \mid_{n=0})
\]

3.2. Calibration

For the calibration of the measurement set-up, two calibration standards and one sample plate measurements are performed.

3.2.1. Measurements

The first measurement is performed in an object free room as a reference measurement for the zero reflection (\( \Gamma = 0 \)) and is represented by \( E \), where \( \Gamma \) is the reflection coefficient including all effects of test antenna, cable and chamber interactions. The second measurement is performed with a metallic plate at reference point for the full reflection condition (\( \Gamma = -1 \)) and is represented by \( P \). These two measurements resemble the two calibration standards. Finally, in the third measurement, dielectric plate sample is placed in the reference plane; measurements are completed represented by \( S \) in the equations. \( E, P \) and \( S \) are reflection ‘\( S_{11} \)’ measurements in complex form. Set-up is described in Section 2. The samples used for the measurement are obtained from the already available products in the commercial market.

3.2.2. Time domain response by IFT

Subsequently, the empty room, conductive plate and sample dielectric measurement data, \( E, P \) and \( S \), respectively, are converted to time domain using IFT. In order to simplify the calibration procedure for use in quick tests, the most simplest windowing technique (rectangular window) is preferred, and the IFT is applied on all data without any special windowing.

3.2.3. Normalization

Once the time domain responses are obtained by using IFT on the frequency domain data from two calibration standards and the sample dielectric plate, the conductive plate time response measurement, \( P_t \), and the sample dielectric plate time response, \( S_t \), are normalized to empty room time response, \( E_t \). Those are expressed as in Equations (3 and (4).
\[
P_n = P_t - E_t
\]
\[
S_n = S_t - E_t
\]
where \( P_n \) and \( S_n \) represent the normalized complex data. Subtraction performed with linear scale in complex domain. Since the conductive plate and sample dielectric plate time responses (\( P_t \) and \( S_t \)) are normalised to empty room time response, \( E_t \), it is certain that by subtracting the reference time
domain response of ‘empty room measurement’, \( E_r \), from \( P_t \) and \( S_t \) data, the peak point of each normalised \( P_t \) and \( S_t \) data in time domain gives the relative reflection coefficient information about the reference point. With this confidence, it is not necessary to think about the time domain gating centre or span values. This property reflects our technique’s simplicity. It is also observed that if only amplitude values of frequency domain data entered to the IFT instead of the complex values, then obtained reflection coefficient will be similar to the ones obtained with complex values. So with this technique, a scalar network analyser or a signal generator with a spectrum analyser combination can be used.

### 3.2.4. Extracting the reflection coefficient

Using the normalised time domain data, one can obtain the peak values \( P_{\text{max}} \) and \( S_{\text{max}} \) as in Equations (5) and (6) to represent the relative reflected levels.

\[
P_{\text{max}} = \max |P_n| \tag{5}
\]

\[
S_{\text{max}} = \max |S_n| \tag{6}
\]

By dividing the \( S_{\text{max}} \) to \( P_{\text{max}} \), it is possible to get the absolute value of the reflection coefficient of the sample dielectric plate, \( R_{\text{dB}} \) (in dB).

\[
R_{\text{dB}} = 20 \log_{10} \left( \frac{S_{\text{max}}}{P_{\text{max}}} \right) \tag{7}
\]

The specified frequency bandwidth used in one single IFT to obtain single reflection coefficient. To get wideband reflection coefficients for the full frequency bandwidth of interest (e.g. 3–18 GHz), this whole band is divided into sub-frequency ranges. Each sub-frequency range provides a single reflection coefficient value in time domain through the aforementioned steps. This single value is then assigned to the middle frequency of this sub-frequency range. This sub-frequency range window is then step-by-step moved from the lower side (3 GHz) to the higher side (18 GHz) of the whole frequency band. In each movement step, a unique reflection coefficient is calculated from the corresponding sub-frequency range, constructing a wide band reflection coefficient data for the full frequency bandwidth (e.g. 3–18 GHz). Figure 3 illustrates the step-by step movement of sub-frequency windows.

The bandwidth of this sub-frequency range represented with \( b \), whereas the step size is presented with \( s \). The total bandwidth of the frequency of interest is represented by \( B_w \), and it can be calculated by means of Equation (8).

\[
B_w = f_e - f_s \tag{8}
\]

where \( f_e \) is the overall end frequency (it is 18 GHz), while \( f_s \) is the overall start frequency (it is 3 GHz). Overall bandwidth, \( B_w \) is sampled by \( N \) points where \( N=20,001 \).

![Figure 3. Shifting frequency window.](image-url)
Therefore, each separate frequency step, $s_f$, determined by Equation (9).

$$s_f = \frac{B_w}{(N - 1)} \tag{9}$$

The number of sub-frequency ranges, namely $i$, in the total data points $N$ is obtained by Equation (10).

$$i = \left\lceil \frac{B_w - b}{s} \right\rceil + 1 \tag{10}$$

The first data point, $f_{sk}$, last data point, $f_{ek}$, and midpoint frequency, $f_{mk}$, are defined for each $k$ by means of Equations (11)–(13).

$$f_{sk} = f_s + s_f(k - 1)s \tag{11}$$

$$f_{ek} = f_{sk} + s_f s \tag{12}$$

$$f_{mk} = \frac{f_{sk} + (s_f s)}{2} \tag{13}$$

where $k=1,2,3,\ldots,i$.

In Figure 3, shifting of frequency window is shown.

### 3.2.5. Selection of optimum bandwidth for IFT

The bandwidth of this sub-frequency range, $b$, is important for the accuracy of the time domain reflection coefficient data. For example, if an inadequate number of points are used, then one cannot obtain the relative reflection coefficient at reference plane since required minimum resolution in time domain could not be supplied. On the other hand, if the bandwidth of this sub-frequency range is selected so wide, then the resolution in the frequency domain is decreased, therefore max and min points of the reflection coefficients could not be detected correctly. Therefore, a prior calculation should be done for every different sample to define the optimum bandwidth. After a prior study, it has been observed that the upper limit of the bandwidth of the sub-frequency range is related with the thickness and the dielectric constant of the sample dielectric plate whereas lower limit is related with the distance of the measurement antenna and the reference plane, $l$, which defines also the minimum resolution limit at $z$ axis. To find the optimum sub-frequency range bandwidth, $b$, some additional measurements are done. Four different thickness of the same dielectric plate is measured separately and tried to find the optimum sub-frequency range bandwidth value, $b$, for each sample. As a performance check, the theoretical reflection coefficient, $\Gamma_{db}$, calculated by Equation (A.1) given in the Appendix is subtracted from the reflection coefficient value, $R_{db}$, obtained from measurements, therefore for that midpoint of the related sub-frequency range, $f_{mk}$, an absolute error term, $e_k$, (in dB) is expressed as in Equation (14).

$$e_k = |R_{db} - \Gamma_{db}| \tag{14}$$

The total RMS error, $e_T$, by using the error terms at each frequency point is calculated by Equation (15).

$$e_T = \frac{1}{i} \sqrt{\sum_{k=1}^{i} (e_k)^2} \tag{15}$$

The comparative results are summarized in Table 1. The best values are taken when the $b$ is equal to around $T/4$. For the upper limit, increasing the sub frequency bandwidth more than one-third of $T$ also causes to lose the resolution at frequency domain in a manner that the general sinusoidal form of the reflection coefficient will start to demolish. So as the maximum bandwidth of
the sub-frequency range, $b_{\text{max}}$, it should not bigger than the $T/3$ for this measurement set-up, and it can be calculated by Equation (16).

$$b_{\text{max}} = \frac{T}{3} = \frac{c}{6d\sqrt{\varepsilon_r}}$$  \hspace{1cm} (16)

As a practical concern in RAM measurements, $\varepsilon_r$ is not known generally before the measurements since one of the reason of the measuring the reflection coefficient is also to define the relative dielectric constant of the RAM. As a solution to this problem, at least one prior calculation is taken before the final one with $b_{\text{max}}/2$ value for the sub-frequency range bandwidth whereas assuming the $\varepsilon_r = 1$. Then, the period of the maximum points in the frequency domain $T$ is obtained from these results by finding maximum values in whole frequency range. Once $T$ is obtained, then Equation (16) can be used to define the $b_{\text{max}}$. For the lower limit, the equivalent resolution in time domain should not lower than the required two-way travelling time of the waves from antenna to the reference plane at least. Then, the minimum bandwidth of the sub-frequency range, $b_{\text{min}}$, is expressed by Equation (17) obtained from well-known radar range resolution equations. So this formula gives the theoretical limit of the bandwidth for required minimum resolution in range or in time domain. If the sub-frequency bandwidth is chosen at its limit value the reference position could not be detected accurately, and maximum error is observed at this point. In order to decrease the observed error, it is necessary to increase the sub-frequency bandwidth. As will be discussed through the paper, proposed method empirically suggests the optimum bandwidth should be around $T/4$. For examined plates in the study, the optimum bandwidth is at least 1.5 times bigger than the calculated minimum sub-frequency bandwidth.

$$b_{\text{min}} = \frac{c}{2l}$$  \hspace{1cm} (17)

On the other hand, the step size, $s$, is taken as $b/4$ for the simplicity since its value is not as critical as $b$. Since the time response is obtained from a limited number of samples, the exact peak value could not be detected accurately. Additional comparative measurements are done in order to define the effect of the number of sample points in frequency domain to reflect the performance of the IFT. Four different values are assigned as the number of sample points without changing the total bandwidth. Then, the error term, $e_k$, is calculated with Equation (14) and results are summarized in Table 2. In Table 2, calculated error values between theory and the measurement for different number of samples is tabulated. The start and stop frequency of the related frequency band are 4.7 GHz and 6.3 GHz respectively. (d=10 mm and $\varepsilon_r=2.26$).

<table>
<thead>
<tr>
<th>$d$ (mm)</th>
<th>$e_T$ ($b=T/2$)</th>
<th>$e_T$ ($b=T/3$)</th>
<th>$e_T$ ($b=T/4$)</th>
<th>$e_T$ ($b=T/5$)</th>
<th>$e_T$ ($b=T/6$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>0.36</td>
<td>0.26</td>
<td>0.14</td>
<td>0.17</td>
<td>0.16</td>
</tr>
<tr>
<td>15</td>
<td>0.36</td>
<td>0.30</td>
<td>0.29</td>
<td>0.29</td>
<td>0.29</td>
</tr>
<tr>
<td>30</td>
<td>0.40</td>
<td>0.32</td>
<td>0.32</td>
<td>0.33</td>
<td>0.32</td>
</tr>
<tr>
<td>40</td>
<td>0.85</td>
<td>0.39</td>
<td>0.34</td>
<td>0.39</td>
<td>0.47</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>$N$</th>
<th>$\Delta \phi$, MHz</th>
<th>$e_k$</th>
</tr>
</thead>
<tbody>
<tr>
<td>800</td>
<td>2</td>
<td>0.29</td>
</tr>
<tr>
<td>1600</td>
<td>1</td>
<td>0.28</td>
</tr>
<tr>
<td>3200</td>
<td>0.5</td>
<td>0.27</td>
</tr>
<tr>
<td>6400</td>
<td>0.25</td>
<td>0.26</td>
</tr>
</tbody>
</table>

Table 1. Total average error.

Table 2. Error for different $N$ in same bandwidth.
by using two subsequent frequency range, The final reflection coefficient calculation is shown in Figure 4.

The sharp nulls could be missed because the averaging property of technique on reflecting coefficient data over this subsequent region while this is not a big disadvantage since the lowest range is around $-23$ dB and this value is enough for practical set-ups.

3.2.6. Repeatability and uncertainty of this technique

The set-up is being used for the past three years, and measurements with the same sample have been carried out at yearly basis to be sure about the stability of setup. An acceptable deviation has been observed at yearly basis measurements that stability of set-up is in reasonable range. Those measurements are shown in Figure 5. The fluctuations are due to measurement error and not due to stability of set-up.

The maximum variation occurs under $-24$ dB limit as 6 dB. For the other regions where the reflection coefficient is bigger than $-24$ dB, the biggest variation is in 1.5 dB. The most important components of uncertainty come from the placement of the sample at reference place, roughness at the surface and thickness of the sample and alignment of the antenna. The contribution of each factor is investigated separately.

3.2.7. Placement of the sample at reference plane

The conductive plate and the sample plate are fixed to the reference plane by adhesive plastic tape at the beginning of this work. It is realized that sometimes the upper two corners of the plates are tilted itself backward from reference plane because of the weight of the sample. It causes a
variation at angle of normal vector of the sample plate front plane. And also the conductivity of the conductive plate to the reference plane is decreased. The level of uncertainty of this problem is investigated by fixing the conductive and sample plate within 3° angular tilt and the results are evaluated whereas the other contribution of the uncertainty is fixed. The results show 0.5 dB variation.

3.2.8. Roughness at surface and thickness of the sample

The samples are obtained from industry market. Since these samples are not designed for sensitive laboratory measurements, the thickness of the sample is not accurate as it presented in the market and measured within 10% tolerance. A calibrated micrometer is used for the final thickness of the sample at five times. And the average value of it is used in calculations. On the other hand, the measurements can be done with a micrometer only at the edges of the plate whereas some samples include the thickness variation within the other middle region of the rectangular shape. One of the solutions to this problem is to use a smaller size of the sample but it changes some parameters of the set-up such as distance of the antenna to the reference plane. Finally, three different samples with same thickness and dielectric constant are measured whereas the other contribution of the uncertainty is fixed. The variation between each calculated result is in 0.4 dB.

3.2.9. Alignment of the antenna

Since it is possible to obtain additional error due to mis-alignment of antennas, separate measurements with new antenna installation are held whereas the other contribution of the uncertainty is fixed. Once all the above parameters are fixed and stabilized for one test set-up, the variation between the time measurement results decreases to 0.1 dB which is expected as a result of equipment un-stability and non-linearity properties.

4. Measurement results

The acceptance criterion of the measurement error is taken as 1 dB. As seen in Figure 6, Figure 7 and Figure 8 those measurements are in good track with theoretical values in quite a good manner. Minima observed in those figures are due to the sample thickness as mentioned before. Calculated RMS error is about 0.35 dB for the sample having a thickness of 15 mm, 0.45 dB for the sample having a thickness of 30 mm, and 0.55 dB for the sample having a thickness of 40 mm as shown in Figure 6. These RMS values are in the acceptable range. It is observed that increased thickness of sample results in increased RMS error, but this increment is at negligible levels.

Even the sample thickness is changed from 15 to 30 mm and then to 40 mm there is still a variation as 1.65 dB at 8.47 GHz for 40 mm and 2.19 dB for 15 mm. Those peak error values may be due to the chamber characteristics. In order to check the performance of the technique on

![Figure 6](image_url)

**Figure 6.** Calculated reflection coefficient for d=15 mm and \( \varepsilon_r=2.26 \) (solid: measured, dashed: theoretical).
different dielectric plates, two different materials were also measured and results are given in Figure 9 and in Figure 10. Maximum variation of error for these samples has been calculated as

Figure 7. Calculated reflection coefficient for $d=30\,\text{mm}$ and $\varepsilon_r=2.26$ (solid: measured, dashed: theoretical), RMS error=0.45 dB.

Figure 8. Calculated reflection coefficient for $d=40\,\text{mm}$ and $\varepsilon_r=2.26$ (solid: measured, dashed: theoretical), RMS error=0.55 dB.

Figure 9. Calculated reflection coefficient for $d=4.1\,\text{mm}$ and $\varepsilon_r=2.85$ (solid: measured, dashed: theoretical), RMS error=0.50 dB.
0.55 dB. But for both samples, calculated RMS errors are 0.5 and 0.45 dB, respectively, in acceptable limits.

From the above various graphical results, it can be concluded that the total error level is increasing while the thickness of the sample is increasing. This is also confirmed by results shown in Table 1. On the other hand, the minimum bandwidth limit $b_{min}$ can also be used to define the thickness limit of the system. This is the value that the optimum bandwidth, $b$, namely $T/4$ reaches to the level of $b_{min}$, for a combination of $d=60$ mm with $\varepsilon_r = 2.26$. Also, there is a mechanical concern that increased sample thickness results in increasing weight. Therefore, for the current set-up, fixing that much heavy sample firmly to the reference plate is not an easy task, and those samples are introducing more errors to uncertainty budget.

5. Conclusion

We have shown a new simple and fast error reduction method for the measurement of scalar reflection coefficient in a portable two-sectioned metallic chamber for use in quick laboratory tests. Additional measurements are performed in order to see the performance of the method. On the other hand, we have used a well-known free-space calibration technique from literature (Smith et al., 1992) and found that this technique is not so suitable for our chamber because of its compactness. Followings are benefits of offered method;

- Offered method predicts very good results that they are in good agreement with the far field reflection measurements with reasonable RMS error.
- It is much faster in final measurement steps than the three standard methods.
- It is simple because it does not need to consider the possible phase differences which may occur due to calibration plane positioning Therefore, it is not required to deal with detailed time domain gating process.
- It is quite useful method for practical applications especially at R&D stages that they do not require exact results (presented method has still a reasonable RMS error at the level 0.55 dB). If more accurate levels of the reflection coefficient are required, then the error sources of the chamber should be decreased with an extra cost by increasing the mechanical stability at reference plane and antenna placement.
Acknowledgments

We would like to thank the Directorate of Akdeniz University Industrial Based Microwave Research Center (EMUMAM- Grant Number: 2007K120530-DPT) and Akdeniz University BAP.

Disclosure statement

No potential conflict of interest was reported by the authors.

Funding

This work was supported by the Devlet Planlama Örgütü [2007K120530-DPT].

ORCID

Selçuk Helhel  http://orcid.org/0000-0002-1401-3297

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Appendix

Reflection Coefficient from Dielectric Plate

Theoretical reflection coefficient of finite thickness dielectric plate, $\Gamma_{db}$, is obtained by using multiple reflections at the boundary of the plate which is expressed in Equation (A1). In this equation, $d$ is the thickness of the dielectric plate, $\varepsilon_r$ is the relative dielectric constant of the media and $\lambda_0$ is the wavelength in the free space environment.

$$\Gamma_{db} = 20\log_{10}\left(\frac{r_{12}(1 - e^{-j\Theta})}{1 - r_{12}e^{-j\Theta}}\right)$$

(A1)

where,

$$\Theta = k_s d$$
$$k_s = \frac{2\pi}{\lambda_0} \sqrt{\varepsilon_r}$$
$$r_{12} = \frac{1 - \sqrt{\varepsilon_r}}{1 + \sqrt{\varepsilon_r}}$$

(A2)