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Improving Scan Gain of Sparse Vivaldi Array with Parasitic Scatterers

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Abstract—This paper presents a densely populated, but sparsely fed antipodal Vivaldi array. Every second element of the array is fed, and the others, parasitic scatterers, are terminated to passive loads. The appropriate terminations of the parasitic elements for improving the scan gain of the array are computed using a principal component analysis-based algorithm. Compared to a reference array without parasitic elements, the main beam realized gain is increased by up to 1.8 dB inside a desired beam-steering sector. The results are verified by measurements of manufactured prototypes.

Index Terms—antenna arrays, antipodal Vivaldi antennas, parasitic scatterers, principal component analysis

I. INTRODUCTION

Transition to higher frequencies has increased the electrical sizes of radio frequency integrated circuits (RFICs) which are responsible for controlling antenna arrays [1]. To avoid complex signal routing and decrease losses, the antenna array’s elements are tightly integrated with the RFIC. Consequently, the antenna array becomes sparse at high frequencies.

In the sparse arrays, there exists additional space between the antenna elements. To fulfill the tightest design requirements for gain, side-lobe level, or impedance matching, for instance, this additional space can be utilized for improving the characteristics of the array and the elements. An effective way to utilize the additional space is to add parasitic elements between the driven elements. Recently, passive parasitic elements have been used for instance in mitigation of voltage standing wave ratio [2] and synthesis of flat-topped or highly directive radiation patterns [3], [4].

Although parasitic elements have been considered as a promising approach to improve antenna arrays, there exists no systematic nor straightforward strategy for obtaining the terminations of the parasitic elements. Typically, the terminations are solved through numerical optimization, either with genetic algorithm [3] or with semi-definite programming [5], [6], [7]. Due to the NP-hard nature of the optimization problem, the globally optimal solution is extremely difficult to find.

In [8], the terminations of the parasitic scatterers suppressing grating lobes are obtained using principal component analysis (PCA). The idea is to approximate the multi-dimensional optimization problem as a single-variable line-search problem. Such approximation seems to give feasible results, and the computational demand is low even with a large number of parasitic elements.

In this work, we design a sparse Vivaldi array that has parasitic Vivaldi elements between the driven ones. Using the parasitic elements, the goal is to increase the realized gain of the antenna array’s main beam when steering the beam inside a given sector. We use the PCA-based approach for solving the appropriate terminations of the parasitic elements. The simulation and prototype measurement results show the feasibility of the PCA-based approach for designing a parasitically loaded Vivaldi array. The parasitically loaded array is compared to a reference array without parasitic elements to evaluate the benefit of using parasitic scatterers.

II. DESIGN METHOD

A. Design goal

Fig. 1 illustrates the idea. Assuming that the antenna elements are smaller than the space available for each element, we can append additional parasitic elements between the driven elements. The driven elements couple to the parasitic ones, and the coupled electromagnetic waves reflect back from the terminations of the parasitic elements. The superposition of the primary waves generated by the driven elements and the waves scattered from the parasitic elements forms the radiation properties of the array. By manipulating the parasitic elements’ loads the radiation properties can be tuned and improved.

We design a five-element antipodal Vivaldi array with 0.8-wavelength inter-element distances. The array is supplemented with six parasitic elements, which are Vivaldi elements as well. Fig. 2a shows the parasitically-loaded antenna array and Fig. 2b shows a zoomed view of a termination of a parasitic element.

The aperture widths of the driven elements are 30 mm. The parasitic elements are 18 mm wide. The total size of the printed circuit board (PCB) is 80 mm x 281 mm. The substrate material is 1.55-mm thick Rogers Ro4350B. 3.36-mm wide
microstrip lines feed the elements. The array is designed for 5 GHz point frequency.

The goal is to find loading for parasitic elements, i.e. the lengths of the shorted microstrip lines, $l_n$, which strengthen the realized gain of the array towards a given beam-steering sector. The beam-steering sector is desired to be $\theta \in [-38^\circ, 38^\circ]$ on $\varphi = 0^\circ$ plane. The sector is chosen based on the fact that grating lobes do not appear in that sector when the main beam is steered inside it.

B. Terminations of parasitic elements

We use the PCA-based algorithm for computing suitable terminations for the parasitic elements. The details of the algorithm are described in [8], and the main workflow is summarized here.

First, the antenna array is simulated in CST Microwave Studio [9] so that each element is excited with a 50-Ω discrete port. In parasitic elements, the ports are placed at the ends of the parasitic elements’ transmission lines, which are located 17.1 mm from the PCB edge. All elements’ ports are between the top and bottom copper. From the CST simulation, we obtain the embedded element patterns (EEP) and the scattering parameters (S-parameters) related to each port.

Then, we determine feeding coefficients for the parasitic elements focusing the radiation toward the beam-steering sector. Let us discretize the beam-steering sector by three directions $(\theta, \varphi) \in \{(-38^\circ, 0^\circ), (0^\circ, 0^\circ), (38^\circ, 0^\circ)\}$. For each beam-steering direction $(\theta_i, \varphi_i)$ separately, we compute the feeding coefficients of driven and parasitic elements, $a^D_n \in \mathbb{C}^5$ and $a^P_n \in \mathbb{C}^6$, which focus the radiation beam toward the referred direction. The feeding coefficients are illustrated in Fig. 1. They are chosen so that they maximize the realized gain and have unit magnitudes, as described in [10].

Next, we compute the reflection coefficients of the parasitic element terminations which would implement the optimal feeding coefficients. In parasitic element $n$, the reflection coefficient

$$r_{nl} = \frac{a^P_n}{\sum_{j=1}^5 s^P_{nj} \hat{a}^D_j + \sum_{j=1}^6 s^P_{nj} \hat{a}^P_j}$$

realizes the feeding coefficient $\hat{a}^P_n$. The S-parameter $s^P_{nj}$ describes the coupling between $j$-th driven and $n$-th parasitic element, and $s^P_{nj}$ describes the coupling between $j$-th parasitic and $n$-th parasitic element. The coupling between the driven and parasitic elements is illustrated in Fig. 1.

The reflection coefficients (1) are not realizable in practice for two reasons. First, their magnitudes may be larger than unity, requiring then active loads for the parasitic elements. Secondly, they depend on the beam-steering direction, requiring tunable loads. We want to have passive and fixed beam-independent loads.

We determine suitable beam-independent reflection coefficients using principal component analysis. Let $R \in \mathbb{C}^{6 \times 3}$ be the matrix of beam-dependent reflection coefficients (1) of the parasitic elements. If $p_1 \in \mathbb{C}^3$ is the first principal component vector of $R$, the variance in the beam-dependent reflection coefficients is maximized along $p_1$. Thus, we may assume that a point on a complex plane $\mathcal{C} = \{ r = e^{j\phi} p_1 : c \in \mathbb{C} \}$ is feasibly close to all the points of $R$. The problem is then to find the feasible reflection coefficient vector $r$ from $\mathcal{C}$.

To terminate the parasitic elements reactively, we force the magnitudes of the reflection coefficients to unity. Let us introduce the unit-magnitude principal component

$$\hat{p} = \frac{p_1}{|p_1|},$$

where the division and absolute value are taken element-wise. When a reflection coefficient vector is on the complex plane $\hat{\mathcal{C}} = \{ r = \hat{e}^{j\phi} \hat{p} : \phi \in [0, 2\pi] \}$, it can be implemented with reactive terminations.

Finally, the vector of unit-magnitude beam-independent reflection coefficients is

$$\hat{r} = \hat{e}^{j\hat{\phi}} \hat{p},$$

where

$$\hat{\phi} = \arg \max_{\phi \in [0, 2\pi]} \left| \sum_{i=1}^L g_i(\hat{e}^{j\phi} \hat{p}) \right|.$$

The function $g_i(r)$, defined in [8], gives the realized gain of the array toward the direction $(\theta_i, \varphi_i)$ when the parasitic elements are terminated with reflection coefficients $r$.

The terminations are then implemented using shorted microstrip lines with appropriate lengths. The shorted transmission line in element $n$ realizes the reflection coefficient $\hat{r}_n$, when the transmission line length is

$$l_n = \frac{1}{\beta} \tan^{-1} \left( \frac{z_{0n} \, 1 + \hat{r}_n}{j z_{0n} \, 1 - \hat{r}_n} \right),$$

where $\beta$ and $z_{0n}$ are phase constant and line impedance of the microstrip line at 5 GHz, respectively. $z_{0n} = 50 \Omega$ is the impedance of the discrete port in the simulation. Negative $l_n$
TABLE I
LENGTHS OF THE TRANSMISSION LINE TERMINATIONS OF THE PARASITIC ELEMENTS.

<table>
<thead>
<tr>
<th>n</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>l_n (mm)</td>
<td>-8.3</td>
<td>-7.4</td>
<td>-5.1</td>
<td>-4.3</td>
<td>-5.7</td>
<td>-8.0</td>
</tr>
</tbody>
</table>

**Fig. 3. Manufactured prototypes.** a) Sparse reference array from the top. b) Parasitically loaded array from top. c) Sparse, bottom. d) Parasitics, bottom.

indicates that the line should be shortened from its initial length, and any imaginary part of \( l_n \) can be neglected. A shortened transmission line termination of parasitic element \( n \) is illustrated in Fig. 2b.

The obtained terminations for the parasitic elements are described in Table I. The indexing of the elements is ascending towards positive \( x \)-axis, and starting from 1 for both driven and parasitic elements.

The reason for using principal component analysis instead of directly optimizing the realized gain as a function of the phases of the reflection coefficients is that the optimization problem can be reduced to the one-dimensional line-search problem (4). Examples of how to use direct optimization for determining the terminations are presented for instance in [2], [3], [4], [5], [6], [7]. The issue in direct optimization is the NP-hardness of constant modulus-constrained optimization. Using proper relaxation, the problem can be solved but with a large number of parasitic elements the optimization is computationally very demanding. The PCA approach may not always give optimal terminations, but often feasible, and is always computationally lightweight regardless of the number of parasitic elements.

### III. RESULTS

The designed parasitically-loaded array is manufactured and characterized by both simulations and measurements. In addition, we design and manufacture a sparse reference array that has a similar topology as the proposed design but no parasitic elements. Comparing the simulation and measurement results of these two arrays, we see the benefit of loading a sparse array parasitically. The two manufactured arrays with coaxial connectors are shown in Fig. 3. The results are analyzed at 5-GHz point frequency.

**Fig. 4.** Simulated and measured embedded element patterns of the center and an edge element in the decibel scale.

**Fig. 5.** Embedded element patterns of the measured arrays.

**Fig. 6.** shows the active reflection coefficients (ARC) of the simulated arrays. When scanning towards the broadside direction, the total active reflection coefficient (TARC) is decreased by 2.5 dB with parasitic elements. The parasitic elements affect more the matching of the edge elements than the center element. Overall, the curves show that the active impedance matching can be significantly improved with appropriately terminated parasitic elements.

Next, we study the beam-steering capabilities of the two simulations. **Fig. 5** presents the measured EEPs of all elements of both arrays. The figure illustrates also the desired beam-steering sector. Compared to the sparse array, the array with parasitic elements radiates less backside, and focuses the radiation into the scanning sector, as desired.
Fig. 6. Total active reflection coefficient and active reflection coefficients of the edge (1) and the center element (3) as a function of the beam-steering direction.

Fig. 7. Array radiation patterns with three different scan angles. The dashed lines refer to the sparse array and solid ones to the parasitically-loaded array. The black lines show the scan gain envelope.

arrays. Fig. 7 shows the radiation patterns of the arrays with three different scan directions. In addition, the scan gain envelope is shown. We see that when the beam is steered closer to the edge of the area, the grating lobe becomes significantly stronger.

The realized gain is increased up to 1.8 dB in the scan area. Considering the broadside scanning case, the back-lobe level compared to the main beam is suppressed by 4.3 dB when the array involves parasitic scatterers. Overall, the radiation patterns are more focused onto the scan window and the beam efficiency is thus improved with the parasitic elements compared to the array without them.

IV. CONCLUSION

We designed a five-element antipodal Vivaldi array with six passively terminated parasitic elements between the driven elements. The reactive terminations were computed based on the computationally lightweight PCA approach to tailor the embedded element patterns of the driven elements. The resulting parasitically-loaded array introduced stronger radiation toward the desired radiation sector, and lower back and side lobes compared to the reference array without the parasitic elements.

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