A Mutual Coupling Study of Linear and Circular Polarized Microstrip Antennas For Diversity Wireless Systems

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Abstract

In this paper, an analysis of mutual coupling is presented to examine the benefits of orthogonal polarizations and patterns for adjacent microstrip antennas. The mutual coupling between two linear polarized antennas orientated in parallel polarizations (E- and H-Plane) is reduced using low dielectric constant materials. The mutual coupling can be reduced an additional 20-35 dB at the same interelement spacing when adjacent elements are orientated in orthogonal polarizations, O-Plane. Similarly, the mutual coupling between two circular polarized antennas orientated in the parallel polarization is reduced using low dielectric constant materials. However, the reduction in mutual coupling between two circular polarized antenna elements orientated in the O-Plane is only an additional 1-6 dB. The mutual coupling between a linear polarized sum beam (1/2λ) and difference beam (1λ) antenna is reduced 20-35 dB below the case when using identical antennas only in the H- and O-Planes. Compact two and four element MultiElement Antennas (MEAs) with interelement spacings less than 0.15λ are fabricated and the S-parameters and radiation patterns are measured.

Keywords

Antenna arrays, antenna array mutual coupling, antenna diversity.

I. INTRODUCTION

Diversity reception techniques are well known methods to reduce the effects of multipath fading in mobile wireless communications [1], [2]. Additionally, recent information theory research has shown that in a Rayleigh fading environment the reliability of communication can be improved and the capacity increased by using Multi-Element Antenna (MEA) systems at the base and mobile in conjunction with appropriate spatio-temporal coding techniques [3], [4]. These techniques are being expanded to Wireless Local Area Networks (WLAN) [5].

There is a growing body of literature on the implementation of antenna diversity for personal devices (mobile stations). One of the topics of investigation is how the mutual coupling between the individual antenna elements relates to the diversity gain. The use of a space diversity antenna at the mobile station necessitates reduced element spacing which can lead to increased mutual coupling. The disadvantages of a large mutual coupling are the reduction in the antenna element radiation efficiency, the decrease in the antenna element effective gain [6], and the reduction in the average received power of the diversity antenna [2], [7]. One advantage is the change in the radiated fields of the individual antenna elements which creates an effective pattern (angle) diversity antenna [8]. The induced pattern diversity in addition to the spatial diversity reduces the correlation coefficients less than the free space prediction \( \rho_c = J_0^2(\beta d) \).
The use of a small number of elements in the MEA is preferred in traditional diversity reception techniques because the greatest degree in improvement in the diversity gain occurs in going from a single branch system to a two or four branch system (two branches give 11.5 dB gain at the 99% reliability level and four branches give 19 dB gain for maximal ratio combining) [1], [2]. In combined transmit/receive diversity there are substantial increases in capacity for a small number of elements (two branches give 7 bit/cycle at the 99% reliability level and four branches give 19 bits/cycle at an average received SNR of 21 dB, while there is only about 1.2 bits/cycle for a single branch) [3]. These considerable benefits for a small number of branches are of interest since the small physical size of the mobile devices will limit the number and placement of the antenna elements.

In this paper we present a mutual coupling study of aperture coupled linear and circular polarized microstrip antennas. The center frequency, \( f_0 = 5.2 \text{ GHz} \), is within the 5-6 GHz Unlicensed National Information Infrastructure (U-NII) band in the United States. This band offers three subbands of 100 MHz each, at 5.15-5.25, 5.25-5.35, and 5.725-5.825 GHz. Some of these bands are available in Europe. The effects of the antenna substrate, interelement distance, polarization, and antenna geometry are studied and analyzed in detail. Additionally, we present the study of mutual coupling between antenna elements with spatially orthogonal patterns. The motivation is to identify array topologies that minimize the mutual coupling under the constraint of small interelement spacing. The fundamental results of this work can be used in the design of compact MEAs utilizing space, polarization, and angle diversity.

The organization of the paper is as follows. First, we briefly review previous works on mutual coupling and discuss our contribution to the literature. Second, the analytical method to calculate the mutual coupling and the theoretical analysis of mutual coupling for dipoles is presented. Third, the mutual coupling between linear polarized antennas is investigated. Fourth, the mutual coupling between circular polarized antennas is investigated. Lastly, the mutual coupling between antennas with spatially orthogonal radiation patterns is presented. Two and four element MEAs are fabricated and the S-parameters and radiation patterns are measured. The realized MEAs are examples of practical designs for the IEEE 802.11a WLAN Standard.

II. PREVIOUS MUTUAL COUPLING RESEARCH

Theoretical research has shown that the radiation properties of printed antennas [9] and mutual coupling between printed dipoles [10] are affected by the radiated fields and the type and number
of surface wave modes propagating within the antenna substrate. Further experimental [11] and theoretical [12] research has investigated the effect of antenna orientation, geometry, and substrate on mutual coupling for the H-Plane and E-Plane. The antennas in all these studies were linear polarized and radiated the same spatial field (maximum directed broadside to the antenna). Research into the mutual coupling between wire antennas of different lengths was conducted in the past [13], [14] and it was shown that the coupling was reduced. The length of the wire antennas was such that they radiated similar radiation patterns and the analysis assumed a free space environment. Theoretical work on mutual coupling between rectangular patch antennas gave two examples for different radiating modes [15] but was incomplete in the antenna diversity context because it did not consider different antenna orientations or mode combinations.

Our contribution to the further understanding of the mutual coupling between printed antennas are as follows. We study the orthogonal orientation between antenna elements. This is relevant in the context of polarization diversity antenna systems. We extend the analysis to circular polarized antenna elements. The advantages of circular polarization to linear polarization in single branch communications systems to reduce the effects of multipath are well known and the extension to multibranch systems is a topic of new research. We also investigate the mutual coupling between antenna elements with spatially orthogonal radiation patterns. This is particularly useful for pattern (angle) diversity antenna systems. Lastly, we consider the effects of the feed substrate.

This work only considers single polarized antenna elements for clarity and conciseness when comparing these results with the fundamental research mentioned previously. The use of dual polarized antenna elements [16], [17] may be a topic of future research.

III. ANALYTICAL METHODS AND THEORETICAL ANALYSIS

A. MPIE MoM Formulation

The analysis and design of the antennas in this research was accomplished using a mixed potential integral equation method of moments software [18]-[20]. A brief description of the theoretical analysis germane to the calculation of the mutual coupling is included for completeness.

The geometry of an aperture coupled microstrip antenna is shown in Figure 1. The electric current \( \vec{J}_1 \) on the lower microstrip line, the tangential electric field \( \vec{E}_2 \) in the aperture, and the electric current \( \vec{J}_3 \) on the upper microstrip antenna are modeled with the triangular patch subdomain basis function, \( \vec{f}_n = t^2 i_n(\vec{r}) \). The total unknown electric and magnetic current distributions
are expressed as,

\[ \vec{J}_1 = \sum_{n=1}^{N_1} A_n \vec{f}_{n1}, \quad \vec{M}_2 = \sum_{n=1}^{N_2} B_n \vec{f}_{n2}, \quad \vec{J}_3 = \sum_{n=1}^{N_3} C_n \vec{f}_{n3} \]  

(1)

An application of the equivalence principle and the use of Galerkin’s procedure transforms the system of coupled integral equations to a system of linear equations \([V] = [Z][I]\) expressed as,

\[
\begin{bmatrix}
[\vec{E}_{inc}, \vec{f}_{m1}] \\
[0] \\
[0]
\end{bmatrix} =
\begin{bmatrix}
[Z_{11}] & [W_{12}] & [0] \\
[U_{21}] & [Y_{22}] & [U_{23}] \\
[0] & [W_{32}] & [Z_{33}]
\end{bmatrix} \cdot
\begin{bmatrix}
[A] \\
[B] \\
[C]
\end{bmatrix}
\]

(2)

where \([Z_{ii}]\) and \([Y_{jj}]\) are the self coupling integral submatrices, and \([W_{ij}]\) and \([U_{ji}]\) are the mutual coupling integral submatrices, between two basis functions located on the same or different layers, respectively.

In this analysis, the excitation mechanism is an impressed electric field \(\vec{E}_{inc}\) generated by a series unit voltage gap source placed across the positive to negative triangles of a single basis function near the end of the microstrip feed line.

In order to extract the S parameters for the 2-port network, the following procedure is adopted. At two consecutive sampling points \(x_i\) and \(x_{i+1}\) along the \(\hat{x}\)-directed port \(p\) in the dominant-mode region, which is far from the excitation and the discontinuities, the currents can be expressed as,

\[
I_i = I_{pq}(x_i) = A_{pq} e^{-j\beta(x_i-x_{ref})} - B_{pq} e^{j\beta(x_i-x_{ref})}
\]

\[
I_{i+1} = I_{pq}(x_{i+1}) = A_{pq} e^{-j\beta(x_{i+1}-x_{ref})} - B_{pq} e^{j\beta(x_{i+1}-x_{ref})}
\]

(3)

(4)

where \(I_{pq}\) is the total current in port \(p\) with the excitation in port \(q\). It can be separated into an incident \((A_{pq})\) and a reflected \((B_{pq})\) component. \(\beta\) is the propagation constant of the dominant mode along a uniform infinite microstrip line, which is determined in advance. \(x_{ref}\) is the location of the reference plane.

The coefficients \(A_{pq}\) and \(B_{pq}\) can be determined from these two sampling points \(I_i\) and \(I_{i+1}\). For the region far away from the discontinuity and the excitation point, a standing wave will be observed. Therefore, \(A_{pq}\) and \(B_{pq}\) are constants, independent of \(x\). We can extract their values by a simple curve-fitting algorithm. In this analysis, typically 5 points are sampled and the extracted constants \(A_{pq}\) and \(B_{pq}\) are averaged in this dominant-mode region. If the excitation is connected at port 1 for a two-port network, an under-determined matrix equation can be expressed as,
where $A_{p1}$ and $B_{p1}$ ($p = 1, 2$) are determined from the total current $I_{pq}$ at each port. To solve the two-port scattering parameters, we individually connect the excitation at each port and leave the other port open. As a result, a $2 \times 2$ matrix can be constructed by extracting $A_{pq}$ and $B_{pq}$ from each port $p$ for each excitation at port $q$ as,

\[
\begin{bmatrix}
A_{11} & A_{12} \\
A_{21} & A_{22}
\end{bmatrix}
\begin{bmatrix}
S_{11} & S_{12} \\
S_{21} & S_{22}
\end{bmatrix}
= \begin{bmatrix}
B_{11} \\
B_{21}
\end{bmatrix}
\]

then the two-port scattering parameters can be solved without difficulties. Since the impedance matrix $[Z]$ is unchanged for different excited ports, it is not necessary to recompute the matrix elements and matrix inversion.

**B. Dipole Mutual Coupling Analysis**

In free space the two dipoles are located along the z-axis of a cartesian coordinate system. The parallel (H-Plane) orientation is defined when the two antennas are located at the same $z$ coordinate but displaced by a distance $\rho$ along the x-y plane. The colinear (E-Plane) orientation is defined when the two antennas are located at the same $x,y$ coordinate but displaced by a distance $\rho$ along the common z-axis. The mutual coupling vs. interelement distance $\rho$ is greater for the H-Plane orientation than for the E-Plane orientation [21]. The mechanism for the mutual coupling is the radiated space wave, whose electric field can be expressed as,

\[
E_{\text{rad}} \propto \frac{\sin \theta}{\rho}
\]

The electric field magnitude is maximum in a plane perpendicular to the dipole and zero along the axis of the dipole.

In a grounded dielectric substrate the number of surface wave modes generated by the antenna depends upon the electrical thickness and dielectric constant of the the substrate. The values of the antenna substrate and the frequency range in this paper will allow only the lowest order $TM_0$ surface wave mode to propagate. For a x-directed ($\phi = 0^\circ$) electric dipole, the electric field spatial variation of the $TM_0$ surface wave mode varies as,
\[ E_{aw} \propto \frac{\cos \phi}{\sqrt{\rho}} \]  

The magnitude is proportional to the dielectric constant and electrical thickness of the substrate [10] and is maximum along the axis of the dipole and zero along the axis perpendicular to the dipole.

IV. LINEAR POLARIZED ANTENNA ELEMENTS

A. Theoretical Analysis and Discussion

The antenna element used in this analysis is a microstrip antenna fed by a symmetrical cross aperture. The aperture is excited by an open end microstrip feedline on the other side of the common ground plane, refer to Figure 1. The advantages of this non-contact vertical transition design are well known [22], there is no physical contact between the antenna and the feed, and the antenna and the feed substrates can be individually optimized. For brevity, this paper does not discuss the analysis and design of the individual antenna elements. The interested reader can review the appropriate references [22], [23] to develop an understanding of the design parameters.

Initially, an analysis was conducted of various isolated canonical antenna geometries. The parameters of interest were physical size, impedance bandwidth, and cross-polarization. The circular ring-slit was chosen since it had the smallest size and lowest cross-polarization. A comparison of a rectangular shape and a cross slot shape demonstrated that the physical size of the aperture is smaller using the latter geometry.

In diversity antennas for portable devices a mutual coupling \((S_{21})\) less than \(-15\) dB has generally been considered adequate. Figure 2 shows the three orientations the antenna pair can take. In the H-Plane and E-Plane case both Port 1 and Port 2 transmit (receive) the same sense of linear polarization, while in the Orthogonal Plane, O-Plane, case Port 1 transmits (receives) vertical linear polarization while Port 2 transmits (receives) horizontal linear polarization.

First, we compare the mutual coupling between two circular ring-slit antenna elements for a constant antenna substrate thickness \(d_a = 2.54\) mm \((d_a/\lambda_0 = 0.044)\) and three dielectric constants of \(\varepsilon_{rn} = 1.07\) (rigid foam plastic), 2.2 (PTFE composite), and 4.5 (ceramic polymer composite). Our previous work has demonstrated the same qualitative results for different substrate thicknesses [24].

When the antenna pair is orientated in the H-plane, see Figure 3 (a), the mutual coupling is
lowest for the rigid foam substrate, \(\epsilon_{ra} = 1.07\), at all interelement spacings. The mutual coupling in the H-Plane (broadside) orientation is due mainly to near field behavior (higher order wave) for small distances and to far field behavior (space wave) for larger distances [10]. The nonzero \(y\)-component of the antenna current is the source of surface waves flowing in the direction of the antennas which causes the difference in magnitude in mutual coupling between the three substrates.

When the antenna pair is in the E-plane, see Figure 3 (b) the mutual coupling is lowest for \(\epsilon_{ra} = 1.07\) when the interelement spacings are greater than \(D/\lambda_0 > 0.25\). For smaller interelement spacings, \(D/\lambda_0 \leq 0.25\), the mutual coupling is less when \(\epsilon_{ra} = 2.2\). Additionally, the mutual coupling is less when \(\epsilon_{ra} = 4.5\) than for \(\epsilon_{ra} = 1.07\) for small spacings, \(D/\lambda_0 \leq 0.15\). For small interelement distances the near field is dominant. The larger dielectric constant reduces the equivalent resonant length which decreases the near field strength which leads to reduced mutual coupling. For larger interelement distances the mutual coupling is dominated by the surface wave. The maximum power density of the \(TM_0\) surface wave mode is now in the same direction as the antenna elements, which causes the increased mutual coupling as the dielectric constant is increased and more coupling at the same interelement distance as compared to the H-Plane.

When the antenna pair is orientated in the O-Plane, refer to Figure 3 (c), the mutual coupling is reduced an additional 20-35 dB at the same interelement spacing when compared to the to the H-plane and E-plane. Since the antenna elements are now orthogonal to the radiated and surface wave fields, there is essentially no coupling mechanism between elements. The advantages of the O-Plane orientation are the ability to receive orthogonal polarizations caused by the highly scattering environment [2], [25] or the communication system can be designed for polarization diversity [26].

We next investigate the mutual coupling between two elements for three antenna geometries, circular ring-slit, square ring, and rectangular patch. In the H-Plane the mutual coupling is less for the rectangle patch at all interelement spacings by 3-6 dB. This difference in magnitude has been previously observed [11] and is due to the more symmetric current flowing on the rectangular patch. The difference in mutual coupling in the E-Plane is very small, less than 2 dB, for the three antenna geometries.

Lastly, the mutual coupling was investigated between two rectangular patch antennas for different feed substrate dielectric constants and thicknesses. There were no differences in mutual coupling for different dielectric constants in the three orientations. A similar analysis was conducted using different substrate thicknesses and found no differences.
B. Experimental Results

A 2 × 2 O-Plane MEA of rectangular patch antennas was fabricated and measured. The array geometry is given in Figure 4. The length and width of the rectangular patch element is \( L = 22.7 \text{ mm} \) \((L/\lambda_0 = 0.39)\) and \( W = 35.0 \text{ mm} \) \((W/\lambda_0 = 0.61)\). The side length of the MEA is \( S/\lambda_0 = 1.23 \). The interelement spacing is less than 0.15\( \lambda_0 \). The rectangular patch antenna is used due to the reduced mutual coupling in the H-Plane compared to the circular ring-slit antenna.

The fabrication procedure was as follows. First, the antennas were etched from a single double sided copper clad thin dielectric (RT/Duroid 5880, \( \epsilon_r = 2.2, t = 0.762 \text{ mm} \)). Second, the microstrip feedline and aperture were simultaneously etched from a single double sided copper clad substrate (GML 1000, \( \epsilon_{rf} = 3.2, d_f = 0.762 \text{ mm} \)). Third, cyanoacrylic adhesive was placed between the antenna substrate, hard foam substrate (Rohacell 51-HF, \( \epsilon_{ra} = 1.07, d_a = 2.0 \text{ mm} \)), and the feed substrate and then the three pieces were pressed 30 seconds until the adhesive dried.

The measured scattering parameters for the rectangular patches are shown in Figure 5. The reflection coefficients are centered at 5.20 GHz for the four ports. The impedance bandwidths are 9.5%. The mutual coupling parameters demonstrate the advantages of the O-Plane configuration and a hard foam substrate. The elements in the orthogonal orientation (1,2), (1,3) have mutual coupling magnitudes less than -38 dB while the diagonal elements (1,4) which are in the parallel orientation, have mutual coupling magnitudes nearly -35 dB. Lastly, the measured radiation patterns at the center frequency reveal cross polarization magnitudes less than -25 dB for each port. The maximum magnitude of the main beam is directed broadside with beam squint of less than 5°. The beamwidths are 70° and 60° in the E-Plane and H-Plane, respectively. A symmetric radiation pattern has been shown to reduce the fading depth in polarization diversity systems [26].

In summary, the antenna substrate dielectric constant and the sense of antenna polarization were shown to be the two dominant parameters in the magnitude of mutual coupling. When two adjacent elements are orientated in the H-Plane the mutual coupling is reduced at all interelement spacings as the dielectric constant is reduced. When two adjacent elements are orientated in the E-Plane the mutual coupling is reduced at large interelement spacing as the dielectric constant is reduced. For small interelement spacings the mutual coupling is reduced as the dielectric constant is increased. When two adjacent elements are orientated in the O-Plane the mutual coupling is reduced an additional 20-35 dB as compared to the H- and E-Planes. The value of the dielectric constant has little influence in this case. The feed substrate does not influence the mutual coupling.
A 2 x 2 MEA oriented in the O-Plane was fabricated. The interelement spacing was $D = 0.15\lambda_0$ and the mutual couplings were less than -35 dB. The radiation patterns were slightly perturbed.

V. CIRCULAR POLARIZED ANTENNA ELEMENTS

A. Theoretical Analysis and Discussion

Space and polarization diversity reception techniques have only used linear polarization [6]-[8], [26]. Recent literature has considered circular polarization [27]-[30] and suggests the potential benefits of circularly polarized antennas. The antenna geometry in this analysis is a circular ring with two diametrically positioned rectangular stubs, see Figure 6. Circular polarization is produced by perturbing the antenna geometry to excite two orthogonal modes and placing the feed such that these two modes are excited in phase quadrature [31]. When the stubs are placed at the angles $\Phi = 45^\circ$ and $-135^\circ$ the fields are right hand circular polarized (RHCP) and when the stubs are placed at the angles $\Phi = -45^\circ$ and $135^\circ$ the fields are left hand circular polarized (LHCP). Additional theoretical analysis of the antennas can be found in our previous work [32].

Figure 6 defines the two orientations the antenna pair can take. In the Parallel Plane, P-Plane, both Port 1 and Port 2 transmit (receive) the same sense of circular polarization, while in the Orthogonal Plane, O-Plane, Port 1 transmits (receives) LHCP while Port 2 transmits (receives) RHCP.

Like the linear polarization case, we first compare the the mutual coupling between the antenna elements for a constant antenna thickness $d_a = 1.27$ mm (0.022$\lambda_0$) and three dielectric constants $\epsilon_{ra} = 1.07, 2.2, 4.5$. The electrical parameters of the substrates allow only the lowest order $TM_0$ surface wave mode to propagate. Our previous work show the same qualitative results for increased substrate thicknesses [33]. Independent work presented at the same conference show the same qualitative results for the P-Plane [34].

When the antenna element pair is in the P-Plane, see Figure 7 (a), the mutual coupling is minimum for $\epsilon_{ra} = 1.07$ and maximum for $\epsilon_{ra} = 4.5$ for all interelement spacings. At large interelement distances the mutual coupling is mainly due to the surface wave mode whose magnitude is proportional to the dielectric constant and electrical thickness of the substrate [9]. Similar results were previously shown for two linear polarized antennas orientated in the H-Plane, see Figure 3 (a). The effect of the mutual coupling on the axial ratio (AR) is non-negligible for small interelement spacings ($D/\lambda_0 < 0.3$) and for larger antenna substrate dielectric constants. We also observe an
oscillatory value of the AR even for large interelement distances due to the surface wave mode.

When the antenna pair is in the O-Plane, see Figure 7 (b), the mutual coupling is minimum for $\varepsilon_{ra} = 1.07$ and maximum for $\varepsilon_{ra} = 4.5$ for larger interelement spacings, $D/\lambda_0 > 0.4$. For small interelement distances, $D/\lambda_0 < 0.4$, the mutual coupling when $\varepsilon_{ra} = 2.2$ is less than when $\varepsilon_{ra} = 1.07$. Additionally, for interelement distances $D/\lambda_0 < 0.1$ the mutual coupling for $\varepsilon_{ra} = 1.07$ and $\varepsilon_{ra} = 4.5$ are nearly the same. For small interelement spacings the near field is the dominant coupling mechanism. The larger dielectric constant reduces the equivalent resonant length which decreases the near field strength which leads to reduced mutual coupling. Similar results for small interelement spacings were demonstrated previously for two linear polarized antennas orientated in the E-Plane, see Figure 3 (b). The axial ratio has the same response as in the P-Plane orientation.

Second, we compare the mutual coupling of the antenna pair for a constant antenna substrate dielectric constant and thickness. We observe that for $\varepsilon_{ra} = 1.07$ the mutual coupling is increased 3-8 dB when the antenna pair is placed in the O-Plane. The increase is due to the constructive addition of the reflected space wave from the ground plane to the direct space wave component. For $\varepsilon_{ra} = 2.2$ and 4.5, the mutual coupling is decreased 2-6 dB and 1-3 dB, respectively, when the antenna pair is placed in the O-Plane. This suggests that the mutual coupling due to the surface wave is nearly constant of the sense of circular polarization.

Similar to the linear polarized case the the analysis was repeated for different feed substrates. The results were unchanged.

B. Experimental Results

The reduced mutual coupling observed in the O-Plane for small interelement distances, see Figure 7 (b), can be utilized to design compact diversity antennas. Two $2 \times 1$ O-Plane MEAs with interelement distance of $D = 0.15\lambda_0$ were fabricated on antenna substrates of foam ($\varepsilon_{ra} = 1.07$) and RT/Duroid ($\varepsilon_{ra} = 2.2$). The geometry is shown in Figure 8. Using a foam substrate the size of the array is $0.4\lambda_0 \times 1.0\lambda_0$ while the PTFE substrate gives an array size of $0.3\lambda_0 \times 0.8\lambda_0$. The measured S-parameters are displayed in Figure 9. The impedance bandwidth is slightly larger using the foam substrate, 7.2%, than the PTFE substrate, 5.9%. However, the mutual coupling is reduced using the PTFE substrate, -25 dB, as compared to -18 dB for the hard foam substrate. The theoretical axial ratio 3 dB bandwidths, 1.5%, are nearly the same for both substrates. The fabrication procedure is the same as the linear polarized case, see IV-B.
In summary, the antenna substrate dielectric constant and the sense of antenna polarization were shown to be the two dominant parameters in the magnitude of mutual coupling. When two adjacent elements are oriented in the P-Plane the mutual coupling is reduced at all interelement spacings as the dielectric constant is reduced. When two adjacent elements are orientated in the O-Plane the mutual coupling is reduced at large interelement spacing as the dielectric constant is reduced. For small interelement spacings the mutual coupling is reduced as the dielectric constant is increased. For an antenna substrate with unity dielectric constant the mutual coupling is less when the adjacent elements are orientated in the P-Plane (5-8 dB). However, for a non-unity dielectric constant the mutual coupling is less when the adjacent elements are orientated in the O-Plane (3-10 dB). The feed substrate does not influence the mutual coupling. Two $2 \times 1$ MEAs oriented in the O-Plane were fabricated. The interelement spacings were $D = 0.15\lambda_0$ and the mutual couplings were -18 dB and -25 dB, for $\epsilon_{ra} = 1.07, 2.2$, respectively.

VI. LINEAR POLARIZED ANTENNA ELEMENTS WITH SPATIALLY ORTHOGONAL RADIATION PATTERNS

A. Theoretical Analysis and Discussion

The use of antenna elements with spatially orthogonal radiated fields have been previously shown to reduce the correlation coefficients [35], [36]. The antenna element in this paper is the rectangular patch antenna. Two orthogonal spatial fields can be radiated by properly choosing the resonant length. If the resonant length is $1/2$ wavelength a single main beam is radiated normal to the antenna (sum beam). If the resonant length is $1$ wavelength two main beams are radiated at an angle $\theta \approx \pm 45^\circ$ with a null normal to the antenna (difference beam). For brevity we study the mutual coupling for a single antenna substrate thickness ($d_a = 1.524$ mm, $d_a/\lambda_0 = 0.026$) and dielectric constant ($\epsilon_{ra} = 3.38$).

The first study is when the antennas are orientated in the H-Plane. The mutual coupling as a function of interelement distance is shown in Figure 10 (a). The three curves correspond to different combinations of antenna elements. We observe that the mutual coupling is similar when the two antennas are the same, either two $1/2\lambda$ or two $1\lambda$ antennas. This is expected in this orientation. The mutual coupling is mostly due to the radiated space wave and both pairs have identical elements that radiate the same spatial patterns. When the antenna pair consists of one $1/2\lambda$ and one $1\lambda$ the mutual coupling is significantly reduced, greater than 20 dB. The reason being that the radiated
fields are spatially orthogonal to each other.

The second study is when the antennas are orientated in the E-Plane. The mutual coupling as a function of interelement distance is shown in Figure 10 (b). In this case the mutual coupling is dominated by the excited surface wave. All three pairs show the same coupling trend. Interestingly, the pair of antennas consisting of two 1\lambda antennas has more coupling (10 dB) at the same interelement distance than two 1/2\lambda antennas. The antenna pair consisting of the mixed pair have mutual coupling in between the other pairs consisting of identical elements.

Additionally, when the antennas are orientated in the O-Plane, the mutual coupling is reduced for the three combinations of antennas. This is similar to the previous case in Section IV.

B. Experimental Results

Using the results of the H-Plane analysis, the concept of a multibeam MEA is presented. The geometry of a fabricated 4 \times 1 H-Plane MEA is shown in Figure 11. The interelement spacing is \( D = 0.10\lambda_0 \). The size of the 4 \times 1 H-Plane MEA is \( S = 1.42\lambda_0 \) (8.2 cm) \( \times \) \( H = 0.53\lambda_0 \) (3.1 cm). The substrate material used for both antenna and feed substrates is a relatively new substrate material that is gaining industry acceptance, Rogers RO4003 High Frequency Material. The dielectric constant is 3.38, it is low loss (\( \tan \delta = 0.0027 \) at 10 GHz), low cost, and processes the same as standard epoxy/glass. RO4003 materials are proprietary woven glass reinforced hydrocarbon/ceramics (non PTFE).

The measured S-parameters of the MEA is shown in Figure 12. The impedance bandwidths are 4.3% and 1.8% for the sum and difference ports, respectively. The mutual couplings are greatest for the element pairs consisting of identical elements (3,1) and (4,2) with \( S_{31} \approx S_{42} = -23 \) dB, as predicted by the previous analysis. The mutual coupling between the mixed elements, (2,1), (4,1), and (3,2), are all less than -35 dB.

Lastly, a 2 \times 1 H-Plane MEA composed of 1/2\lambda and 1\lambda rectangle patch antennas was fabricated on a high dielectric constant antenna substrate (Rogers RO3006, ceramic filled PTFE composite, \( \varepsilon_r = 6.15, \tan \delta = 0.0025 \) at 10 GHz). The geometry and scattering parameters are shown in Figure 13. The high dielectric constant reduces the physical size of the MEA, \( S = 0.68\lambda_0 \) (3.9 cm) \( \times \) \( H = 0.39\lambda_0 \) (2.3 cm). The impedance bandwidths are also reduced, 2.4% (125 MHz) and 1.1% (55 MHz) for the sum and difference port, respectively. The theoretical antenna element radiation efficiency \( \eta_a = P_{rad}/(P_{rad} + P_{sw}) \) is 50%. This is caused by the greater percentage of available power
being transferred to the surface wave mode. The increased surface wave power is not converted to an increased mutual coupling as in the case of two antennas of the same radiation pattern, refer to Figure 3 (a). The orthogonality of the radiated fields lead to a measured mutual coupling of $S_{21} = -31$ dB. The measured copolar radiation patterns are shown in Figure 14. The 3 dB beamwidths for the sum port are 110° and 80° in the E-Plane and H-Plane, respectively. The cross polarization is less than -20 dB in both planes. The null at boresight in the E-Plane for the difference port is -25 dB. The fabrication procedure is the same as the linear polarized case, see IV-B.

In summary, the mutual coupling of sum beam (1/2λ) and difference beam (1λ) antennas was investigated to determine the optimum element placement and orientation. The mutual coupling between a sum beam and difference beam antenna is lowest when the pair is orientated in the H-Plane and O-Plane. In the E-Plane there is significant mutual coupling due to the propagating surface wave mode. A 4 × 1 H-Plane MEA was fabricated using ceramic (non PTFE) substrates for the antenna and feed. The interelement spacings were $D = 0.10\lambda_0$ and the mutual couplings were less than -20 dB. A 2 × 1 H-Plane MEA was fabricated on a high dielectric constant PTFE antenna substrate. The interelement spacing was $D = 0.10\lambda_0$ and the mutual coupling was less than -30 dB. The antenna radiation patterns were unperturbed.

VII. CONCLUSION

This paper has presented a study of independent microstrip antennas to form a MultiElement Antenna (MEA) to realize diversity antennas. Linear and circular polarized antenna elements with identical radiation patterns (maximum at boresight) were considered. The use of antenna substrates with low dielectric constants for the MEA was shown to reduce the mutual coupling between elements. The use of orthogonal polarization (O-Plane) in the MEA was shown to further reduce the mutual coupling and allow a reduction in the interelement spacing. The mutual coupling was independent of the feed substrate. Linear polarized antenna elements with spatially orthogonal radiation patterns were also considered. The mutual coupling of sum beam (1/2λ) and difference beam (1λ) antennas is lowest when the pair is orientated in the H-Plane and O-Plane. In the E-Plane there is significant mutual coupling due to the propagating surface wave mode. Two and four element MEAs with interelement spacing $D \leq 0.15\lambda_0$ and mutual couplings less than -30 dB were realized. The MEAs were fabricated on low cost and low loss hard foam, ceramic, and PTFE substrates. A two element single polarized multibeam MEA was fabricated on a high dielectric
constant (PTFE) material. The spatial and polarization diversity of the linear and planar MEAs will be used in future studies of correlation coefficients and channel capacities.

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REFERENCES


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Fig. 1. Geometry of a single aperture coupled microstrip antenna.

Fig. 2. Theoretical $2 \times 1$ element arrays of linear polarized (LP) antennas with definitions of three orientations of elements. In the H-plane the radiated electric field of both ports are parallel to the x-axis. In the E-plane the radiated electric field of both ports are parallel to the y-axis. In the O-plane case the radiated electric field of Port 1 is parallel to the y-axis and the radiated electric field of Port 2 is parallel to the x-axis.
Fig. 3. Comparison of theoretical mutual coupling as a function of distance between two circular ring-slit LP antenna elements for three antenna substrate dielectric constants. (a) H-Plane, (b) E-Plane, (c) O-Plane. $f_0 = 5.2$ GHz, $d_a = 2.54$ mm ($d_a/\lambda_0 = 0.044$), $\varepsilon_{r_a} = 10.8$, $d_f = 0.635$ mm.
Fig. 4. Fabricated geometry of a 2 x 2 O-Plane LP MEA of rectangular patches. \( f_0 = 5.2 \) GHz.

Fig. 5. Measured S-parameters of a 2 x 2 O-Plane LP MEA of rectangular patch antennas, \( D/\lambda_0 = 0.13, f_0 = 5.2 \) GHz. (a) \( S_{ii} \), (b) \( S_{ij} \), \( d_a = 2.0 \) mm, \( \epsilon_{ra} = 1.07 \), \( d_f = 0.762 \) mm, \( \epsilon_{rf} = 3.2 \).
Fig. 6. Theoretical 2 × 1 element arrays of circular polarized (CP) antenna elements with definitions of two orientations of elements. In the P-Plane (Parallel Orientation) the radiated electric fields are of the same sense of polarization. In the O-Plane (Orthogonal Orientation) the radiated electric fields are of the opposite sense of polarization.

Fig. 7. Comparison of theoretical mutual coupling ($S_{21}$) and axial ratio ($AR$) as a function of interelement distance between two CP antenna elements for three antenna substrate dielectric constants. (a) P-Plane, (b) O-Plane. $d_a = 1.27$ mm, $d_f = 0.635$ mm, $\varepsilon_r = 10.8$, $f_0 = 5.2$ GHz.
Fig. 8. Geometry of a fabricated 2 x 1 O-Plane CP MEA for \( d_a = 2.0 \) mm and \( \epsilon_{ra} = 1.07 \). For \( d_a = 1.57 \) mm and \( \epsilon_{ra} = 2.2 \), \( W/\lambda_0 = 0.3 \) and \( L/\lambda_0 = 0.8 \). \( d_f = 0.762 \) mm and \( \epsilon_{rf} = 3.2 \) for both cases.

Fig. 9. Comparison of measured S-parameters of 2 x 1 O-Plane CP arrays for two antenna substrate dielectric constants, \( \epsilon_{ra} = 1.07, 2.2 \). (a) \( S_{11} \), (b) \( S_{21} \). \( D/\lambda_0 = 0.15 \). \( d_a = 2.0 \) mm for \( \epsilon_{ra} = 1.07 \), \( d_a = 1.57 \) mm for \( \epsilon_{ra} = 2.2 \). \( d_f = 0.762 \) mm, \( \epsilon_{rf} = 3.2 \).
Fig. 10. Comparison of theoretical mutual coupling as a function of distance between two rectangle patch LP antennas for three different antenna pairs. (a) H-Plane, (b) E-Plane. $\epsilon_r = 3.38, d_a = 1.524 \text{ mm}, \epsilon_r = 3.38, d_f = 0.508 \text{ mm}$. $f = 5.2 \text{ GHz}$.

Fig. 11. Geometry of a fabricated 4 x 1 H-Plane LP MEA composed of 1/2$\lambda$ and 1$\lambda$ rectangular patch antennas. Ports 1 and 3 radiate a sum pattern and Ports 2 and 4 radiate a difference pattern. $D/\lambda_0 = 0.10$, $f_0 = 5.20$ GHz. $\epsilon_r = 3.38, d_a = 1.524 \text{ mm}, \epsilon_r = 3.38, d_f = 0.508 \text{ mm}$. 
Fig. 12. Measured S-parameters of a 4 x 1 H-Plane LP MEA composed of 1/2λ and 1λ rectangle patch antennas. (a) $S_{ii}$, (b) $S_{ij}$. $D/\lambda_0 = 0.10$, $f_0 = 5.20$ GHz, $d_a = 1.524$ mm, $\varepsilon_{r_a} = 3.38$, $d_f = 0.508$ mm, $\varepsilon_{r_f} = 3.38$.

Fig. 13. (a) Geometry of a fabricated 2 x 1 H-Plane LP MEA composed of 1/2λ and 1λ rectangle patch antennas, (b) Measured scattering parameters. $D/\lambda_0 = 0.10$, $f_0 = 5.20$ GHz. $\varepsilon_{r_a} = 6.15$, $d_a = 1.27$ mm, $\varepsilon_{r_f} = 3.38$, $d_f = 0.508$ mm.
Fig. 14. Measured co-polar radiation patterns, (a) Port 1 (sum), (b) Port 2 (difference), of a $2 \times 1$ H-Plane LP MEA. $f = 5.2$ GHz. $\epsilon_r = 6.15, d_a = 1.27$ mm, $\epsilon_r f = 3.38, d_f = 0.508$ mm.