Evaluation of the Radiation of Rectangular Microstrip Antenna Covered with a Dielectric Layer

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Abstract: In this work, a rigorous full-wave analysis is applied to investigate the effect of dielectric superstrate on radiation pattern of rectangular microstrip patch antenna. Using a matrix representation of each layer, the far-field pattern of the substrate-superstrate configuration is efficiently determined by the (TM,TE) representation. Numerical results indicate splitting of the mainlobe into three lobes with an increase in superstrate thickness.

Key words: Dielectric cover, printed circuit antennas, radiation, spectral-domain.

INTRODUCTION

Microstrip antennas have been employed in airborne and spacecraft systems [Bahl 1982] due to their excellent advantages. Such advantages include small size, light weight, low production cost, conformal nature, and good aerodynamic characteristics [Dmitriev 02]. Superstrate dielectric layers are often used to protect printed circuit antennas from environmental hazards, or may be naturally formed (e.g. ice layers) during flight or severe weather conditions. Theoretical research on the effect of dielectric superstrate on the resonant frequency, half-power bandwidth and quality factor of a microstrip patch antenna are abundant [Bahl 1982] [Row 1993] [Losada 1999] [Bouttout 00], however, there is no theoretical report on the effect of superstrate on the radiation pattern of a microstrip patch antenna.

In this paper, a theoretical investigation of the effect of dielectric superstrate on radiation pattern of rectangular microstrip patch antenna is presented. This paper is organized as follows. In section 1, using a matrix representation of each layer, a new approach to derive the far-field pattern of the substrate-superstrate configuration is proposed. Numerical results for the effect of dielectric superstrate on radiation pattern of rectangular microstrip patch antenna are presented in section 2. Finally, conclusions are summarized in section 3.
with constitutive parameters $\mu_0$ and $\varepsilon_0$. All fields and currents are time harmonic with the $e^{i\omega t}$ time dependence suppressed. The transverse fields inside the $j$th layer ($j=1,2$) can be obtained via the inverse vector Fourier transforms as [Chew 1986]

$$
\mathbf{E}(r,z) = \left[ \begin{array}{c} E_x(r,z) \\ E_y(r,z) \end{array} \right] = \frac{1}{4\pi} \int_{\infty}^{\infty} \int_{\infty}^{\infty} \tilde{\mathbf{F}}(k_x,r_x) \cdot \mathbf{e}(k_x,z) \, dk_x \, dk_y
$$

(1)

$$
\mathbf{H}(r,z) = \left[ \begin{array}{c} H_x(r,z) \\ -H_y(r,z) \end{array} \right] = \frac{1}{4\pi} \int_{\infty}^{\infty} \int_{\infty}^{\infty} \tilde{\mathbf{F}}(k_x,r_x) \cdot \mathbf{h}(k_x,z) \, dk_x \, dk_y
$$

(2)

where $\tilde{\mathbf{F}}(k_x,r_x)$ is the kernel of the vector Fourier transform [Chew 1986], and

$$
\mathbf{e}(k_x,z) = \mathbf{A}_j(k_x) e^{-ik_j z} + \mathbf{B}_j(k_x) e^{ik_j z}
$$

(3)

$$
\mathbf{h}(k_x,z) = \tilde{\mathbf{G}}_j(k_x) \left[ \begin{array}{c} \mathbf{A}_j(k_x) e^{-ik_j z} - \mathbf{B}_j(k_x) e^{ik_j z} \end{array} \right]
$$

(4)

In (3) and (4), $\mathbf{A}_j$ and $\mathbf{B}_j$ are two-component unknown vectors and

$$
\tilde{\mathbf{G}}_j(k_x) = \text{diag} \left[ \frac{\omega \varepsilon_0 \varphi_{j0}}{k_j}, \frac{k_j}{\omega \mu_0} \right], \quad k_j^2 = \left( \frac{\varepsilon_j}{\varepsilon_0} k_0^2 - k_j^2 \right)^2
$$

(5)

with $k_0^2 = \omega^2 \varepsilon_0 \mu_0$ and $k_j$ is the propagation constant in the $j$th layer. Writing (3) and (4) in the planes $z=z_{j-1}$ and $z=z_j$, and by eliminating the unknowns $\mathbf{A}_j$ and $\mathbf{B}_j$, we obtain the matrix form

$$
\begin{bmatrix}
\mathbf{e}(k_x,z_j^-) \\
\mathbf{h}(k_x,z_j^-)
\end{bmatrix} = \mathbf{T}_j \cdot
\begin{bmatrix}
\mathbf{e}(k_x,z_{j-1}^+), \\
\mathbf{h}(k_x,z_{j-1}^+)
\end{bmatrix}
$$

(6)

with

$$
\mathbf{T}_j = \begin{bmatrix}
\mathbf{T}_{j11} & \mathbf{T}_{j12} \\
\mathbf{T}_{j21} & \mathbf{T}_{j22}
\end{bmatrix} = \begin{bmatrix}
\mathbf{I} \cos \theta_j & -i \mathbf{I}^{-1} \sin \theta_j \\
-i \mathbf{I} \sin \theta_j & \mathbf{I} \cos \theta_j
\end{bmatrix}
$$

(7)

which combines $\mathbf{e}$ and $\mathbf{h}$ on both sides of the $j$th layer as input and output quantities. In (7), $\theta_j = k_j d_j$ and $\mathbf{I}$ stands for the 2x2 unit matrix. The matrix $\mathbf{T}_j$ is the matrix representation of the $j$th layer in the (TM,TE) representation. The continuity equations for the tangential field components are

$$
\mathbf{e}(k_x,z_j^-) = \mathbf{e}(k_x,z_j^+), \quad j = 1, 2
$$

(8)

$$
\mathbf{h}(k_x,z_j^-) - \mathbf{h}(k_x,z_j^+) = \delta_{j1} \mathbf{j}(k_x), \quad j = 1, 2
$$

(9)

$\mathbf{j}(k_x)$ is the vector Fourier transform of the current $\mathbf{J}(r_x)$ on the patch, it accounts for the discontinuity of the transverse magnetic field at the interface $z=d_1$. $\delta_{j1}$ is the Kronecker symbol. Using (7)-(9) yields

$$
\begin{bmatrix}
\mathbf{e}(k_x,d_1^+) \\
\mathbf{h}(k_x,d_1^+)
\end{bmatrix} = \mathbf{T}_1 \cdot
\begin{bmatrix}
\mathbf{e}(k_x,0^+) \\
\mathbf{h}(k_x,0^+)
\end{bmatrix}
$$

(10)

$$
\begin{bmatrix}
\mathbf{e}(k_x,d_1^+) \\
\mathbf{h}(k_x,d_1^+)
\end{bmatrix} = \mathbf{T}_2 \cdot
\begin{bmatrix}
\mathbf{e}(k_x,d_1^+) \\
\mathbf{h}(k_x,d_1^+)
\end{bmatrix}
$$

(11)

The transverse electric field must necessarily be zero on a perfect conductor, so that for the perfectly conducting ground plane we have

$$
\mathbf{e}(k_x,0^-) = \mathbf{e}(k_x,0^+) = \mathbf{e}(k_x,0) = 0
$$

(12)

In the unbounded air region above the superstrate of the structure ($d_1 + d_2 \gg \infty$ and $\varepsilon_c = 1$) the electromagnetic field given by (3) and (4) should vanish at $z \rightarrow \infty$ according to Sommerfeld’s condition of radiation, this yields

$$
\mathbf{h}(k_x,d_1^+) = \tilde{\mathbf{G}}_0(k_x) \cdot \mathbf{e}(k_x,d_1^+)
$$

(13)

where $\tilde{\mathbf{G}}_0(k_x)$ can be easily obtained from the expression of $\tilde{\mathbf{G}}_j(k_x)$ given in (5) by allowing $\varepsilon_j = 1$. From (10)-(13), we obtain the following relationship:

$$
\mathbf{e}(k_x,d_1) = \tilde{\mathbf{G}}(k_x) \cdot \mathbf{j}(k_x)
$$

(14)

where $\tilde{\mathbf{G}}(k_x)$ is the dyadic Green’s function in the (TM,TE) representation, it is given by

$$
\tilde{\mathbf{G}}(k_x) = \left[ \begin{array}{c} \mathbf{T}_{22} \cdot \mathbf{T}_{12} \cdot \mathbf{T}_{21} \cdot \mathbf{I} \cdot \mathbf{G}_0 \cdot \mathbf{T}_{12} \cdot \mathbf{T}_{21} \cdot \mathbf{T}_{22} \end{array} \right]^{-1}
$$

(15)

Now that we have the necessary Green’s function, it is relatively straightforward to formulate the moment method solution for the complex resonant frequency of the antenna [Boutout 00]. Once the complex resonant frequency is determined, the eigenvector corresponding to the minimal eigenvalue of the impedance matrix gives the coefficients of the current on the rectangular patch. The current density is thus obtained in numerical form. This current density can be used for computing the radiation electric field in the region $z \geq d_1 + d_2$ of Figure 1. Using the stationary phase method [Colli 1985], we can obtain the far-zone field above the superstrate of Figure 1 in terms of the transverse electric field at the plane

- 2 -
$z = d_1 + d_2$ as follows:

$$
\begin{bmatrix}
E_p(r', \theta', \phi') \\
E_\phi(r', \theta', \phi')
\end{bmatrix} = i k_0 \frac{e^{-i k r}}{2 \pi r} \begin{bmatrix}
-1 & 0 \\
0 & \cos \theta'
\end{bmatrix} \cdot e(k_x, d_1 + d_2)
$$

where $\{r', \theta', \phi'\}$ is a local set of spherical coordinates defined with respect to the Cartesian system $\{x', y', z'\}$ with an origin placed at the plane $z = d_1 + d_2$ of Figure 1. From equations (11) and (13), we can obtain the transverse electric field $e(k_x, d_1 + d_2)$ at the plane $z = d_1 + d_2$ in terms of the transverse electric field $e(k_x, d_1)$ at the plane $z = d_1$ as follows:

$$
e(k_x, d_1 + d_2) = \left[ \mathbf{T}_{22}^{12} - \mathbf{g}_0, \mathbf{T}_{12}^{12} \right]^{-1} \cdot e(k_x, d_1)
$$

(17)

Using (14), (15) and (17), equation (16) becomes

$$
\begin{bmatrix}
E_p(r', \theta', \phi') \\
E_\phi(r', \theta', \phi')
\end{bmatrix} = i k_0 \frac{e^{-i k r}}{2 \pi r} \begin{bmatrix}
-1 & 0 \\
0 & \cos \theta'
\end{bmatrix} \cdot \mathbf{T}_{12}^{12} \cdot \left[ \mathbf{f}_{22}^{22} - \mathbf{g}_0, \mathbf{f}_{12}^{12} \right]^{-1} \cdot j(k_x)
$$

(18)

where

$$\mathbf{T} = \mathbf{T}_{22}, \mathbf{T}_{12}
$$

(19)

In equations (16) and (18), $k_x$ and $k_y$ are evaluated at the stationary phase point as

$$k_x = -k_0 \sin \theta' \cos \phi'
$$

(20)

$$k_y = -k_0 \sin \theta' \sin \phi'
$$

(21)

Note that the radiation field given by the expression (18) is expressed in the (TM,TE) representation, which is not the case in the other formulations [Losada 1999] [Pozar 1987]. Contrary to the equivalent boundary method, where the radiation pattern is calculated via recurrent expressions [Losada 1999], the new mathematical expression shown in equation (18) allows the computation of the radiation field of the substrate-superstrate configuration using simple matrix multiplications. It is worth noting that equation (18) can be generalized to the case of a multilayered structure (arbitrary number of layers below and/or above the patch) by cascading the matrices by simple multiplication.

2. Results and discussion

In order to confirm the computation accuracy, our calculated resonant frequencies are compared with previously published experimental data [Bahl 1982]. Table 1 summarizes the measured and computed resonant frequencies for different superstrate materials and differences between numerical and experimental results of less than 1.17% are obtained. As a consequence, excellent agreement between theory and experiment is achieved. This validates the theory presented in this work. We have also obtained the current distribution for the microstrip structures considered in table 1, when the Custom High-K is used as a superstrate. Figure 2 shows the current distribution for two different superstrate thicknesses. Note that the current plotted in Figure 1 is the current in the y direction; since the considered mode in this work is the TM$_{01}$ mode, which has a dominant current in the y direction.

Table 1. Comparison of measured and calculated resonant frequencies; $a = 1.9\text{ cm}$, $b = 2.29\text{ cm}$, $\varepsilon_r = 2.32$, $d_1 = 1.59\text{ mm}$.

<table>
<thead>
<tr>
<th>Superstrate</th>
<th>$\varepsilon_r$</th>
<th>$d_2$ (mm)</th>
<th>Measured [Bahl 1982]</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air</td>
<td>1</td>
<td>$\infty$</td>
<td>4.104</td>
<td>4.123</td>
</tr>
<tr>
<td>Duroid</td>
<td>2.32</td>
<td>0.8</td>
<td>4.008</td>
<td>4.033</td>
</tr>
<tr>
<td>Plexiglass</td>
<td>2.6</td>
<td>3.18</td>
<td>3.874</td>
<td>3.887</td>
</tr>
<tr>
<td>Mylar</td>
<td>3</td>
<td>0.064</td>
<td>4.070</td>
<td>4.108</td>
</tr>
<tr>
<td>Custom High-K</td>
<td>10</td>
<td>1.54</td>
<td>3.482</td>
<td>3.518</td>
</tr>
<tr>
<td>High-K</td>
<td>1.54</td>
<td>3.482</td>
<td>3.518</td>
<td></td>
</tr>
</tbody>
</table>

Figure 2. Current distribution for the microstrip structures considered in table 1; when the Custom High-K is used as a superstrate.
Figure 3. Behaviour of radiation pattern of rectangular microstrip antenna with increase in thickness of superstrate.

Table 2. Rectangular microstrip patch antenna parameters with in-touch protecting Perspex superstrate.

<table>
<thead>
<tr>
<th>Superstrate thickness (mm)</th>
<th>Resonant frequency (GHz)</th>
<th>Number of lobes</th>
<th>Angle of lobes (degrees)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>9.777</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>2.54</td>
<td>9.410</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>5.30</td>
<td>9.452</td>
<td>3</td>
<td>0, ± 77.25</td>
</tr>
<tr>
<td>5.80</td>
<td>9.481</td>
<td>3</td>
<td>0, ± 82.25</td>
</tr>
</tbody>
</table>

The effect of dielectric superstrate on radiation pattern of an X-band rectangular microstrip patch antenna is studied in Figure 3. The rectangular microstrip patch of length \( a = 7.95 \) mm and width \( b = 4.58 \) mm is printed on alumina substrate \( \varepsilon_r = 9.8(1 - i 10^{-4}) \), \( d_1 = 0.635 \) mm, and covered by a Perspex superstrate with relative permittivity \( \varepsilon_r = 2.6 \) with various thickness. The radiation of the TM_{01} mode in the H plane is considered. The Figure 3 shows the radiation patterns for antennas with and without superstrate. We observe that the radiation pattern with \( d_2 = 2.54\)mm have the same shape that of the structure without superstrate. When the patch is covered by the thick superstrate \( d_2 = 5.30 \) mm \( (d_2 = 5.80 \) mm\), the radiation pattern split into three lobes having the centre lobe higher (lower) than for the other two lobes. In table 2, we give more details concerning the number of lobes and their angular positions for each thickness of the superstrate. Concerning the influence of the dielectric superstrate on the operating frequency of the rectangular microstrip antenna, although this effect has been considered in numerous work [Bahl 1982] [Bouttout 00] [Losada 1999] [Row 1993], which is important to note here is that for thick superstrate the resonant frequency increases with the increase of the thickness of the dielectric cover.

3. Conclusion

In this paper, a rigorous full-wave analysis has been applied to investigate the effect of dielectric superstrate on radiation pattern of rectangular microstrip patch antenna. Using a matrix representation of each layer, a new explicit expression has been derived for the calculation of the far-field pattern of the substrate-superstrate configuration. We have shown that this expression can be easily extended for the case of a multilayered structure. The numerical results obtained have shown that using thick superstrate as a protecting layer will drastically change the shape of radiation pattern into three lobes. Higher thickness of superstrate can be used to split the mainlobe of a single microstrip patch antenna into three lobes for special purposes.

REFERENCES