Modeling Planar Arbitrarily Shaped Microstrip Elements in Multilayered Media

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Abstract—Microstrip elements of arbitrary shape are modeled in multilayered media. The Green’s function for the multilayered structure is developed in a form useful for efficient computation for interacting microstrip elements, which may be located at any substrate layer and separated by an arbitrarily large distance. This result is of significant value to a variety of applications in wave propagation besides those discussed in this paper. The mixed-potential integral-equation (MPIE) method is developed in the spatial domain. Examples for regularly/arbitrarily shaped geometries in single and multilayered media are presented. These involve the optimization of an open-end microstrip, a radial-stub microstrip, a five-section overlay-gap-coupled filter, and a circular-patch proximity-coupled microstrip antenna. Very good agreement with measurement and other published data is observed.

I. INTRODUCTION

In the design of microwave monolithic integrated circuits (MMIC’s) and millimeter-wave integrated circuits, electromagnetic (EM) modeling of microstrip elements (interconnects, antennas, and circuits) becomes important as the operating frequency becomes higher. Full-wave analysis includes the effects of EM coupling, surface waves, and radiation loss while traditional quasi-static methods [1], [2] and equivalent waveguide models [3] fail to yield sufficiently accurate results. Moreover, the ability to analyze circuits of arbitrary shape embedded in a multilayered medium allows for more versatile designs with higher density. The authors have selected the mixed-potential integral equation (MPIE) method [4]–[7] combined with triangular-patch function [8] to solve this general configuration. The MPIE applied in the spatial domain was determined to be more efficient for numerical modeling of arbitrarily shaped printed circuits than the electric-field integral equation (EFIE) techniques applied in the spectral domain [9]. In addition, this method gives a better physical interpretation of current flow and field distribution.

The triangular basis function has been successfully used to solve irregular scatterers [8], [10], and [33]. It describes the vector surface current, matches the boundary conditions with no normal components along the boundary edges, and has the potential to represent nonplanar current flow. The unknown current distribution on microstrip elements can be modeled very accurately with triangular basis functions used in the method of moments (MoM’s).

Derivation of the spatial-domain dyadic Green’s function begins with closed-form formulation in the spectral domain. Several papers using the MPIE formulation have given explicit expressions for single-layered or double-layered structures [12]–[15]. The field and source points were assumed to be embedded in the same layer, which is not applicable for cases with more than three layers. In this paper, the authors derive two different formulations for multilayered structures, and the position of the field point dictates which formulation is chosen. This Green’s function can be evaluated without difficulty for any number of layers. To obtain the Green’s function in the spatial domain, Sommerfeld integrals must be evaluated. Recently, the complex image method (CIM) [14], [15] was proposed to approximate the integrand with several complex exponential terms, making the computation very efficient. However, the following parameters must be determined before applying the CIM: sampling range, expansion terms, and surface-wave poles. The first two parameters depend strongly on the properties of the layers, and the surface-wave poles of a medium with more than three layers are difficult to find. Therefore, the CIM is not well suited for general geometries with multiple layers. Instead, the authors employ the traditional quadrature method to evaluate Sommerfeld integrals, and apply several techniques to speed up the computations.

The MPIE–MoM analysis presented here starts by meshing the whole geometry with small triangular facets using the meshing algorithm developed by [16]. The spatial-domain Green’s function is evaluated over the range of microstrip elements using Sommerfeld integrals, then stored as numerical tables for interpolation. Galerkin’s procedure is followed to create a matrix equation from the MPIE. These procedures will be discussed in Section II. Four examples have been investigated and compared with either measurements or published data in Section III.

II. MPIE FORMULATION

A generalized multilayered medium is shown in Fig. 1. Each substrate layer \( i \) has a thickness \( d_i \), and relative permit-
tivity and permeability $\epsilon_r$ and $\mu_r$, respectively. The upper and lower ground planes are removable to represent either a shielded, semi-open, or open structure. The medium is assumed to be infinite in the $x$–$y$ plane, and microstrip patterns are assumed to be of infinitesimal conductor thickness.

An EFIE can be set up by applying the boundary condition of the zero tangential field on metal surface $S$, as

$$\hat{n} \times \mathbf{E}^{\text{inc}}(\vec{r}) = -\hat{n} \times \int_{S} \mathbf{G}_E(\vec{r}|\vec{r}_s) \cdot \mathbf{J}_s(\vec{r}_s) dS_s$$

where $\mathbf{G}_E$ is the dyadic electric-field Green’s function, and $\mathbf{J}_s$ is the surface current distribution. For simplicity, the authors neglect conductor and dielectric losses. If the magnetic vector potential ($\vec{A}$) and charge scalar potential ($V$) as $\mathbf{E} = \mathbf{j} \omega \vec{A} - \nabla V$ are introduced, then (1) can be rewritten as an MPIE

$$\hat{n} \times \mathbf{E}^{\text{inc}}(\vec{r}) = \hat{n} \times [\mathbf{j} \omega \int_{S} \mathbf{G}_E(\vec{r}|\vec{r}_s) \cdot \mathbf{J}_s(\vec{r}_s) dS_s + \nabla \int_{S} \mathbf{G}_q(\vec{r}|\vec{r}_s) \mathbf{E}_s(\vec{r}_s) dS_s]$$

where $\mathbf{G}_E$ and $\mathbf{G}_q$ are the dyadic Green’s functions for $\vec{A}$ and $V$, respectively. Both Green’s functions are first derived analytically in the spectral domain, then evaluated in the spatial domain using Sommerfeld integrals, discussed in the following section. $\mathbf{J}_s$ and $\mathbf{q}_s$ are the unknown distributions of the electric surface current and charge. They are connected to each other by the continuity equation. The reason the authors use the MPIE is to avoid the two-dimensional (2-D) infinite integrals with highly oscillating and slowly decaying integrands that must be evaluated in the EFIE. After removing the singular term from the MPIE’s integrand, which occurs when source and field points coincide and can be computed in closed-form [17], only surface integrals with well-behaved functions over small areas need to be calculated.

A. Dyadic Green’s Function

Before applying the MoM, the dyadic Green’s function for a multilayered medium is derived. In [11]–[15] only the Green’s function with the field point in the source layer is discussed. As a result, planar circuits in conventional microstrip with two dielectric layers can be investigated, but circuits with three or more dielectric layers cannot be dealt with. Here, the spectral-domain Green’s function is derived in a general form. It can be written as the sum of TE and TM $\pm \hat{z}$ and $-\hat{z}$ propagating waves due to the reflection and transmission between different layers, no matter where the field point is. The generalized reflection and transmission coefficients [18] are obtained from the boundary conditions. For planar microstrip configurations, only $G_{TE}^{xz}$ and $G_{TM}^{xz}$ due to a horizontal electric dipole (HED) are involved. They are listed as follows, depending on the location of the field point.

1) Source Layer:

$$G_{TE}^{(i)xx} = \frac{\lambda_0 \mu_r}{2 \gamma_i} \left\{ e^{-\gamma_i z} I_{\infty}^{TE} + N_{\infty}^{TE} e^{-\gamma (2h_i - z)} + N_{\infty}^{TE} e^{-\gamma (z + 2h_i - z_{\infty})} + e^{-\gamma (2d_{\infty} + z_{\infty})} \right\}$$

$$G_{TM}^{(i)xx} = \frac{1}{2 \gamma_i \epsilon_0 \omega \mu_0} \left\{ e^{-\gamma_i z} I_{\infty}^{TM} + \frac{\lambda_0^2}{\mu_0} \tilde{N}_{\infty}^{TM} e^{-\gamma (2h_i - z)} + \tilde{N}_{\infty}^{TM} e^{-\gamma (z + 2h_i - z_{\infty})} + e^{-\gamma (2d_{\infty} + z_{\infty})} \right\}$$

These equations are used to calculate the dyadic Green’s functions for the multilayered medium, allowing for accurate modeling of arbitrarily shaped microstrip elements in multilayered media.
2) Outside Source Layer \((i < j; z > z_s)\):

\[
G_A^{(ij)xx}(r) = \frac{\mu_0 \rho_{ij} e^{j\phi_{ij}}}{2\gamma_{ij}} \hat{M}_{ij}^{TE} \cdot \left[ e^{-j\gamma_{ij}(z_a - z_s)} + \hat{R}_{TE}^{ij} e^{-j\gamma_{ij}(z_a + h_{ij} - h_s)} \right] \\
\cdot \left[ e^{-j\gamma_{ij}(z_s - h_{ij})} + \hat{R}_{TE}^{ij} e^{-j\gamma_{ij}(2h_{ij} - h_s - z_s)} \right] \\
G_q^{(ij)xx}(r) = \frac{1}{2\gamma_{ij} \rho_{ij} \kappa_{ij}^2} \left[ e^{-j\gamma_{ij}(z_a - z_s)} + \hat{R}_{TM}^{ij} e^{-j\gamma_{ij}(z_s + h_{ij} - h_s)} \right] \\
\cdot \left[ e^{-j\gamma_{ij}(z_s - h_{ij})} + \hat{R}_{TM}^{ij} e^{-j\gamma_{ij}(2h_{ij} - h_s - z_s)} \right] \\
\cdot \left[ e^{-j\gamma_{ij}(h_{ij} - 1 - z_a)} - \hat{R}_{TM}^{ij} e^{-j\gamma_{ij}(z_s + h_{ij} - h_s - 1)} \right] \\
\cdot \left[ e^{-j\gamma_{ij}(z_s - h_{ij})} - \hat{R}_{TM}^{ij} e^{-j\gamma_{ij}(2h_{ij} - h_s - z_s - 1)} \right] \tag{5}
\]

where \(\gamma_{ij} = k^2_{ij} - k^2_s\) and \(k^2_{ij} = k^2_s + k^2_{ij}\) with \(k_z\) as the wave number of the \(i\)th layer. The HED is embedded in the \(i\)th layer, and the field point is in the \(j\)th layer. The subscripts \(+\) and \(-\) represent \(+z\) and \(-z\) propagating waves, respectively. The generalized reflection/transmission coefficients of TE/TM waves are denoted as \(\hat{R}_{TE}^{ij}\) and \(\hat{R}_{TM}^{ij}\). They are defined in the Appendix. Equations (3) and (4) are the Green’s functions with the field point embedded in the source layer (in this case, \(i = j\)). They are composed of three parts: a direct term, a single reflection, and a multiple-reflection from both upper and lower interfaces. Equations (5) and (6) are the Green’s function with the field point embedded in the layer \(j\) differing from the source layer \(i\). Because of reciprocity and the continuity of the Green’s function, only the case of an HED below the field point (\(i < j\)) is considered here. Both \(G_A^{(ij)xx}\) and \(G_q^{(ij)xx}\) are the multiplication of the transmission coefficient and +\(z\) propagating components at the \(h_i\) and \(h_{i-1}\) interfaces. With this spectral-domain Green’s function, any configuration of planar conductors in a multilayered structure can be analyzed.

To convert the spectral-domain Green’s function into the spatial domain, Sommerfeld integrals need to be evaluated. This is a very time-consuming step since the integrands are both highly oscillating and slowly decaying. Recently, several papers have suggested using complex images to approximate the remaining integral, after extracting the quasi-static and surface-wave contribution [14], [15], [19], [20]. However, the expansion terms and sampling region are strongly dependent on the layer’s thickness and dielectric constant. Also, it is difficult to find the surface-wave poles for more than three layers. Therefore, in this analysis, Sommerfeld integrals, after extracting the quasi-static terms, are still evaluated by numerical integration [21]–[23]. The quasi-static terms, corresponding to the asymptotic behavior of the integrand, can be obtained by letting \(k_{\rho}\) approach infinity [24]. They can be transformed into analytic expressions by using the identity

\[
e^{-j\gamma_{ij} |\rho|} = \int_{0}^{\infty} e^{-j\gamma_{ij} \rho} J_0(k_{\rho} \rho) k_{\rho} \, dk_{\rho} \tag{7}
\]

where \(R = \sqrt{\rho^2 + z^2}\) is the observation distance. The remaining integrals, which converge very fast, are computed by Gaussian quadrature integration along a contour deformed off the real axis in the \(k_{\rho}\) plane, to avoid surface-wave poles. In order to accelerate the MoM computation, tables of \(G_A\) and \(G_q\) versus distance between source and field points are constructed. These tables are stored as a database and interpolated repeatedly during the MoM procedure. The same Green’s function tables are valid for any new conductor shape, as long as the layer parameters remain the same.

B. The MoM’s

The MoM’s is applied to convert the MPIE to an algebraic linear system. The first step is to expand the unknown current distribution with a set of basis functions. To model the arbitrarily shaped microstrip geometries, the authors adopt the triangular-patch subdomain function, shown in Fig. 2, and defined as

\[
\bar{\sigma}(\rho) = \begin{cases} 
\frac{\ell_n}{2A_n} (\rho^2 - \rho_n^2) & \rho^2 \leq \rho_n^2, \\
\frac{\ell_n}{2A_n} (\rho^2 - \rho_n^2) & \rho_n^2 < \rho^2 \leq \rho^2_T, \\
0 & \rho^2 > \rho^2_T,
\end{cases} 
\tag{8}
\]

This basis function can describe vector current flow. Its divergence is constant over the associated triangular face, and normal components do not exist at the exterior edges. The meshing algorithm developed in [16] is used to grid the whole circuit into small triangular facets. The same current cell is chosen as the testing function (Galerkin’s procedure) and vector identities are used to reformulate (2) as a matrix equation \([\bar{Z}] [\bar{I}] = [V]\), containing the following matrix elements:

\[
Z_{mn} = j\omega \int_{S_m} ds \int_{S_n} dS_d \bar{\sigma}_{n} \cdot \bar{G}_A(\rho) \cdot \bar{G}_A(\rho) \cdot \bar{G}_q(\rho) \\
\cdot [\nabla \cdot \bar{G}_q(\rho)] 
\tag{9}
\]

In this analysis, the excitation mechanism is a series delta-gap voltage source, attached at the open end of the input line [25]. Therefore, only one nonzero element exists in the excitation vector \([V]\), and it is normalized to be 1 V. The output port(s) are open circuited.

Several numerical techniques are applied to accelerate the evaluation of matrix elements [16]. In addition to construction of Green’s function tables (Section II-A), \(Z_{mn}\) is divided into nonsingular and singular parts. The singular part comes from
the singularity of the spatial-domain Green’s function for field points which are very close to the source point. It can be evaluated by four seven-point quadratures combined with the analytic expressions in [16] and [17]. The nonsingular part is approximated by a three-point average. The process of filling the impedance matrix is based on facets instead of interior edges to eliminate many redundant calculations [8], [16]. For example, it takes only 1 min of central processing unit (CPU) time to solve a 625-unknown problem on an IBM RS/6000 workstation.

C. Extraction of Scattering Parameters

After evaluating all elements of \([Z]\) and \([V]\), the linear system is solved for the unknown coefficients \([I]\), and scattering parameters are extracted from the current distribution. Consider the generalized two-port network in Fig. 3, \(I_1\) and \(I_2\) are the total current in ports 1 and 2, respectively, and can be separated into two parts: incident \((A, C)\) and reflected \((B, D)\) components [26]. The current distribution along the microstrip lines can be recovered from the coefficients \([I]\). At two consecutive sampling points \(x_i\) and \(x_{i+1}\) along the port 1 line, the currents can be expressed as

\[
I_i = I_1(x_i) = A e^{i\beta_p x_i} - B e^{i\beta_p x_i} \quad (10)
\]

\[
I_{i+1} = I_1(x_{i+1}) = A e^{i\beta_p x_{i+1}} - B e^{i\beta_p x_{i+1}} \quad (11)
\]

where \(\beta_p\) is the propagation constant along the microstrip line, determined in advance. Therefore, the coefficients \(A\) and \(B\) can be extracted from \(I_i\) and \(I_{i+1}\). For the region far away from the discontinuity and the excitation point, a standing wave will be observed. Therefore, \(A\) and \(B\) are constants, independent of \(x\). The authors can extract their values by a simple curve-fitting algorithm. \(C\) and \(D\) can be extracted from the port 2 current distribution in a similar way. To solve the scattering parameters, there are four unknowns in total, but the authors have only two equations, although more conditions should be included. The simplest way is to recalculate the current distribution with the excitation at port 2 and leave port 1 open. Following the above procedure, \(A^*, B^*, C^*, \) and \(D^*\) can be determined. Finally, the scattering parameters can be found from

\[
S_{11} = \frac{BC^* - B^*C}{AC^* - A^*C} \quad (12)
\]

\[
S_{21} = \frac{DC^* - D^*C}{AC^* - A^*C} \quad (13)
\]

This extraction technique can be generalized easily to multi-port networks, and is demonstrated to be useful and effective by the numerical examples in this paper.

III. NUMERICAL RESULTS AND DISCUSSION

In this section the authors present several numerical results for different configurations: regular/arbitrarily shaped, single/double-layered circuits, and antennas. The multilayered dyadic Green’s function is used to solve these general geometries.

A. Microstrip Open-End Optimization

Several different calibration methods can be used to establish reference planes for measurements of microstrip circuits, some of which require an open-circuit standard. The accuracy of the measurements depends on how precisely the microstrip open is described. Also, the behavior of a true open-circuit may be more closely approximated by modifying the shape of the microstrip open. The simply mitered open end was first discussed in [27] for a coplanar waveguide (CPW) by using a stair-step approximation. The following comparison between numerical and measured results shows that the authors’ analysis method provides the accuracy and flexibility needed to describe both simple (rectangular) and compensated (shaped) microstrip opens.

First the authors consider the simple microstrip open-end discontinuity [see Fig. 4(a)]. A semi-infinite microstrip line is located on the top of a single-layered substrate with \(\varepsilon_r = 9.9\), thickness 0.635 mm, line width 0.6 mm (for 50 \(\Omega\)), and line length 12 mm. Fig. 5 compares the phase of the computed reflection coefficient to the measurement in [28]. The computed results agree with the measurement to within 1.5%. To compensate for the end effect, simply mitered as well as elliptically shaped ends were investigated [see Fig. 4(b)]. The computed performance is shown in Fig. 5. The end effect can be compensated by choosing proper dimensions for either the mitered or elliptical end. To validate this optimization, a simply-mitered microstrip open was fabricated on a 1.5-mm quartz substrate \((\varepsilon_r = 3.47)\). A 0.2-\(\mu\)m nichrome layer was evaporated first to guarantee adhesion on the quartz, followed by a 3-\(\mu\)m copper layer on top [29]. The relevant dimensions of the microstrip lines are shown in Fig. 4(c). The
authors chose a thick substrate with a low dielectric constant to exacerbate the end effect problem, and emphasize optimization of the geometry to minimize this phenomenon. Also, large geometries are less sensitive to fabrication tolerances. Fig. 6 shows excellent agreement between the measured and computed phase of the reflection coefficient. The simply mitered open circuit provides the best approximation to a true open circuit.

B. Microstrip Radial Stubs

The second example demonstrates the authors’ ability to analyze a single-layered microstrip circuit of arbitrary shape. Fig. 7 shows a microstrip shunt-connected radial stub. The substrate has the following parameters: dielectric constant \( \epsilon_r = 9.9 \), thickness \( d = 0.635 \) mm, \( w = 0.6 \) mm, as in Fig. 4(a) and (b). Measurement from [28]. Simply mitered: \( b = 0.25 \) mm; elliptical: \( a = 0.9 \) mm.

Fig. 6. The phase of reflection coefficient for uncompensated and simply-mitered open-end configurations, \( \epsilon_r = 3.37, d = 1.5 \) mm, as in Fig. 4(c).

higher frequencies. The radius of the stub is 7.5 mm and the radial angle is 60°. To verify convergence, computations are presented for two different discretizations: the first one uses 212 triangular facets and 265 internal edges. The second one uses two cells along the transverse direction of the microstrip line and more facets along the radial direction of the stubs, making a total of 488 facets and 644 edges. The computed transmission coefficient is compared with the measurement reported in [30] and the result is shown in Fig. 8. Good
C. Doubled-Layered Bandpass Filter

The next example is a five-section overlap-coupled bandpass filter (see Fig. 9) [31]. It was fabricated from alumina and dudroid substrates with the following tolerances: \( \varepsilon_r = 9.6 \pm 0.2 \) and \( \varepsilon_r = 2.2 \pm 0.02 \). The thicknesses \( d_1 \) and \( d_2 \) are both 0.254 mm. The line widths are \( w_1 = 0.812 \) mm and \( w_2 = 0.458 \) mm. The fabricated resonator and overlap lengths are listed in Table I. Equations (5) and (6) are used here to construct the Green’s function. Fig. 10 is a broad-band plot comparing the computed and measured responses. Good agreement can be seen for the low-side rejection, passband ripple, upper stopband, and second passband. However, there is a small discrepancy in center frequency. Fig. 11 (the dashed line) shows that the measured upper and lower band edges are shifted up by 220 and 130 MHz, respectively. The authors suspect two possible causes: 1) a 0.0076-mm (0.3-mL) air gap was introduced at the alumina/dudroid interface by the 0.0076-mm metallization thickness of the resonators located there; and/or 2) the actual dielectric constants were below their nominal values. Therefore, the filter was analyzed again using the lower bound of the manufacturer’s tolerance for each dielectric constant as well as an additional layer of air (\( d_2 = 0.0076 \) mm): \( \varepsilon_r = 9.6 \), \( \varepsilon_r = 2.2 \), \( d_1 = 0.200 \) mm, \( d_2 = 0.0076 \) mm, and \( d_3 = 0.2514 \) mm. These modifications produced better agreement, as seen in Fig. 11 (the dotted line).

D. Proximity-Coupled Circular-Patch Antenna

To illustrate the versatility of this analysis, a circular-patch antenna is considered as the final example. The configuration is shown in Fig. 12. In this example, the circular-patch antenna is fed by a proximity-coupled microstrip line, embedded in the bottom layer. The two substrate layers have the same parameters: \( \varepsilon_r = \varepsilon_r = 2.2 \), \( d_1 = d_2 = 1.20 \) mm. Three transverse cells are used on the feed line, and 128 triangular facets compose the patch, making a total of 493 internal edges. Fig. 13 shows that the calculated input impedance agrees well with the measurement from [32]. It proves that this full-wave MPIE–MoM analysis using triangular basis functions can be applied to arbitrarily shaped multilayered microstrip antennas as well as circuits.
Fig. 12. Geometry of a proximity-coupled circular patch antenna. 128 facets are used to represent the patch ($\varepsilon_{r1} = \varepsilon_{r2} = 2.62$ and $d_1 = d_2 = 1.59$ mm).

Fig. 13. The input impedance of a proximity-coupled circular patch antenna (See Fig. 12 for dimensions).

IV. CONCLUSION

In advanced MMIC technology, multilayered configurations have become necessary to achieve higher density. The authors have proposed this full-wave spatial-domain analysis to develop a generalized dynamic model for arbitrarily shaped circuits and antennas in a multilayered configuration. The derived dyadic Green’s function can be applied to structures with any number of layers. Moreover, it is developed so that the field point and source point can be at different layers without limitation on the distance between them. In addition, triangular-patch functions allow any arbitrarily shaped planar geometry to be modeled without difficulty. Several numerical results have been presented that show excellent properties and may find promising applications in MMIC designs. This analysis provides an accurate and flexible algorithm to model microstrip antennas as well as microstrip circuits.

APPENDIX

The generalized reflection/transmission coefficients in (3)–(6) were introduced in [18]. They are listed here in our notation as follows:

\[
R_{\text{TE, TM}}^{i+} = \frac{R_{\text{TE, TM}}^{i-} + R_{\text{TE, TM}}^{i+} e^{-2\gamma_{i-} d_{i-1}}}{1 + R_{\text{TE, TM}}^{i-} + R_{\text{TE, TM}}^{i+} e^{-2\gamma_{i+} d_{i+1}}},
\]

\[
R_{\text{TE, TM}}^{i-} = \frac{R_{\text{TE, TM}}^{i-} + R_{\text{TE, TM}}^{i+} e^{-2\gamma_{i+} d_{i+1}}}{1 + R_{\text{TE, TM}}^{i-} + R_{\text{TE, TM}}^{i+} e^{-2\gamma_{i+} d_{i+1}}},
\]

\[
S_{k+}^{\text{TE, TM}} = \left[1 - R_{\text{TE, TM}}^{i+} R_{\text{TE, TM}}^{i-} e^{-2\gamma_{i-} d_{i-1}}\right]^{-1}
\]

\[
T_{k+}^{\text{TE, TM}} = \prod_{k=j}^{i+1} e^{-\gamma_{ik} d_{k-1}} s_{k+}^{\text{TE, TM}} \quad \text{with } d_j = 0
\]

where $S_{k+}^{\text{TE, TM}}$ is

\[
s_{k+}^{\text{TE, TM}} = \frac{1 + S_{\text{TE, TM}}^{\text{TE, TM}}}{1 - R_{\text{TE, TM}}^{i+} R_{\text{TE, TM}}^{i+} e^{-2\gamma_{i+} d_{i+1}}}
\]

with half space with ground plane otherwise.

REFERENCES


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