Considerations for Millimeter Wave Printed Antennas

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Abstract—Calculated data are presented on the performance of printed antenna elements on substrates which may be electrically thick, as would be the case for printed antennas at millimeter wave frequencies. Printed dipoles and microstrip patch antennas on polytetrafluoroethylene (PTFE), quartz, and gallium arsenide substrates are considered. Data are given for resonant length, resonant resistance, bandwidth, loss due to surface waves, loss due to dielectric heating, and mutual coupling. Also presented is an optimization procedure for maximizing or minimizing power launched into surface waves from a multielement printed antenna array. The data are calculated by a moment method solution.

I. INTRODUCTION

There has been rapid growth in printed antenna theory and technology during the last decade [1], [2]. Most of this work was for antennas operating in the UHF to microwave frequency bands (300 MHz to 10 GHz), and characteristics of printed antennas such as low-cost, low-profile, conformability, and ease of manufacture were often found to outweigh the electrical disadvantages, such as narrow bandwidth and low-power capacity, for many applications.

Currently there is increasing interest in millimeter wave systems and applications, such as aircraft-to-satellite communications and imaging array antennas [3], [4], as well as interest in complete monolithic systems which combine antenna elements or arrays on the same substrate as the integrated RF/IF front-end detector and amplifier circuits. Thus, printed antennas are being seriously considered for use at frequencies well above 30 GHz, while past applications were generally below 30 GHz. In these applications, substrates are often much thicker and have higher dielectric constants than at lower frequencies. It is the purpose of this paper to present performance data for and discuss the applications of printed antennas on these types of substrates.

As will be seen, the electrical performance of these antennas can be severely degraded, due to surface waves or mutual coupling. Bandwidth and input impedance are additional properties that are strongly affected by substrate thickness, often in a desirable way. Other factors such as dielectric loss and feeding techniques can also be significantly different at millimeter wave frequencies.

This paper will consider two particular types of printed antennas: rectangular microstrip (patch) elements, and printed dipoles. The intrinsic differences between these two elements and their comparative electrical performances will be discussed. Three types of substrate materials will be considered: polytetrafluoroethylene (PTFE), quartz, and gallium arsenide. These materials represent substrates which may be used for millimeter wave antennas, and cover the range of relative permittivity from 2.55 to 12.8. The following characteristics will be presented in Section II.

1) General element characteristics.
2) Substrate properties.
3) Resonant element length.
4) Resonant input resistance.
5) Bandwidth.
6) Losses due to surface waves.
7) Losses due to dielectric.

Section III will discuss mutual coupling and surface wave effects in an array environment and present a procedure for the optimization of array efficiency (minimization or maximization of $e = P_{\text{rad}}/(P_{\text{rad}} + P_{\text{sw}})$, where $P_{\text{rad}}$ is the desired radiated power and $P_{\text{sw}}$ is surface wave power). Although the results presented here are for printed dipole or microstrip patch elements, some of the trends and conclusions will apply to other types of printed antenna elements.

All of the data presented in this paper were calculated using a moment method solution of a printed rectangular radiating element on a grounded dielectric slab. A detailed description of this method, with comparisons of calculated and measured results for input and mutual impedances, has already appeared in the literature [5], so only a brief description will be given here.

Important factors for antennas on electrically thick substrates include surface waves and mutual coupling [6]. The lowest order surface wave ($TM_0$) has a zero cutoff frequency, and thus is excited to some degree even on very thin substrates. As the substrate becomes thicker, more surface wave modes can exist, and more power can be coupled into these waves. Mutual coupling between elements in arrays involves the transfer of power from one element to a nearby element via space waves (direct radiation) or by surface waves. Coupling levels greater than roughly 20-30 dB may have a deleterious effect on array performance, unless specifically included in the design procedure. Thus, it is desirable for the theoretical solution to account for fields exterior to the radiating element, i.e., to account for surface wave power and mutual coupling. The cavity model and transmission line model cannot do this. In addition, neither of these models have yet successfully treated the printed dipole element, and neither are valid for antennas on thick substrates.

The moment method solution uses the rigorous dyadic Green's function for the grounded dielectric slab, and so includes the exterior fields making calculations for surface wave excitation and mutual coupling possible. Because of the general nature of the moment method formulation, printed dipoles as well as probe-fed or microstrip line-fed patches can be handled. Dielectric loss can be easily included by using a complex permittivity in the solution. The price paid for this versatility is a somewhat more complicated solution, primarily due to the Sommerfeld-

II. PRINTED DIPOLE AND MICROSTRIP PATCH ANTENNA ELEMENTS

This section will discuss properties of printed dipoles and patch antenna elements, and present data on their electrical characteristics.

A. General Element Characteristics

Printed dipole radiating elements have been extensively studied by Alexopoulos et al. [8], [9], [10], via a similar moment method procedure. Fig. 1 shows a typical center-fed printed dipole element. In practice, the feed may take the form of a parallel two-wire line printed on the substrate. This feed line may carry RF energy or, if a detector is placed at the dipole gap, IF energy. In either case, this type of feed is balanced with respect to the ground plane, which can be a serious disadvantage in some applications. The use of parallel microstrip feed lines to couple the radiating dipole, as in [7], can alleviate this difficulty at the expense of a more complicated feeding structure, possibly involving printed conductors on two substrate levels. Advantages of the printed dipole are that it uses less substrate area compared to patch elements (particularly important in arrays), and that it can be used near its first or second resonances without deleterious higher order mode effects.

The microstrip patch antenna, also shown in Fig. 1, however, is inherently unbalanced with respect to the ground plane. The patch can be fed with a microstrip line, or with a probe from the bottom of the substrate. An unbalanced antenna element is probably advantageous when RF or IF circuitry is to be combined with the antenna in a hybrid or monolithic configuration. Some disadvantages are that the rectangular patch uses more substrate area than the dipole, and that a probe-type feed may be difficult to fabricate on monolithic substrates, or even on quartz substrates. A problem also exists with microstrip line feeds, since a microstrip line's characteristic impedance determines the feed line width, and is relatively constant with frequency. The size of a resonant patch antenna, however, decreases with increasing frequency, so that a given microstrip feed line on a substrate thickness d for \( \varepsilon_r = 2.55, W = 0.3 \lambda_0 \) for the patch. Also shown are the printed dipole and patch geometries.

![Fig. 1. Resonant lengths of a printed dipole and a microstrip patch versus d for \( \varepsilon_r = 2.55, W = 0.3 \lambda_0 \) for the patch. Also shown are the printed dipole and patch geometries.](image)

TABLE I

<table>
<thead>
<tr>
<th>Substrate</th>
<th>( \varepsilon_r )</th>
<th>( \tan \delta )</th>
</tr>
</thead>
<tbody>
<tr>
<td>PTFE</td>
<td>2.55</td>
<td>0.001-0.003</td>
</tr>
<tr>
<td>Quartz</td>
<td>3.78</td>
<td>0.001</td>
</tr>
<tr>
<td>GaAs</td>
<td>12.8</td>
<td>0.002</td>
</tr>
</tbody>
</table>

TABLE I ELECTRICAL PROPERTIES OF MICROWAVE SUBSTRATES

For a dipole, the feed couples to the electric field component along the dipole axis, while a coax or microstrip line-fed patch antenna is coupled by the electric field component normal to the substrate. Printed dipoles and patches, however, have similar current distributions, thus the radiation patterns are similar [11], [8].

B. Substrate Characteristics

Three substrate materials—PTFE, quartz, and gallium arsenide—were chosen for comparison in this paper. This choice was based on the fact that the permittivities range from 2.55 to 12.8, and that these materials are either in use today or are expected to be used for millimeter wave antenna systems. Because of time and space considerations, not all materials available today could be compared here, but it is felt that other typical substrate materials have properties roughly in range of those considered. Table I summarizes the typical electrical properties of these three substrates.

Polytetrafluoroethylene and related products like Rexolite and Duroid have been used extensively in the microwave band. Quartz substrates have very good dimensional stability and are often used in microwave integrated circuits. Gallium arsenide is probably the preferred substrate material for monolithic microwave integrated circuits.

C. Resonant Frequency

Clearly one of the first considerations in the design of a printed antenna element is the length L of the element required for resonance. This length is a function of substrate thickness d and dielectric constant \( \varepsilon_r \), and, in the case of a microstrip patch, a function of the patch width W. Because the dielectric fills only part of the region surrounding the antenna element, the resonant length does not scale with dielectric constant as \( L = \varepsilon_r \), as an antenna in a homogeneous medium would.

Fig. 1 shows the required lengths for the first resonance of a printed dipole and a rectangular microstrip patch element versus substrate thickness d for a PTFE material. The patch width is \( W = 0.3 \lambda_0 (\lambda_0 \) is the free-space wavelength). The dipole length varies less than 6 percent for \( 0 < d < 0.5 \lambda_0 \) and is slightly longer than the patch length, which varies somewhat more with
d. An interesting feature of the patch antenna is that it stops resonating for substrate thicknesses greater than 0.11 \( \lambda_0 \). With increasing substrate thickness, the trend is to an entirely inductive input impedance locus. This effect occurs for both probe-type and microstrip line-type feeds. This situation is probably undesirable in most cases, and so the use of patches on thick substrates may not be practical unless some way of countering this inductive trend, by using a capacitive-gap coupling from a feed line, for example, is used. The dotted continuation of the patch length curve in Fig. 1 is given only to show the length chosen for the calculation of other data presented in Figs. 4, 6, and 8. An increase in patch width \( W \) can reduce the resonant length by a few percent; the length reduction is greater for thicker substrates.

Fig. 2 shows the resonant lengths for half- and full-wave printed dipoles on a quartz (\( \epsilon_r = 3.78 \)) substrate. (The usefulness of the full-wave dipole is discussed in Section II-D.) Note that the length for the first resonance has decreased from the corresponding value for PTFE, but not by the factor \( \sqrt{2.55/3.78} \), which would be the case if scaling could be applied with dielectric constant.

Fig. 3 shows the required lengths for half-wave and full-wave printed dipoles and a microstrip patch element on GaAs. Again, the patch element does not resonate for substrate thicknesses greater than 0.08 \( \lambda_0 \).

D. Resonant Resistance

Fig. 4 shows the input resistance of a half-wave printed dipole and a microstrip patch on a PTFE substrate versus thickness. As previously pointed out, the patch element does not strictly resonate for \( d > 0.11 \lambda_0 \); the patch resistance shown in Fig. 4 for \( d > 0.11 \lambda_0 \) is the real part of the input impedance for a patch length of 0.270 \( \lambda_0 \). Since the printed dipole's first resonance is a series-type resonance, the input resistance is very small for small \( d \), since electrically thin substrates imply high-Q resonances. The microstrip patch, having a parallel-type resonance, shows a high input resistance for small \( d \).

Fig. 5 shows the resonant resistance of half-wave and full-wave printed dipoles and a microstrip patch on a GaAs substrate. Again, the patch resistance for \( d > 0.08 \lambda_0 \) is taken as the real part of the input impedance for \( L = 0.105 \lambda_0 \).

The full-wave dipole has a parallel-type resonance, with high input resistance for small \( d \), similar to a full-wave dipole in free space. This element has interesting advantages in some applica-
E. Bandwidth

Bandwidth is defined here as the half-power width of the equivalent circuit impedance response. For a series-type resonance, as shown in [10], this bandwidth (BW) is

\[
BW = \frac{2R}{\frac{dX}{d\omega}}_{|\omega=\omega_0},
\]

where \( Z = R + jX \) is the input impedance at the resonant frequency \( \omega_0 \). For a parallel-type resonance (1) is used with \( R \) replaced by \( G \) and \( X \) replaced by \( B \), where \( Y = G + jB \) is the input admittance at resonance. This definition of bandwidth implies a standing wave ratio of about 2.4, for a transmission line of characteristic impedance \( R \) or \( 1/G \), \( \Omega \). The derivative in (1) can be evaluated by calculating the input impedance at two frequencies near resonance and using a finite difference approximation.

Fig. 6 shows the calculated bandwidths of a half-wave printed dipole and a microstrip patch versus substrate thickness for a PTFE substrate. The patch width is 0.3 \( \lambda_0 \) and is fed by a probe at a point \( L/4 \) from the patch edge, although the feeding method or position does not affect the intrinsic patch bandwidth. The bandwidth increases rapidly with increasing substrate thickness, so that bandwidths of 10-20 percent can be obtained for substrate thicknesses in the range of \( d = 0.1 \lambda_0 \) to 0.2 \( \lambda_0 \). Also note that the bandwidth of a patch is significantly greater than that of a printed dipole, at least over the range for which the patch actually resonates (\( d < 0.11 \lambda_0 \)). These facts are consistent with the antenna gain/bandwidth relation to antenna size, as discussed by Harrington [11]. The lowest achievable \( Q \) of an antenna is inversely related to antenna volume; since the patch antenna encompasses a greater volume than does the printed dipole, its \( Q \) can be lower than the \( Q \) of the dipole, hence the bandwidth can be greater. Also shown in Fig. 6 is a bandwidth calculation for the patch using the cavity model [1]. The cavity model approximation is seen to be useful for substrate thicknesses \( d < 0.04 \lambda_0 \).

Fig. 7 shows the bandwidth of a printed dipole (half-wave) and a microstrip patch on a gallium arsenide substrate, versus \( d \). Again the patch bandwidth is greater than the dipole bandwidth.

F. Power Lost to Surface Waves

Both TE and TM surface waves can be excited on a grounded dielectric substrate. The cutoff frequency of these modes is given by [11].

\[
f_c = \frac{nc}{4d\sqrt{\varepsilon_r} - 1},
\]

where \( c \) is the speed of light, and \( n = 0, 1, 2, 3, \ldots \) for the TM\(_0\), TE\(_1\), TM\(_2\), TE\(_3\) \ldots surface mode. Note that the TM\(_0\) mode has a zero cutoff frequency, so that it can be generated for any substrate thickness \( d \). As the substrate becomes electrically thicker, more surface modes can exist and the coupling to the lower order modes can become stronger. For thin substrates \( d < 0.01 \lambda_0 \), surface wave excitation is generally not important. For thicker substrates surface waves may have a detrimental effect on printed antenna performance. Surface wave power launched in an infinitely wide substrate would not contribute to the main beam radiation and so can be treated as a loss mechanism. A radiation efficiency can then be defined as

\[
e = \frac{P_{rad}}{P_{rad} + P_{sw}},
\]

where \( P_{rad} \) is the power radiated and \( P_{sw} \) is the power lost to surface wave modes.
where $P_{rad}$ is the power radiated via space wave (direct main beam power), and $P_{sw}$ is the power coupled into surface waves. $P_{rad} + P_{sw}$ is then the total power delivered to the printed antenna element. Dielectric loss is ignored here. The effect of a finite-sized substrate would be to diffract the surface waves from the substrate edges, possibly causing undesirable effects on sidelobe level, polarization, or main beam shape. Surface waves could also be diffracted by or coupled to feed lines or components on the substrate.

In the moment method formulation surface wave fields and space wave fields are easily separated from the Sommerfeld-type integral expression for the total fields of an elemental current source on a grounded dielectric slab—the surface waves come from the residues of the contour integral. Since the moment method impedance matrix elements are expressed in terms of integrals of the fields from the expansion modes, these elements can be broken up as

$$Z_{mn} = Z_{mn}^{rad} + Z_{mn}^{sw},$$  \hspace{1cm} (4)

where $Z_{mn}$ represents the matrix element using the total field and $Z_{mn}^{rad}$ and $Z_{mn}^{sw}$ represent the direct radiation (space wave) and surface wave contributions, respectively. Then, if $I_n$ represents the current on the $n$th expansion mode, the total input power can be written as

$$P_{tot} = \text{Re} \sum_n \sum_m I_n^* Z_{mn} I_m.$$  \hspace{1cm} (5)

and the radiated power can be written as

$$P_{rad} = \text{Re} \sum_n \sum_m I_n^* Z_{mn}^{rad} I_m.$$  \hspace{1cm} (6)

Fig. 8 shows efficiency (3) versus substrate thickness $d$ for a half-wave printed dipole and a microstrip patch, for $\varepsilon_r = 2.55$. Observe that $e \rightarrow 1.0$ as $d \rightarrow 0$, since surface wave excitation is negligible for very thin substrates. As $d$ gets larger the TM$_0$ surface mode becomes stronger, reducing $e$. However, the radiated power becomes greater as $d$ increases, so that $e$ levels off and starts to increase for $d > 0.1 \lambda_0$. At $d = 0.2 \lambda_0$, the next surface mode (TE$_1$) starts to propagate, causing a slope discontinuity in $e$ and a decrease in $e$ as this mode becomes more strongly coupled. This type of slope discontinuity is also seen in related dielectric covered antenna problems [12].

An interesting feature of Fig. 8 is the similarity between $e$ for the dipole and the patch. Also, $e$ does not depend on the feed location of the dipole or patch, or on the patch width $W$.

Fig. 9 shows the efficiency $e$ for half-wave and full-wave dipoles and a microstrip patch on GaAs substrate. Note that surface wave power accounts for over half of the total input power for $d > 0.045 \lambda_0$.

G. Losses Due to Dielectric

Power loss due to dielectric heating can be calculated by using the loss tangent and complex permittivity for the particular dielectric material. For the half-wave dipole (a series-type reso-
nance), for example, the radiation efficiency based on dielectric loss can be calculated as
\[
\eta = \frac{R_r}{R_r + R_l}
\]
where \(R_r\) is the radiation resistance at the input terminals and \(R_l\) is the loss resistance. \(R_r\) and \(R_l\) can be found from two calculations of input impedance; one with \(\tan \delta = 0\), and one with \(\tan \delta \neq 0\). The radiation resistance is \(R_r = \text{Re}(Z_{in})\) for \(\tan \delta = 0\), and the loss resistance is found from \(R_l + R_l = \text{Re}(Z_{in})\) with \(\tan \delta \neq 0\). This is an accurate procedure for small losses. For full-wave dipoles or microstrip patches (anti-resonances), the efficiency is calculated using conductances in (7). Note that efficiency as defined by (7) does not include power loss to surface waves (although it does include heating loss from surface wave fields).

Efficiency based on dielectric loss for a half-wave dipole and a microstrip patch versus \(d\) is shown in Fig. 10 for \(\varepsilon_r = 2.55\). The lengths of the dipole and patch are chosen for resonance, and the patch width is again \(W = 0.3 \lambda_0\) for the patch.

It is seen that the patch efficiency is greater than the dipole efficiency, and that efficiency improves rapidly as substrate thickness increases. Both of these effects can be explained by noting that, for a given power level, the fields are more concentrated for thin substrates or narrow antenna elements, thus more power is lost to dielectric heating than in cases of thicker substrates or wider elements.

III. PRINTED ANTENNA ELEMENTS IN AN ARRAY ENVIRONMENT

This section will discuss some aspects of the printed antenna element in an array configuration—in particular, mutual coupling between array elements and array effects on surface wave power.

A. Mutual Coupling

When printed antenna elements are in an array environment, large mutual coupling levels can degrade sidelobe levels, main beam shape, and possibly cause array blindness. Foreknowledge of the coupling between array elements and proper inclusion into the array design procedure can minimize these effects [7], [13].

The calculation of mutual coupling as a two-port transfer impedance by the moment method is described in [5]. The method yields magnitudes as well as phase, and comparisons with measured data for patches are shown in [5]. Mutual coupling between parallel and collinear half-wave dipoles versus separation is shown in [9]. Calculations for full-wave dipoles have also been made, with the result that full-wave dipole coupling is about 10 dB less than the half-wave dipole coupling for both parallel and collinear configurations. (These data are not shown here for lack of space.)

Fig. 11 shows the coupling between parallel half-wave dipoles and microstrip patches versus substrate thickness. The elements are resonant, and the spacing between elements is \(0.5 \lambda_0\). Mutual coupling is shown for both parallel and collinear configurations. For thin substrates the coupling levels are very low but increase rapidly with increasing thickness and then tend to oscillate for thicknesses greater than about \(0.5 \lambda_0\).

The dominant coupling mechanism for the parallel configuration is via space wave fields; since these fields are stronger in the broadside than in the endfire directions, the coupling levels between parallel dipoles are fairly large for close spacings, but
drop off quickly as the spacing increases. Surface waves are launched in the dipole’s endfire direction and so have most effect for the collinear configuration.

B. Array Efficiency—an Optimization Procedure

Section II-F discussed printed antenna element efficiency e, based on power lost to surface waves. This section will discuss what happens to e when elements are combined in an array, where it will be seen that e can be increased or decreased, depending on the element excitations, from the efficiency of an isolated element. An example will be given for two collinear half-wave dipoles spaced \( \lambda_0/2 \) apart on a quartz substrate. Microstrip patches can be treated in the same manner, and the procedure can be applied to arrays with more than two elements.

Since all mutual coupling terms between array elements would be included, (5) and (6) apply to the total input power and total radiated power, respectively, for an array of printed elements. The overall array efficiency, based on power lost to surface waves, can then be written in matrix form as

\[
e = \frac{[\mathbf{\Phi}]^T \mathbf{Re}[\mathbf{Z}_{\text{rad}}][\mathbf{I}]}{[\mathbf{\Phi}]^T \mathbf{Re}[\mathbf{Z}][\mathbf{I}]},
\]

(8)

where \([\mathbf{Z}]\) is the total (square) impedance matrix, \([\mathbf{Z}_{\text{rad}}]\) is the contribution to the radiated field, \([\mathbf{I}]\) is a column vector of expansion mode currents, and the superscript \(t\) denotes transpose. If we let \([\mathbf{R}] = \mathbf{Re}[\mathbf{Z}]\) and \([\mathbf{R}_{\text{rad}}] = \mathbf{Re}[\mathbf{Z}_{\text{rad}}]\), then (8) can be written as

\[
e = \frac{[\mathbf{\Phi}]^t[\mathbf{R}_{\text{rad}}][\mathbf{I}]}{[\mathbf{\Phi}]^t[\mathbf{R}][\mathbf{I}]}.
\]

(9)

To illustrate the method an example of two printed dipoles, with three expansion modes on each dipole, will be discussed. The \([\mathbf{Z}]\) and \([\mathbf{R}]\) matrices are then \(6 \times 6\), and \([\mathbf{I}]\) is a six-element column vector. If the expansion modes are numbered consecutively down each center-fed dipole, the terminal currents for the first and second dipoles will be \(I_2\) and \(I_5\), respectively. If \(V_1\) is the (port) voltage applied to dipole 1 and \(V_2\) is the (port) voltage applied to dipole 2, then

\[
\begin{bmatrix}
0 \\
V_1 \\
0 \\
V_2 \\
0
\end{bmatrix} = [\mathbf{Z}][\mathbf{I}].
\]

(10)

Now let \([\mathbf{I}_{\text{lo}}] = [\mathbf{I}]\) for \(V_1 = 1, V_2 = 0\) and let \([\mathbf{I}_{21}] = [\mathbf{I}]\) for \(V_1 = 0, V_2 = 1\). Then by superposition the dipole currents caused by port excitation voltages \(V_1\) and \(V_2\) can be written as

\[
[\mathbf{I}] = [\mathbf{S}][\mathbf{V}^p].
\]

(11)

where \([\mathbf{V}^p]\) is a two-element port voltage vector and \([\mathbf{S}] = [\mathbf{I}_{10}, [\mathbf{I}_{01}]]\) is a \(6 \times 2\) matrix. The array efficiency can then be expressed in terms of the port voltages as

\[
e = \frac{[\mathbf{V}^p]^t [\mathbf{S}^*]^t [\mathbf{R}_{\text{rad}}][\mathbf{S}][\mathbf{V}^p]}{[\mathbf{V}^p]^t [\mathbf{S}^*]^t [\mathbf{R}][\mathbf{S}][\mathbf{V}^p]}.
\]

(12)

Note that (12) expresses the performance index \(e\) as a ratio of two quadratic forms. Thus, (12) can be optimized by solving the eigenvalue problem,

\[
[A][\mathbf{V}^p] = e[B][\mathbf{V}^p],
\]

(13)

where \([A] = [\mathbf{S}^*]^t[\mathbf{R}_{\text{rad}}][\mathbf{S}]\) and \([B] = [\mathbf{S}^*]^t[\mathbf{R}][\mathbf{S}]\) are Hermitian \(2 \times 2\) matrices. A similar procedure has been used for free-space array optimization [14]. If the dipoles are identical then, by symmetry, it can be shown that \([A]\) and \([B]\) have the form

\[
[A] = \begin{bmatrix}
a_1 & a_2 \\
a_2 & a_1
\end{bmatrix}; \quad [B] = \begin{bmatrix} b_1 & b_2 \\
b_2 & b_1
\end{bmatrix},
\]

with \(a_1, a_2, b_1, b_2 \) real. The eigenvectors, representing the feed voltages for optimum \(e\), are then either even or odd:

\[
[V^p]_e = \begin{bmatrix} 1 \\ 1 \end{bmatrix}; \quad [V^p]_o = \begin{bmatrix} 1 \\ -1 \end{bmatrix}.
\]

(14)

The corresponding eigenvalues are then the efficiencies resulting from the above excitations:

\[
e_e = \frac{a_1 + a_2}{b_1 + b_2}; \quad e_o = \frac{a_1 - a_2}{b_1 - b_2}.
\]

(15)

Generally the even mode produces maximum \(e\) while the odd mode produces minimum \(e\).

The optimized efficiency \(e\) for two collinear half-wave dipoles \(\lambda_0/2\) apart versus substrate thickness \((\epsilon_r = 3.78)\) is shown in Fig. 12, for even and odd mode excitations. Also shown is the efficiency of an isolated dipole. As can be seen, the efficiency can be improved by as much as 30 percent for even mode excitation. A similar calculation for two parallel dipoles results in a 10 percent improvement for even mode excitation. For the data shown here, maximum calculation occurs for cophasal excitation of the array elements—a very practical result for broadside arrays. Odd-mode excitation generally produces a reduced efficiency, which means more power is being coupled to surface waves—a result which may be of interest for surface wave antennas. In Fig. 12 the \(e_e\) and \(e_o\) curves cross at about \(d = 0.19 \lambda_0\).

The change in efficiency for printed antenna elements in an array can be partially explained in terms of the phasings of the
surface wave fields. Surface waves, launched endfire from each dipole, are significantly out of phase (because of the $\lambda_0/2$ dipole spacing) and tend to cancel. It is hypothesized that an element spacing exists such that maximum cancellation occurs and $e \to 1$, at least for substrate thicknesses where only one surface wave mode exists. This situation would probably not occur when more than one surface mode is present, since the different phase constants would preclude total cancellation.

IV. CONCLUSION

This paper has presented data on resonant length, input resistance, bandwidth, surface wave power, dielectric loss, and mutual coupling for printed dipoles and microstrip patch antennas. Emphasis has been on the effect of the substrate, which may have a high dielectric constant and may not be electrically thin at millimeter wave frequencies.

Many other configurations of array elements could be studied for mutual coupling effects and surface wave power optimization. It is conceivable that for an array with many elements surface wave power can be made negligible for particular element spacings and excitations.

A very important need is to verify more of the theoretical calculations with measurements, particularly for printed dipole input resistance and surface wave losses for dipoles and patches. Some type of modified Wheeler cap method [15] may work for the latter measurement.

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