The Planar Ultrawideband Modular Antenna (PUMA) Array

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The authors humbly dedicate this work in memory of Professor Benedikt A. Munk.

Abstract—A fully planar ultrawideband phased array with wide scan and low cross-polarization performance is introduced. The array is based on Munk's implementation of the current sheet concept, but it employs a novel feeding scheme for the tightly coupled horizontal dipoles that enables simple PCB fabrication. This feeding eliminates the need for "cable organizers" and external baluns, and when combined with dual-offset dual-polarized lattice arrangements the array can be implemented in a modular, tile-based fashion. Simple physical explanations and circuit models are derived to explain the array's operation and guide the design process. The theory and insights are subsequently used to design an exemplary dual-polarized infinite array with 5:1 bandwidth and VSWR < 2.1 at broadside, and cross-polarization ≈ -15 dB out to $\theta = 45^{\circ}$ in the D- plane.

Index Terms—Dipole arrays, phased arrays, planar arrays, ultra wideband antennas.

I. INTRODUCTION

U LTRAWIDEBAND antenna arrays with wide-scan ability and polarization diversity are in great demand for multifunctional systems, [1]. Such systems use phased arrays that can simultaneously accommodate multiple frequency bands, beams, and polarizations in order to consolidate multiple antennas into a single ultrawideband aperture. Ultrawideband apertures are typically non-modular and expensive to build and assemble, and often require external baluns to integrate with RF front-ends.

One of the most popular wideband arrays is the tapered-slot (Vivaldi) array, which has been utilized in many UWB systems [2]. Extensive work spanning three decades has resulted in Vivaldi arrays that achieve bandwidths in excess of 10:1 at wide-

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scans [3]. Additionally, the integrated Knorr balun of the Vivaldi elements provides an unbalanced 50Ω feed interface. Despite excellent electrical performance, wideband Vivaldis consist of deep, vertically-integrated notches that must be electrically connected to avoid resonances [4], resulting in complicated dual-polarized arrays that are often expensive to manufacture and assemble. Modular Vivaldi array variations, such as the body of revolution (BOR) Vivaldi array [5], or the Mechanotch array [6] exist, but rely on special machining and are difficult to implement at high frequencies (above X-band).

Several dipole-like elements have been developed as low-profile and modular alternatives to Vivaldi arrays, and consist of vertically integrated PCB cards backed with a ground plane. These low-profile (depth $<\lambda_{high}/2$, independent of bandwidth) elements do not require electrical connection, and are often tapered vertically from the feed-lines. The Bunny-ear [7], balanced antipodal Vivaldi antenna (BAVA) [8], doubly-mirrored BAVA (Dm-BAVA) [9] and Banyan tree antenna (BTA) [10], [11], [33] arrays belong to this class. The Bunny-ear array has demonstrated a 5:1 bandwidth, using external baluns and resonance-suppression resistors between its element arms and ground [7]. The Dm-BAVA alleviates these resonances using element mirroring in E-and H- planes, but requires 180° hybrids to maintain beam collimation. The BTA array has achieved bandwidths up to 4:1-without baluns or hybrids—by using shorting strips between the arms and the ground plane.

Vertically-integrated arrays are difficult to integrate in dual-polarized arrangements at higher frequencies, thus fully planar wideband topologies are desirable. To date, several "quasi-planar" wideband arrays have been published. The current sheet antenna (CSA) array [12], which cleverly realizes Wheeler's current sheet, [13], uses tightly-coupled horizontal dipoles above a ground plane to achieve bandwidths of 9:1 with a printable element layer. Alternatively, array apertures formed from periodically fed continuous sheets or slots, such as the fragmented aperture array (FAA) [14], [15], and the long slot array, [16], can produce a stable impedance over wide bandwidths; however, both arrays exhibit bidirectional radiation, and when backed by a ground plane suffer from catastrophic resonances that limit their performance to approximately 4:1 [17]. Alternatively, these resonances can be suppressed using Jaumann screens [18] or ferrite loading [19], which decreases radiation efficiency and power handling, and increases the array's noise figure. More importantly, these array technologies are not fully printable and planar because they require elaborate, non-planar 3D "cable organizers" [20] between the ground plane and the printed element layers. These "cable organizers" are critical in preventing scan-induced resonances [21] by electrically shielding the vertical balanced feed lines. Additionally, all of these arrays require external wideband baluns/hybrids at each element.

The need for wideband external baluns poses a non-trivial problem. Wideband passive baluns do exist, e.g., Marchand/ compensated baluns, [22], but are electrically large, [23], and have high insertion loss. Conversely, active baluns can readily achieve wide bandwidths, but are unidirectional, noisy, and have low power-handling that limits their use to receive-only applications. Moreover, non-zero phase and amplitude imbalances, [24], finite common-mode rejection ratios (CMRR), [25], and low active gains (< 0 dB) at high frequencies are some other practical active balun limitations.

Unlike other UWB arrays, the planar ultrawideband modular array (PUMA) is fabricated with planar etched circuits and plated vias, thus it can be fabricated as a simple multilayer microwave PCB, and does not require external baluns. In addition, the array has a low profile (depth $\approx \lambda_{high}/3$) and can be constructed modularly. The array consists of a dual-offset dual-polarized version of Munk's tightly-coupled dipoles above a ground plane, fed by a novel unbalanced feed-line scheme. The PUMA [26] has shorting vias at its dipole arms, enabling direct connection to standard RF interfaces and modular construction. The placement of the plated vias controls the frequency of a catastrophic common mode that would otherwise occur near mid-band since the array is fed unbalanced. This topology modification, along with the dual-offset dual-polarized arrangement, gives rise to new phenomenology that is explained using simple physical models and equivalent circuits. Thus far, the PUMA array has demonstrated low VSWR and good scan performance out to $\theta = 45^{\circ}$ over a bandwidth of 3:1 when fed directly from a 50Ω unbalanced interface [27], [28]. When the array is used in conjunction with a specially designed planar matching network printed on the opposite side of the ground plane, it achieves a 5:1 bandwidth with VSWR < 2.1 at broadside, VSWR < 2.9out to 45° scan in the H- plane, and approximately -15 dBcross-polarization at $\theta = 45^{\circ}$ in the D- plane.

The remainder of the paper is organized as follows. Section II describes the baseline structure of the PUMA array. Section III develops the theory, starting with an overview of Munk's current sheet array theory, and a detailed discussion on the commonmode and loop-mode resonances supported by the unbalanced fed dipoles. A simple method for controlling surface waves on the thick grounded dielectric layers is also discussed. Section IV presents the design of a dual-polarized infinite PUMA array that validates the wideband performance of this topology. The paper concludes in Section V.

II. THE PUMA ARRAY

Though the PUMA array can be implemented in either singleor dual-polarized versions, this paper focuses on the dual-polarized form. The radiating layer of the PUMA array is comprised of printed dipoles in a dual-offset dual-polarized lattice with strong capacitive coupling between cross-polarized elements, as shown in Fig. 1(a).

A thin layer of PTFE substrate material, ϵ_{r2} , is placed above and below the dipole layer, shown in Fig. 1(b), which provides



Fig. 1. The PUMA array topology. (a) Top view of dipole layer; (b) crosssectional view of a unit-cell, showing the location where a module split occurs; (c) isometric view of a $2 \times 2 \times 2$ PUMA module with exploded dielectric cover layers.

a printable substrate and increases the power handling of the array [29] by preventing dielectric breakdown between capacitor gaps. An electrically thick ($t \approx \lambda_{mid}/4$, where λ_{mid} is the wavelength at mid-band) cover layer with low permittivity $(1 \leq \epsilon_{r1} \leq 2.2)$ is used for wideband tuning and acts as a wide angle impedance matching (WAIM) layer. The bottom dielectric layer, ϵ_{r3} , with thickness $d \approx \lambda_{mid}/4$, is also a low permittivity $(1 \leq \epsilon_{r3} \leq 2.2)$ PTFE material that can support plated vias that form the feed lines and shorting posts, as discussed next.

Together, the dipole elements, ground plane, and dielectric layers provide wideband performance, based upon the current sheet principle, [12]. However, the feed and dipole arrangements of the PUMA array are unique, and are what chiefly differentiate this topology from all other implementations of the CSA. A cross-sectional view of a PUMA unit cell is depicted in Fig. 1(b), which shows the unbalanced feed and the shorting vias. This fundamental change in topology allows unbalanced feed lines to be utilized without exciting the catastrophic common-mode resonance found in 2 D unbalanced fed arrays [11]. More importantly, this feeding method avoids "cable organizers," since the unbalanced feed lines do not support the scan-induced common-modes typical of balanced fed arrays. This allows the entire array (radiating elements and feed lines) to be fabricated as a single microwave multilayer PCB, with the feed lines and shorting posts implemented as plated vias, shown in Fig. 1(c). Also, the unbalanced feed lines readily connect to standard 50 Ω interfaces (coax, stripline, microstrip, CPW, etc.) without an external balun. An additional advantage derived from the unbalanced feed arrangement and the dual-offset, dual-polarized offset (egg-crate) lattice is modularity. As shown in Figs. 1(b) and 2, array modules can be formed by

intersecting planes passing between the feed line vias, therefore a PUMA array can be built and assembled modularly.

III. THEORY

The PUMA array's wideband performance depends on careful tuning of its complex structure, and design is carried out using full-wave numerical analysis; nevertheless, key physical models are presented to develop insights and design strategies.

A. The Current Sheet Concept

As early as 1968, Staiman, [30], recognized that tightly spaced dipole elements placed over a ground plane exhibit wideband impedance characteristics. In 2003, Munk, [12], exploited this concept to develop the wideband CSA array, using tightly-coupled printed horizontal dipoles arranged over a ground plane. Munk developed an insightful transmission line model, [31], illustrating that the capacitive impedance of the dipole layer is compensated by the inductive reactance of the ground plane, resulting in a wideband impedance. Placing a dielectric layer in front of the array is shown to further improve the impedance match. Part of the PUMA array design relies on both the groundplane and dielectric cover compensation principles. Since the main innovation of this work is not the design principle of tightly coupled arrays, the interested reader is referred to [31]. However, in an attempt to show the challenges associated with the feeding of tightly-coupled dipole arrays above a ground plane, e.g., PUMA, a simplified analysis from [29] is used to show that the resistance is much higher than 50Ω .

Consider an infinite, single-polarized dipole array placed a distance d above a ground plane, as shown in Fig. 3. The dipoles reside in vacuum ($\epsilon_r = 1$), and are fed by ideal delta-gap sources at their center. The active resistance of the array in Fig. 3 is, [29],

$$R_{A} = \frac{\eta_{\circ}}{2D_{x}D_{y}} \frac{1}{\cos\theta} [P_{\perp}T_{\perp}P_{\perp}^{*} + P_{\parallel}T_{\parallel}P_{\parallel}^{*}]$$
(1)

where η_{\circ} is the free-space wave impedance, and D_x and D_y are the E- and H-plane element spacings. $P_{\perp} = \hat{\mathbf{y}} \cdot P\hat{\mathbf{n}}_{\perp}$ and $P_{\parallel} = \hat{\mathbf{y}} \cdot P\hat{\mathbf{n}}_{\parallel}$ are the normal and parallel scalar pattern function components (referenced to the scan plane), and T_{\perp} and T_{\parallel} are the associated transmission coefficient terms that account for the presence of the ground plane. The scalar pattern function is given by $P = (1/I(0)) \int_{-L/2}^{L/2} I(y') e^{jk_{\circ}y' \sin\theta \sin\phi} dy'$, where I(y') is the current along the dipole length L. The current on closely-spaced, tightly-coupled dipole elements is approximately constant, as observed by Hansen, [32], thus I(y') = 1. The transmission coefficient terms are two-element array factors consisting of the excited dipole at z = d and its image located at z = -d, where $T_{\perp} = T_{\parallel} = 1 - e^{-j2dk_{\circ}\cos\theta}$. The resulting input resistance at broadside ($\phi = 0^{\circ}, \theta = 0^{\circ}$) becomes

$$R_A = \frac{\eta_o}{2D_x D_y} L^2 (1 - e^{-j2dk_o}).$$
 (2)



Fig. 2. Top view of the dual-offset, dual-polarized (egg-crate) PUMA array lattice, showing two of the many possible module sizes, $1 \times 1 \times 2$ and $2 \times 2 \times 2$, where the module split locations are shown dashed. Circles indicate feed line vias.



Fig. 3. Sketch of infinite dipole array located in the x-y plane, above a ground plane, scanned in the $\hat{\mathbf{k}}_{00}$ direction.



Fig. 4. Broadside resistance of a single-pol dipole array spaced $d = \lambda/4$ above a ground plane versus periodicity, with $L = D_y$. The element current is assumed to be constant.

In Fig. 4, R_A is plotted versus D_x for various D_y values and for $d = \lambda/4$. For most wideband arrays at mid-band, $D_x = D_y = \lambda/4$, leading to $R_A \approx 377\Omega$. This high resistance level poses a major challenge when matching to standard (unbalanced) 50 Ω interfaces.

B. Feeding

A successful feeding mechanism for a tightly-coupled dipole array above a ground plane should be able to connect the dipole arms to a feed network or connectors at the back side of the ground plane and transform the high dipole impedance to 50Ω .



Fig. 5. Feeding approaches for tightly-coupled dipole arrays (only one unit cell is shown). (a) Balanced feeding, with 3D cable organizer; (b) unbalanced feeding.



Fig. 6. Typical broadside VSWR of single-pol, tightly-coupled dipole arrays with balanced and unbalanced feed arrangements (as shown in Fig. 5). Neither design is optimized for impedance match or bandwidth.

This challenging task could be achieved with either balanced or unbalanced vertical feed lines, as shown in Fig. 5.

1) Balanced Vs. Unbalanced Excitation: Exciting the dipoles with a vertical balanced line requires an external balun at each port, as shown in Fig. 5(a). The balun forces a 180° phase difference between currents and voltages on each line, with the ground plane acting as a third reference conductor with 0V. Assuming ideal balun operation and broadside array excitation, this arrangement results in resonance-free performance over a wide bandwidth, as shown in Fig. 6 (it is noted that ideally fed (e.g., delta gap) tightly-coupled dipole arrays could achieve bandwidths up to 4:1 with broadside VSWR < 2, [31], though this example is not optimized). However, when the array is scanned along the E- plane the push-pull currents become unbalanced (due to mutual coupling), resulting in scan-induced anomalies if the feed lines are not appropriately shielded [21]. Practically, the shielding of these lines is achieved using a vertical "cable organizer" [20]. Cable organizers are 3D metallic structures that require machining, assembly, and soldering, rendering the array non-planar. More importantly, this feeding approach is not easily scaled to very high frequencies.

Unbalanced excitation is accomplished using the feed shown in Fig. 5(b), where one dipole arm is connected to the ground with a vertical line, and the other dipole arm is excited directly with the inner conductor of a coax line. This results in unequal magnitude currents on the two feed lines, producing a net vertically polarized current. This net current excites a problematic resonance, as shown in Fig. 6, referred to as a common-mode resonance in Section IV.

2) Common-Mode on Unbalanced Fed Tightly-Coupled Dipole Arrays: When driving the dipoles with unbalanced feeds, the net vertical current distribution couples into a resonance that results in a short circuit in the input impedance of the elements. This resonance is referred to throughout the rest of this paper as the common-mode resonance and occurs near mid-band, at the frequency f_{cm} (see [33] for an in depth analysis of the common-mode). This resonance is unique to unbalanced-fed arrays, and is different from the common-mode issues reported in certain types of balanced-fed arrays, [34].

Away from f_{cm} , there are no vertically polarized (\hat{z} -polarized, normal to the ground plane) electric fields present in an array unit cell. Conversely, observing the fields at f_{cm} reveals a strong vertical electric field distribution with a cosine amplitude variation along the diagonal feed-to-feed direction, with nulls at the grounded line of each feed, as shown in Fig. 7. This field distribution is observed along both diagonal planes, and forms a distribution as shown in the overhead view of Fig. 8, which is a contour map sketch of the vertically-polarized electric field amplitude. The common-mode resonance occurs when the diagonal plane path length between shorted lines is equal to half a wavelength, i.e., $L = \lambda_{cm}/2 = \sqrt{D_x^2 + D_y^2}$, where D_x an D_y are the E- and H- plane spacings respectively, leading to the common-mode frequency

$$f_{cm} \approx \frac{c_{\circ}}{2\sqrt{\epsilon_{r3}}\sqrt{D_x^2 + D_y^2}} \tag{3}$$

where c_{\circ} is the speed of light, and ϵ_{r3} is the bottom dielectric layer, shown in Fig. 1(b). Equation (3) shows that f_{cm} will always occur below the grating lobe onset frequency, inside the operating band.

To validate (3), Table I shows the analytic predictions of f_{cm} compared with results obtained using Ansoft/Ansys HFSS, [35], for a single-polarized (along $\hat{\mathbf{x}}$) unbalanced-fed tightly-coupled dipole array for various D_x , D_y , and ϵ_{r3} , with d = 5 mm ($\approx \lambda_{g,mid}/4$, where $\lambda_{g,mid}$ is the guided wavelength in dielectric layer 3). The $\hat{\mathbf{z}}$ -polarized E- fields at f_{cm} are mostly confined to dielectric layer 3, thus layers 1 and 2 have negligible effect on f_{cm} and are set to vacuum for these simulations ($\epsilon_{r1} = \epsilon_{r2} = 1$). Good agreement is shown, with error less than 7.60%.

Equation (3) implies strategies to tune f_{cm} out of the band. A very large element spacing, D_x , D_y , can move f_{cm} below the lowest frequency of the band, but f_g (the grating lobe onset frequency when scanned to the horizon, $\lambda_g = \max\{2D_x, 2D_y\}$) will also move to the low end of the band, resulting in a narrow usable bandwidth. Alternatively, very small element spacing can move f_{cm} above the operating band, but this results in an over-populated aperture that unnecessarily increases the



Fig. 7. Electric field distribution at the common-mode frequency f_{cm} in an unbalanced fed tightly-coupled dipole array.



Fig. 8. Overhead view of the \hat{z} -polarized electric field distribution E_z at the common-mode frequency f_{cm} in an unbalanced fed tightly-coupled dipole array.

TABLE I COMPARISON OF COMMON-MODE RESONANCE THEORY WITH NUMERICAL SIMULATIONS

D_x	D_y	ϵ_{r3}	Analytic	Numerical	Error
[mm]	[mm]		f_{cm} (3)	f_{cm}	%
9	5	1	14.57	13.54	7.60
7	5	1	17.44	16.41	6.26
5	5	1	21.21	20.89	1.55
5	5	1.96	15.15	15.18	0.18
7	5	1.96	12.46	12.08	3.11
7	9	1.96	9.40	9.90	5.08
10	7	1.96	8.78	9.08	3.33
5	5	2.2	14.30	14.41	0.75
5	5	5	9.49	9.73	2.50

number of T/R modules, and makes connectorization difficult. Finally, increasing ϵ_{r3} reduces f_{cm} only marginally, due to the inverse square root relationship. These limited options are thus not practical solutions for controlling the common-mode resonance.

3) Common-Mode Mitigation in PUMA Arrays: Since the resonant fields are polarized normal to the ground plane, orthogonal to either of the desired transmitting polarizations,



Fig. 9. Top view of a PUMA unit cell, showing new resonant length L due to the shorting posts.

 TABLE II

 COMMON-MODE FREQUENCY CONTROL WITH SHORTING POSTS

8	Analytic	Numerical	Error
[mm]	f_{cm} (4)	f_{cm}	%
(no shorts)	12.08	12.46	3.15
0.75	14.41	14.01	2.81
1	15.15	15.02	0.87
1.25	15.93	16.07	0.89
1.5	16.73	17.06	1.95
1.75	17.55	18.01	2.59
2	18.37	18.90	2.86
2.25	19.17	19.43	1.38

shorting posts can be used to suppress the resonance, much like the shorting posts used to suppress board resonances in PCB design, without significantly disrupting the desired radiating fields.

The shorting posts are shown in Fig. 1, and an overhead view of the array in Fig. 9 shows the placement of the shorting posts along the dipole arms. In this structure (PUMA), the commonmode resonant frequency is

$$f_{cm} \approx \frac{c_{\circ}}{2\sqrt{\epsilon_{r3}}\sqrt{(D_x - 2s)^2 + D_y^2}} \tag{4}$$

where s is the shorting post separation from the center of the unit cell. From (4), one can shift f_{cm} up in frequency and out of band by increasing the shorting post separation s. The results of an infinite single-pol ($\hat{\mathbf{x}}$ -polarized) PUMA array simulation using Ansoft/Ansys HFSS, [35], having $\epsilon_{r3} = 1.96$, $\epsilon_{r1} = \epsilon_{r2} = 1$, $D_x = 7 \text{ mm} (f_g = 21.48 \text{ GHz}), D_y = 5 \text{ mm}$, and with the dipole layer located $d = 5 \text{ mm} (\approx \lambda_{g,mid}/4)$ above the ground plane, are shown in Table II. The full-wave numerical results show good agreement with the analytic predictions of (4), demonstrating less than 4% error for all cases.

Increasing the shorting post spacing s increases f_{cm} , until s = 2.25 mm, where f_{cm} is close to f_g . Thus the novel shorting post arrangement of the PUMA array enables common-mode free, wideband performance at broadside without external baluns or 3D metalized feed structures. As discussed in [33], when unbalanced fed dipole arrays are scanned the problematic common-mode disappears along the D -and H- planes, and is weakly excited along the E-plane, where f_{cm} increases with θ ; thus broadside suppression of the common-mode implies common-mode free scan operation in the PUMA.

C. Low Frequency Loop Resonance in PUMA

Despite alleviating the in-band common-mode problem, and displaying minor impact on impedance over most of the



Fig. 10. Impedance of an unbalanced fed Fig. 5(b) tightly-coupled single-polarized dipole array, without and with shorting posts (PUMA).



Fig. 11. Loop mode resonance model. (a) Current distribution on PUMA elements at the loop-mode resonance f_{loop} ; (b) current distribution at f_{loop} , using image theory to remove ground plane; (c) circuit model showing a small non-resonant loop driving a large resonant loop.

operating band, the shorting posts significantly alter the low-frequency impedance behavior of the array. To demonstrate this, the impedance of a PUMA and its no-short counterpart are plotted in Fig. 10. Without shorting posts, the impedance at the low frequency limit f_L is capacitive, and an extrapolation to DC leads to an open-circuit, suggesting a series resonance typical of dipoles. In contrast, the PUMA array exhibits an inductive low frequency impedance, with an eventual short-circuit impedance at DC, suggesting a parallel resonance typical

TABLE III Comparison of Loop-Mode Resonance Theory With Numerical Simulation

D_x	d	Analytic	Numerical	Error
[mm]	[mm]	f_{loop} (5)	f_{loop}	%
7	2.5	5.10	4.84	5.14
9	2.5	4.29	4.26	0.60
11	2.5	3.69	3.80	2.85
7	5	3.46	3.41	1.34
9	5	3.06	3.18	3.88
11	5	2.75	2.77	0.83
7	7	2.75	2.86	4.10
9	7	2.49	2.67	7.16
11	7	2.28	2.47	8.35

of loop antennas. A schematic of the current distribution at the first resonance, f_{loop} , is shown in Fig. 11(a).

Currents on the excited feed line flow in the opposite direction from those on its adjacent shorting post, and the currents on both feed lines flow in the same direction, forming a loop between elements. Since the structure is placed over a ground plane, image theory allows the full current loops to be revealed, as in Fig. 11(b). Two loops are formed by the circulating currents, where a small "driving loop" couples energy into a large resonant loop formed between the shorting posts of adjacent dipole arms, shown in the circuit model of Fig. 11(c). Using the circumference of the resonant loop, f_{loop} is found to be

$$f_{loop} \approx \frac{c_{\circ}}{2\sqrt{\epsilon_{r3}}(4d + 2D_x - 3s)} - |f(C)| \tag{5}$$

where |f(C)| represents the effect of the capacitive loading due to inter-element capacitance C. The dominant terms of (5) are D_x and d, and Table III demonstrates their effect on f_{loop} , using the same simulation model parameters as in Section III-B.3, and with s = 1 mm. The predicted values agree with full-wave numerical values with less than 8.35% error. (Note that the analytic values in Table III assume |f(C)| = 0; a properly chosen |f(C)| may further reduce the error levels).

From (5), reducing f_{loop} (wider bandwidth) requires minimal short spacing s, and maximal capacitive coupling, leading to an interesting compromise with (4). As shown in Section IV-C, removing the shorting post on the grounded arm increases the loop size sufficiently to move f_{loop} well below the operating band, while maintaining an f_{cm} above the operating band.

D. Surface Wave Mitigation in PUMA

Tightly-coupled dipole arrays such as the CSA and PUMA exhibit a broader frequency response and avoid surface waves when the dielectric between the dipoles and ground is air ($\epsilon_r =$ 1). In the case of the PUMA, such a choice would make fabrication complicated and difficult to realize above X-band, since the shorting posts and feed lines require mechanical support. It is thus desirable to design the PUMA array using PTFE substrates to allow simple fabrication of the feed lines and shorting posts as plated vias. Unfortunately, this results in a thick ($d \approx \lambda/2$ @ f_{high}) grounded dielectric layer that supports surface



Fig. 12. Perforated dielectric arrangement. (a) Top view of substrate, showing spacing D_x , D_y and radius, R, of the holes in the dielectric; (b) top and side view of a single-pol, tightly-coupled dipole array on a perforated dielectric.

waves at certain scan angles, resulting in "scan blindnesses" inside the desired scan volume. The location of such scan blindnesses can be approximately predicted in u, v space, by, [36]

$$\left(\frac{\beta_{sw}}{k_{\circ}}\right)^{2} = \left(\frac{m}{\frac{D_{x}}{\lambda}} + u\right)^{2} + \left(\frac{n}{\frac{D_{y}}{\lambda}} + v\right)^{2} \tag{6}$$

where $\beta_{sw} \geq k_o$ is the grounded dielectric surface wave propagation constant, k_o is the free space wavenumber, and m and n are integers. Since β_{sw} increases with ϵ_r and d in a PUMA array, the scan blindness angle θ_{sw} could potentially move inside the usable scan volume, posing a severe limitation. Since dis an important design parameter in achieving wide bandwidth, [31], the only option is to reduce ϵ_r . Currently, the lowest ϵ_r for PTFE materials is $\epsilon_r = 1.96$ (Rogers 5880LZ), but even that is not sufficient to move surface waves out of the $\theta = 45^\circ$ scan cone.

The effective permittivity can be further reduced by perforating the dielectric layers with a series of holes in the region between the dipoles. The effective permittivity, $\epsilon_{r,eff}$, is reduced by the volume averaging of the permittivities of the dielectric and air-filled holes, and is given (based on a derivation similar to [37]) by

$$\epsilon_{r,eff} = \epsilon_{r3} - \frac{\pi R^2}{D_x D_y} (\epsilon_{r3} - 1) \tag{7}$$

where R is the radius of the holes in a rectangular grid as shown in Fig. 12(a). The effectiveness of the holes is demonstrated using a simple single-polarized, tightly-coupled dipole array placed between two dielectrics, of thicknesses $h_1 = h_2 =$ 3.175 mm with $\epsilon_r 1 = \epsilon_r 3 = 2.2$, and backed by a ground plane, shown in Fig. 12. The ideal, gap-fed dipoles are of length $\alpha = 6.6 \text{ mm}$ and width $\beta = 1.2 \text{ mm}$, with spacing $D_x =$ $D_y = 6.8 \text{ mm}$. The array was analyzed at f = 21.5 GHzas a doubly-periodic, infinite array using Ansoft/Ansys HFSS, [35], for scanning in the E-plane. The hole radius is varied from R = 0 - 2.5 mm. The results in Fig. 13 show the unperforated substrate dipole array has a scan blindness at $\theta_{sw} = 39^{\circ}$, while the dielectrics with holes of increasing radius lead to progressively larger scan blindness angles. At R = 2.5 mm, the



Fig. 13. Normalized reflection coefficient, $|\Gamma|$, variation versus θ for E-plane scan angles at f = 21.5 GHz and various perforated dielectric hole radii. All curves assume the array is conjugately matched to the source impedance at broadside.

TABLE IV SCAN BLINDNESS ANGLES: THEORY VS. NUMERICAL SIMULATIONS

		Theory ((6)&(7))	Numerical (HFSS)
$R \ [mm]$	$\epsilon_{r,eff}$	θ_{SW} [deg]	θ_{SW} [deg]
0.00	2.20	39.7	39.3
0.50	2.17	40.4	39.8
1.50	2.02	44.5	43.7
2.50	1.69	54.8	53.8

surface wave occurs at $\theta_{sw} = 54^{\circ}$. These full-wave results are compared with the theoretical predictions from (6) and (7) in Table IV, and the results are in excellent agreement. Experience has shown that this method can move θ_{sw} by approximately $10-20^{\circ}$.

IV. DUAL-POLARIZED INFINITE ARRAY DESIGN

A dual-polarized PUMA array design is presented to provide a representative example of how the theory and insights presented thus far can be applied to achieve a 5:1 PUMA array with good scan performance out to $\theta = 45^{\circ}$ in all planes.

A. Array Geometry

The overall array arrangement is shown in Fig. 1, where the elements are arranged in a dual-pol, dual-offset (egg-crate) lattice. As arranged in Fig. 1(a), the interdigited capacitors form a rhombic shaped "loop". When large capacitance is needed to extend the low frequency limit of the operating band, the mean circumference of this "loop" becomes approximately half a wavelength, and a new resonance appears when the array is scanned in the H- plane. To concurrently maximize capacitance and avoid this scan resonance, the proposed PUMA design places each polarization on a separate dielectric layer, with parallel plate capacitors formed by the overlapping portions of orthogonally polarized dipole arms, as shown in Fig. 14(a). As elucidated in Section IV-C, the shorting post on the grounded dipole arm is removed to further enhance the low frequency performance. Dielectric layer 3 is perforated with cylindrical holes of radius R = 10 mm and is comprised of Rogers 5880LZ $(\epsilon_r = 1.96)$, while Dielectric layer 1 is an un-perforated layer



Fig. 14. A 5:1 dual-pol PUMA array. (a) Top view of unit cell; (b) side view; (c) bottom view of the matching network; (d) circuit model of the matching network.

of Rogers 5880 ($\epsilon_r = 2.2$). Dielectric layers 2 and 4 are thin dielectric bonding layers (Gore SPEEDBOARD C Prepreg bond layer), with $\epsilon_r = 2.6$, and serve as both a means of joining the dielectric layers, as well as a useful dielectric substrate, which is critical to a backplane matching network, described in Section IV.

B. Backplane Matching Network

Due to the high impedance of the dipole aperture layer $(Z_{in} \approx 377\Omega)$, obtaining an array impedance well matched to $Z_o = 50\Omega$ can be accomplished in two ways. The first is to carefully design the array and vertical feed lines to act as impedance transformer sections to provide matching to 50Ω at the ground plane interface; this method is typically limited to $\approx 3:1$ bandwidths, [27], [28]. The second option is to use an impedance transformer below the ground plane. Instead



Fig. 15. Full-wave analysis of array and matching network for broadside scan, showing impedance loci for: the array impedance seen at the ground plane, the impedance seen after the capacitor, and finally the impedance seen at the end of the impedance transformer. All results referenced to $Z_{\circ} = 50\Omega$.

of tuning the array to $Z_o = 50\Omega$, the array is tuned to have a specific "mismatched" impedance profile that, when fed through the backplane matching network, results in a wideband impedance well-matched to 50Ω . The matching network employed in this design is inspired by a classic wideband impedance matching concept, which uses the combination of a series open-circuit stub and an impedance transformer [38]. However, the proposed matching network avoids the open stub with a substantially smaller series LC network, formed by a parallel plate capacitor and high impedance T-line.

The array backplane (PCB 2 in Fig. 14(b) and (c)) contains a printed matching network, comprised of a series capacitor, series inductor, and a transformer section. The capacitor is formed by a pair of circular plates of radius C_R , which are separated by the thin dielectric layer 4. One plate is attached to the end of the feed line plated via, and the other plate is attached to a narrow, high impedance ($\approx 80\Omega$) microstrip feed line of length L_{t1} and width W_{t1} that approximates an inductor. This is followed by a microstrip line of width W_{t2} , with an impedance greater than 50Ω , and has a length L_{t2} .

The circuit schematic of this structure is shown in Fig. 14(d). The series capacitor, C, introduces a reactance $jX_C = 1/j\omega C$ that partially cancels out the inductive reactance of the antenna; additionally, two parasitic shunt capacitors are shown dashed in Fig. 14(d), representing the parasitic capacitances between each plate and the ground plane, though these have second-order impact on performance. Next, a series inductor introduces the reactance $jX_L = j\omega L$ that forms a series LC network with the capacitor. Finally, a transmission line with length L_{t2} and propagation constant $\beta_2 = \sqrt{\epsilon_{r4}k_o}$ transforms the impedance to $Z_{in} = Z_{t2}(Z_L + jZ_{t2} \tan(\beta_2 L_{t2}))/(Z_{t2} + jZ_L \tan(\beta_2 L_{t2}))$, where $Z_L = Z_{ant} + j(X_C + X_L)$, as shown in the circuit schematic of Fig. 14(d). In the schematic, Z_{ant} is the antenna impedance at the ground plane (see Fig. 14(b)).

The entire structure, array and backplane printed matching network, were analyzed full-wave (no approximate T-line models) using Ansoft HFSS, [35], and the results are shown in Fig. 15. At broadside, the antenna impedance Z_{ant} is purposefully tuned to be inductive over most of the operating band, with a poor match to 50 Ω ; however, the combination of



Fig. 16. Comparison of the port coupling (H to V) for an optimized PUMA array with one and two shorting posts.

 $Z_L = Z_{ant} + jX_c$ is shown to be centered near $Z = 75\Omega$, though strongly capacitive. By adding the inductor reactance $jX_L = j \ \omega L$ in series, the entire locus is moved toward the inductive part of the Smith chart (though this step is not shown in order increase the readability of Fig. 15). Finally, the impedance locus $Z_L + jX_L$ is matched to 50Ω by passing through a quarter wave transformer with $Z_t \approx 50 - 100\Omega$ $(Z_t = 65\Omega$ for this design). The final impedance is shown to be well matched, with a VSWR < 2.1 over the band. This simple, planar matching network is used on all ports, and readily fits inside a unit cell with meandering.

As shown in Fig. 14(b), the array can be fabricated as one PCB, labeled "PCB 1," and the backplane matching network is fabricated on a separate PCB, labeled "PCB 2," these PCB's are then joined by a thick bonding layer, dielectric layer 4, without any conductive electrical connection, further reducing the fabrication complexity.

C. Shorting Post Placement

As noted in Section IV-A, this design uses only one shorting post, attached to the non-grounded dipole arm. The loop mode resonance, discussed in Section III-C, can be shifted to lower frequencies by maximizing the resonant loop size—accomplished here by removing one shorting post. The Γ_a of the optimized PUMA with one and two shorting posts behaves very similarly above mid-band. However, the cross-polarized coupling C_{hv} (H-V pol coupling) is quite different. As shown in Fig. 16, the double-short PUMA has strong coupling near 1 GHz, due to the presence of the loop mode. Therefore, using only one shorting post increases impedance bandwidth and reduces cross-polarization coupling.

D. Numerical Results

The array and the backplane matching network were simulated in an infinite array environment using Ansoft/Ansys HFSS, [35], assuming PEC for vias and metal layers, and standard dielectric models for layers 2 and 4 (Gore SPEED-BOARD C Prepreg bond layer) and layers 1 (Rogers 5880) and 3 (Rogers 5880LZ). All results are referenced to $Z_{\circ} = 50\Omega$.

1) Impedance Vs. Scan: Fig. 17 shows the VSWR for various E-plane and H-plane scan angles out to 45°. These results are



Fig. 17. VSWR vs. frequency and scan angle of the infinite dual-polarized PUMA array. (a) E-plane; and (b) H-plane. The D-plane impedance (not shown here) is approximately the average of the two.

calculated by exciting one polarization and terminating the other in 50 Ω . As discussed in Section IV-C, the cross-polarization coupling is very low, below -20 dB over the band, therefore this is a valid means of evaluating performance. The vertical lines in the plot indicate the band edges (1.06 GHz and 5.3 GHz) of this array.

Fig. 17(a) shows a maximum broadside VSWR of approximately 2.1, with a VSWR less than 2 over most of the band. A maximum VSWR of 2.5 is observed at the high frequency band edge for $\theta = 45^{\circ}$ scanning in the E-plane. The H-plane results are shown in Fig. 17(b), indicating a maximum VSWR of 2.9 at $\theta = 45^{\circ}$ at the low end of the band, which is a typical rise for dipole arrays. The VSWR in each of these planes is seen to vary little with scan, a benefit of the array's low-profile and WAIM layer. The D-plane results are omitted since they follow an approximate average of the E- and H-planes.

2) Cross-Polarization Vs. Scan: A dual-polarized infinite array can radiate arbitrarily polarized plane waves, thus the radiated power can be decomposed into two orthogonally polarized plane waves. For this analysis, the polarizations are chosen according to Ludwig's 3rd definition of cross-pol [39]. A surface *S* is placed parallel to the ground plane above the array, and the power flowing through this surface is calculated by integrating the Poynting vector of each plane wave over a unit cell. The coand cross-polarization radiated powers per unit cell are shown in Fig. 18 over the frequency band for scanning in the E- and D-planes; all power levels are plotted in dB and are normalized to the incident power at the input port, and therefore include mismatch, dielectric, and back-radiation losses.



Fig. 18. Co- and cross-polarization (Ludwig's 3rd definition) radiated power vs. frequency and scan angle of the infinite dual-polarized PUMA array. (a) E-plane; and (b) D-plane. The H-plane polarization levels (not shown here) are approximately the same as the E-plane.

The co-polarized power in the E-plane, Fig. 18(a), is nearly 0 dB throughout the band, exhibiting minor decreases only due to mismatch loss, indicating good efficiency. The cross-polarized power is below -20 dB over the majority of the band, and reaches a maximum of -7 dB at the low-end band edge. Overall, the cross-polarization remains very low even at wide scan angles. The H-plane results are omitted, since they are very similar (or better) than the E-plane.

For all angles, the D- plane co-polarized power, Fig. 18(b), is very similar to that of the E-plane at broadside. The cross-polarized power is shown to be below -20 dB over the full band for angles out to $\theta = 30^{\circ}$, and reaches a maximum of -15 dB at $\theta = 45^{\circ}$. The low-profile of the PUMA array helps maintain very low cross-polarization throughout the band.

Finally, to demonstrate that the back-radiated power from the matching network is very small, Table VI shows the ratio of the back-radiated power, P_{back} , to the input power, P_{in} , for various frequencies. The array is excited at broadside. The back radiation loss is very small (< 1%), and as expected, it increases slightly with frequency. This behavior is approximately the same for other scan angles as well. It is worth noting that throughout the band the radiation efficiency (including mismatch, port coupling, conductor, dielectric and back-radiation losses) of this infinite PUMA array was predicted to be above 90%.

V. CONCLUSION

The proposed PUMA array offers wideband and wide-scan performance in a simple, modular, truly planar architecture that

 TABLE V

 5:1 Dual-Pol PUMA Array Dimensions

D_x [mm]	D_y [mm]	C_L [mm]	C_W [mm]
24.50	24.50	2.10	1.75
t_1 [mm]	$t_{\mathscr{Q}}$ [mm]	$t_{\mathcal{J}}$ [mm]	t_4 [mm]
11.80	0.127 (5mil)	14.00	0.127 (5mil)
ϵ_{r1}	ϵ_{r2}	ϵ_{r3}	ϵ_{r4}
1.96	2.6	1.96	2.6
L_1 [mm]	L_2 [mm]	W_1 [mm]	W_2 [mm]
5.60	9.80	3.50	2.50
C_R [mm]	g [mm]	S [mm]	F_g [mm]
1.30	0.50	4.10	1.50
L_{t1} [mm]	$L_{t\mathcal{Z}}$ [mm]	W_{t1} [mm]	$W_{t\mathcal{Z}}$ [mm
6.00	17.30	0.10	0.20
R_f [mm]	R_s [mm]		
0.88	1.30		

 TABLE VI

 BACK-RADIATION LOSS OF THE 5:1 PUMA

f [GHz]	1.50	2.00	2.50	3.00	3.50	4.00	4.50	5.00
$\frac{P_{back}}{P_{in}}$ [%]	0.10	0.20	0.20	0.20	0.30	0.40	0.50	0.70

allows both the array aperture and feeds to be printed using standard microwave fabrication techniques, which results in lowcost and frequency scalable manufacturing. Additionally, the unbalanced feed arrangement removes the necessity for external wideband baluns and cable organizers. The enabling feature is the introduction of shorting posts, which are shown to eliminate a catastrophic resonance due to unbalanced feeding of balanced dipoles. Simple circuit models are presented that provide insight into the shorting post operation, including control of a new low frequency loop mode. The PUMA array uses a thick grounded substrate that supports surface waves, and a method to control the associated scan blindnesses was proposed. Finally, a dual-polarized PUMA array design with a 5:1 bandwidth with VSWR < 2.1 at broadside, and approximately -15 dB cross-polarization out to 45° scan in the D-plane was shown.

REFERENCES

- G. C. Tavik, C. L. Hilterbrick, J. B. Evins, J. J. Alter, J. G. Crnkovich, J. W. de Graaf, W. Habicht, G. P. Hrin, S. A. Lessin, D. C. Wu, and S. M. Hagewood, "The advanced multifunction RF concept," *IEEE Trans. Microw. Theory Tech.*, vol. 53, no. 3, pp. 1009–1020, Mar. 2005.
- [2] L. R. Lewis, M. Fassett, and J. Hunt, "A broadband stripline array element," presented at the Antenna Applications Symp., Allerton Park, Monticello, IL, Sep. 1974.
- [3] M. Stasiowski and D. H. Schaubert, "Broadband array antenna," presented at the Antenna Applications Symp., Allerton Park, Monticello, IL, Sep. 2008.
- [4] D. H. Schaubert, "A gap-induced element resonance in single-polarized arrays of notch antennas," in *Proc. IEEE Antennas and Propagation Society Int. Symp.*, Jul. 1994, vol. 2, pp. 1264–1267.
- [5] H. Holter, "Dual-polarized broadband array antenna with BOR-elements, mechanical design and measurements," *IEEE Trans. Antennas Propag.*, vol. 55, no. 2, Feb. 2007.
- [6] E. W. Lucas, M. A. Mongilio, K. M. Leader, C. P. Stieneke, and J. W. Cassen, "Notch radiator elements," U.S. patent 5 175 560, Dec. 29, 1992.
- [7] J. J. Lee, S. Livingston, and R. Koenig, "A low-profile wide-band (5:1) dual-pol array," *IEEE Antennas Wireless Propag. Lett.*, vol. 2, pp. 46–49, 2003.
- [8] J. D. S. Langley, P. S. Hall, and P. Newham, "Balanced antipodal Vivaldi antenna for wide bandwidth phased arrays," *Inst. Elect. Eng. Proc.-Microw. Antennas Propag.*, vol. 143, no. 2, pp. 97–102, Apr. 1996.

- [9] M. Elsallal, "Doubly-Mirrored balanced antipodal Vivaldi antenna (Dm-BAVA) for high performance arrays of electrically short, modular elements," Ph.D. dissertation, Dept. Elect. Comput. Engirg., Univ. Massachusett, Amherst, 2007.
- [10] S. S. Holland, M. N. Vouvakis, and D. H. Schaubert, "The Banyan Tree Antenna Array," U.S. patent application 61/230,768, Aug. 3, 2009.
- [11] S. S. Holland, M. N. Vouvakis, and D. H. Schaubert, "A new modular wideband array topology," presented at the Antenna Applications Symp., Allerton Park, Monticello, IL, Sep. 2009.
- [12] B. Munk et al., "A low-profile broadband phased array antenna," in Proc. IEEE Antennas and Propagation Society Int. Symp., Jun. 2003, vol. 2, pp. 448–451.
- [13] H. A. Wheeler, "Simple relations derived from a phased-array antenna made of an infinite current sheet," *IEEE Trans. Antennas Propag.*, vol. 13, no. 4, pp. 506–514, 1965.
- [14] J. G. Maloney, M. P. Kesler, P. H. Harms, and G. S. Smith, "Fragmented aperture antennas and broadband antenna ground planes," U.S. 6,323,809, May 26, 2000.
- [15] P. Friederich, L. Pringle, L. Fountain, and P. Harms, "A new class of broadband planar apertures," presented at the Antenna Applications Symp., Allerton Park, Monticello, IL, Sep. 2001.
- [16] A. Neto and J. J. Lee, "Ultrawideband properties of long slot arrays," *IEEE Trans. Antennas Propag.*, vol. 54, no. 2, pp. 534–543, Feb. 2006.
- [17] J. J. Lee, S. Livingston, R. Koenig, D. Nagata, and L. L. Lai, "Compact light weight UHF arrays using long slot apertures," *IEEE Trans. Antennas Propag.*, vol. 54, no. 7, pp. 2009–2015, Jul. 2006.
- [18] W. Croswell, T. Durham, M. Jones, D. Schaubert, P. Friederich, and J. Maloney, "Wideband antenna arrays," in *Modern Antenna Handbook*, C. A. Balanis, Ed. Hoboken, NJ: Wiley, 2008.
- [19] J. J. Lee, S. Livingston, and D. Nagata, "A low profile 10:1 (200–2000 MHz) wide band long slot array," in *Proc. IEEE Antennas and Propagation Society Int. Symp.*, Jul. 5–11, 2008, pp. 1–4.
- [20] M. Jones and J. Rawnick, "A new approach to broadband array design using tightly coupled elements," in *Proc. IEEE Military Communications Conf.*, Oct. 2007, pp. 1–7.
- [21] J. R. Bayard, D. H. Schaubert, and M. E. Cooley, "E-plane scan performance of infinite arrays of dipoles printed on protruding dielectric substrates: Coplanar feed line and E-plane metallic wall effects," *IEEE Trans. Antennas Propag.*, vol. 41, no. 6, pp. 837–841, Jun. 1993.
- [22] G. Oltman, "A compensated balun," *IEEE Trans. Microw. Theory Tech.*, vol. 14, no. 3, pp. 112–119, Mar. 1966.
- [23] J.-S. Lim, U.-H. Park, S. Oh, J.-J. Koo, Y.-C. Jeong, and D. Ahn, "A 800-to 3200-MHz wideband CPW balun using multistage Wilkinson structure," in *IEEE Microwave Symp. Digest*, Jun. 2006, pp. 1141–1144.
- [24] K. Jung, W. R. Eisenstadt, R. Fox, A. Ogden, and J. Yoon, "Broadband active balun using combined cascode-cascade configuration," *IEEE Trans. Microw. Theory Tech.*, vol. 56, no. 8, pp. 1790–1796, Aug. 2008.
- [25] M. Ferndahl and H.-O. Vickes, "The matrix balun-a transistor-based module for broadband applications," *IEEE Trans. Microw. Theory Tech.*, vol. 57, no. 1, pp. 53–60, Jan. 2009.
- [26] S. S. Holland and M. N. Vouvakis, "The planar ultrawideband modular antenna (PUMA) array," U.S. Application 61/230,271, Jul. 31, 2009.
- [27] S. S. Holland and M. N. Vouvakis, "A 7–21 GHz planar ultrawideband modular antenna array," in *Proc. IEEE Antennas and Propagation Society Int. Symp.*, Toronto, ON, Jul. 11–17, 2010, pp. 1–4.
- [28] S. S. Holland and M. N. Vouvakis, "A fully planar ultrawideband array," in *Proc. Antenna Applications Symp.*, Allerton Park, Monticello, IL, Sep. 21–23, 2010, pp. 221–232.
- [29] B. Munk, Frequency Selective Surfaces, Theory and Design. New York: Wiley, 2000.

- [30] D. Staiman, M. Breese, and W. Patton, "New technique for combining solid-state sources," *IEEE J. Solid-State Circuits*, vol. Sc-3, no. 3, pp. 238–243, Sep. 1968.
- [31] B. Munk, Finite Antenna Arrays and FSS. New York: Wiley, 2003.
- [32] R. C. Hansen, "Linear connected arrays," IEEE Antennas Wireless Propag. Lett., vol. 3, pp. 154–156, 2004.
- [33] S. S. Holland and M. N. Vouvakis, "The banyan tree antenna array," *IEEE Trans. Antennas Propag.*, vol. 59, no. 11, pp. 4060–4070, Nov. 2011.
- [34] D. Cavallo, A. Neto, and G. Gerini, "Analysis of common-mode resonances in arrays of connected dipoles and possible solutions," presented at the 6th Eur. Radar Conf., Rome, Italy, Sep./Oct. 20/2, 2009.
 [35] Ansoft HFSS [Online]. Available: www.ansoft.com
- [36] D. M. Pozar and D. H. Schaubert, "Scan blindness in infinite phased arrays of printed dipoles," *IEEE Trans. Antennas Propag.*, vol. 32, no. 6, pp. 602–610, Jun. 1984.
- [37] J. B. Muldavin, T. J. Ellis, and G. M. Rebeiz, "Tapered slot antennas on thick dielectric substrates using micromachining techniques," presented at the IEEE Antennas and Propagation Society Int. Symp., Jun. 1997.
- [38] J. A. Nelson and G. Stavis, "Impedance matching, transformers and baluns," in *Very High-Frequency Techniques*, H. J. Reich and L. S. McDowell, Eds. New York: McGraw-Hill, 1947.
- [39] A. Ludwig, "The definition of cross polarization," *IEEE Trans. An*tennas Propag., vol. 21, pp. 116–119, Jan. 1973.



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