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INTEGRATED DECOUPLING AND MATCHING NETWORK INCORPORATED IN THE GROUND PLANE OF A COMPACT MONOPOLE ARRAY

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ABSTRACT: Compact arrays enable various applications such as antenna beam-forming and multi-input, multi-output (MIMO) schemes on limited-size platforms. The reduced element spacing in compact arrays introduces high levels of mutual coupling which can affect the performance of the adaptive array. This coupling causes a mismatch at the input ports, which disturbs the performance of the individual elements in the array and affects the implementation of beam steering. In this article, a reactive decoupling network for a 3-element monopole array is used to establish port isolation while simultaneously matching input impedance at each port to the system impedance. The integrated decoupling and matching network is incorporated in the ground plane of the monopole array, providing further development scope for beamforming using phase shifters and power splitters on a separate substrate in a double-layered configuration.

Key words: antenna arrays; mutual coupling; decoupling network; matching network; phased arrays

1. INTRODUCTION

Spatial diversity can significantly enhance the system capacity of wireless networks [1]. The adverse effects of mutual coupling in arrays are usually restricted using an interelement spacing of at least half a wavelength (λ/2). For antenna diversity in mobile applications, an element spacing considerably smaller than λ/2 becomes unavoidable. Reduced element spacing results in increased mutual coupling between array elements, which can decrease the antenna gain and cause significant system performance degradation [2, 3].

The effects of mutual coupling can be countered using passive, lossless decoupling, and matching networks. A decoupling network consists of interconnected reactive elements and/or transmission line sections and stubs. It provides an additional signal path between the array elements, which effectively cancels the external coupling between them. Various implementations of decoupling networks have been described in the literature [4–9]. A systematic design approach for larger circulant symmetric arrays has also been proposed [10]. The design procedure involves the repeated decoupling of the characteristic eigenmodes of the array. In most cases, additional matching networks are required to ensure an impedance match at each port.

In this article, a reactive decoupling network for a 3-element monopole array is used to establish port isolation while simultaneously matching the input impedance at each port to the system impedance. The integrated decoupling and matching network (DMN) is incorporated in the ground plane of the monopole array, providing further development scope for compact beamforming circuitry using phase shifters and power splitters on a separate substrate in a double-layered configuration.

2. INTEGRATED DMN SYNTHESIS

The 3-element uniform circular array in Figure 1 is characterized by the scattering matrix $S^a$, given by

$$
S^a = \begin{bmatrix}
S_{11} & S_{12} & S_{13} \\
S_{21} & S_{22} & S_{23} \\
S_{31} & S_{32} & S_{33}
\end{bmatrix}, \quad (1)
$$

The corresponding admittance matrix can be calculated from

$$
Y^a = Y_0 (I - S^a) (I + S^a)^{-1} = \begin{bmatrix}
Y_{11}^a & Y_{12}^a & Y_{13}^a \\
Y_{21}^a & Y_{22}^a & Y_{23}^a \\
Y_{31}^a & Y_{32}^a & Y_{33}^a
\end{bmatrix}, \quad (2)
$$

where, $I$ is a 3×3 identity matrix and $Y_0$ is the characteristic admittance of the system. Impedance transformation can be obtained using the circuit shown in Figure 1. It consists of three identical transmission line with a characteristic impedance $Z_{01}$ and electrical length $\theta_1$. Using transmission line theory the admittance matrix at Ports 1’, 2’, and 3’ can be calculated from [11]

$$
\begin{bmatrix}
Y_{11}' \\
Y_{12}' \\
Y_{13}'
\end{bmatrix} = B A^{-1} = \begin{bmatrix}
Y_{11} & Y_{12} \\
Y_{21} & Y_{22} \\
Y_{31} & Y_{32}
\end{bmatrix}, \quad (3)
$$

with

![Figure 1](https://example.com/figure1.png)
The proposed decoupling network incorporates impedance matching at the ports without the need of additional matching circuits. The design consists entirely of transmission line sections with a characteristic impedance $Z_{\text{TL}}$ and electrical length $\theta_{\text{TL}}$. The admittance matrix of single transmission line section is given by

$$
Y_{\text{TL}} = \begin{bmatrix}
-j\cot(\theta_{\text{TL}})/Z_{\text{TL}} & j\cosec(\theta_{\text{TL}})/Z_{\text{TL}} \\
-j\cosec(\theta_{\text{TL}})/Z_{\text{TL}} & -j\cot(\theta_{\text{TL}})/Z_{\text{TL}}
\end{bmatrix} = \begin{bmatrix} x & y \\ y & x \end{bmatrix}
$$

(6)

The DMN provides feed ports in-between array elements, as seen in Figure 2, with $Z_{\text{TL}} = Z_{01}$ and $\theta_{\text{TL}} = \theta_2/2$. The admittance matrix $Y'$ at Ports 1', 2', and 3' can be calculated from (11)

$$
Y' = Y_1 + Y_3(\mathbf{Y} + Y_1)^{-1}Y_2.
$$

(7)

where

$$
Y_1 = \begin{bmatrix} 2x & y & y \\ y & 2x & y \\ y & y & 2x \end{bmatrix}
$$

(8)

$$
Y_2 = \begin{bmatrix} -y & 0 & -y \\ -y & -y & 0 \\ 0 & -y & -y \end{bmatrix}
$$

(9)

$$
Y_3 = \begin{bmatrix} 0 & y & y \\ 0 & y & 0 \\ y & 0 & y \end{bmatrix}
$$

(10)

The scattering parameters are then obtained using the following transformation

$$
S' = (\mathbf{I} - Z_0Y') \cdot (\mathbf{I} + Z_0Y')^{-1}
$$

(11)

The fraction of radiated power for a symmetrical 3-element array is given by

$$
P_{\text{rad}}/P_{\text{in}} = 1 - |S'_{11}|^2 - 2|S'_{12}|^2.
$$

(12)

To aid the implementation, the electrical lengths $\theta_1$ and $\theta_2$ can be increased by multiples of $180^\circ$ and $360^\circ$, respectively.

3. DESIGN EXAMPLE

The design procedure is illustrated using a 3-element array with an interelement spacing of 0.15\,\lambda at a frequency of 2.45, GHz. Standing monopoles were used as the array elements. These elements were selected for their wide angle steering capability and simplicity in manufacturing. The monopoles shown in Figure 3 have a top radius of 1.5 mm and a bottom radius of 0.79 mm. The top cylindrical section is 23 mm long and the bottom section has a length of 1 mm. The tapered length between the two sections is 5 mm. The monopoles were mounted on Rogers RT 4003 substrate (thickness 0.813 mm, relative permittivity $\varepsilon_r=3.55$ and loss tangent $\tan\delta=0.0027$) and fed using a 50 \,Ω conductor-backed coplanar waveguide (CBCPW) with a gap of 1 mm gap and a track width of 1.7 mm. A photograph of the original array is shown in Figure 4, while the measured scattering parameters for the original array are depicted in Figure 5.

The parameters of the DMN were calculated using the optimization cost function defined in (11), and the results are shown

**Figure 2** Integrated DMN for the 3-element circular array. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
in Table 1. Given the array geometry, the solution for electrical length $h_2$ was too short and was, therefore, extended by $360^\circ/C_{14}$.

After additional tuning, the physical line lengths and track widths for the CBCPW line sections were obtained and are listed in Table 2. The transmission line section which provide the impedance transformation needs to be meandered to keep the DMN dimensions as compact as possible. The coordinates of a horizontal meandered line can be obtained using the following parametric equation

$$
(x(t), y(t)) = \left( R + h \frac{\Delta R}{2} \left[ 1 - \cos \left( 2n\pi t \right) \right] \right), \quad 0 \leq t \leq 1.
$$

(12)

Here, $R$ is the initial horizontal displacement of the feed line from the origin, $l$ is the required total horizontal length of the line, $n$ is the number of bends in the meandering curve, and $\Delta R$ represents the amplitude of the meandering deviation from the straight line connecting the starting point and the endpoint, as illustrated in Figure 6. Similarly, a meandering curved line was implemented to connect two impedance transforming line sections, with the coordinates of points on the curve defined by

$$
(x(\theta), y(\theta)) = \left( R + \frac{\Delta R}{2} \left[ 1 - \cos \left( \frac{2n\pi (\theta - \theta_s)}{\theta_e - \theta_s} \right) \right] \right) + \frac{\Delta R}{2} \left[ 1 - \cos \left( \frac{2n\pi (\theta - \theta_s)}{\theta_e - \theta_s} \right) \right] \sin \theta
$$

(13)

Here, $R$ is the radius to the starting point of the curved line from the origin, $\Delta R$ defines the amplitude of the meandering deviation from a circular curve between the connecting points, $n$ is the number of bends in the meandering curve, while $\theta_s$ and $\theta_e$ define angles of the starting point and end point of the curve, as shown in Figure 7. Using these equations, a line

**Figure 3** Dimensions of the standing monopole elements.

**Figure 4** Photograph of the original 3-element monopole array fed by 50 $\Omega$ CBCPW lines. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

**Figure 5** Measured scattering parameters of the original 3-element monopole array. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

**Figure 6** Meandering line for restricting the overall dimensions of a straight transmission line section

**Table 1** DMN Design Parameters

| Parameter | Value |
|-----------|-------|
| $Z_{01}$  | 49.3 $\Omega$ |
| $\theta_1$ | 106.8$^\circ$ |
| $Z_{02}$  | 61.6 $\Omega$ |
| $\theta_2$ | 75.4$^\circ$ + 360$^\circ$ = 435.4$^\circ$ |

**Table 2** Physical Dimensions of the CBCPW Line Sections

| Parameter | Value |
|-----------|-------|
| $l_1$     | 27.05 (mm) |
| $w_1$     | 2.01   |
| $l_2$     | 89.2   |
| $w_2$     | 1.6    |
of arbitrary length can be defined between the start and end points.

The final implementation of the DMN is shown in Figure 8. The circuit was analyzed using CST Microwave Studio, and the simulated results for the scattering parameters are compared with measured data in Figures 9 and 10. Figure 11 shows measured and simulated azimuth radiation patterns for the case where a single port is driven and the unused ports terminated in matched loads.

4. CONCLUSION

We successfully demonstrated the implementation of an integrated DMN for a 3-element monopole array. The DMN is incorporated in the ground plane of the monopole array. Excellent agreement between simulated and measured results was obtained. The incorporation of the DMN in the ground plane of the array provides further development scope for additional beamforming circuitry in double-layered circuits while restricting the overall size of the system.

Figure 7 Meandering line for restricting the overall dimensions of a circular segment of transmission line. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 8 Integrated DMN incorporated into the groundplane of the monopole array. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 9 Measured and simulated reflection coefficient of the array with the DMN. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 10 Measured and simulated results for the mutual coupling of the array with the DMN. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 11 Measured and simulated azimuth radiation for the decoupled and matched array fed at a single port and the unused ports terminated in matched loads. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
ABSTRACT: In this article, a new wideband bandpass filter (BPF) adopting two different types of resonators is proposed. The filter is constructed of four same open loop resonators and one inverted I-shaped resonator located at the middle. As a result, compared with conventional BPF without inverted I-shaped resonator, the proposed BPF exhibits one new transmission zero around the passband, which can adjust bandwidth and increase the selectivity at high-side edge of the passband significantly. The desired bandwidth can be obtained from the condition $Z_{in} = \infty$, and the fundamental resonant frequency can be extracted: \[ \text{frequency} = \frac{1}{2\pi \sqrt{L_C C}} \]

1. INTRODUCTION

Recently, there has been increased interest in wideband bandpass filter (BPFs) for application in modern wireless communication systems, great efforts have been performed to realize various kinds of wideband BPFs [1–8]. In [1], an ultrawideband (UWB) BPF with one of the outer coupled lines shorted to ground is proposed to exhibit two new transmission zeros and increase the selectivity significantly. However, the interval between the transmission zeros and high-side edge or low-side edge of the passband is still large. To tune resonant frequencies and transmission zeros, a quadruple-mode resonator BPF with tunable bandwidth is proposed using two types of microwave varactors in [2], where the high-side edge and low-side edge of the passband can be separately varied. An UWB BPF designed based on a three-line coupled resonator is presented, four bent stubs are loaded to improve the performance in [3], where the selectivity needs to be further improved. In [4], a tunable wideband bandwidth with two transmission zeros near the passband is realized by changing the characteristic impedance of the symmetric multimode resonator or the coupling gap. To obtain the required cutoff frequency and out-of-band performance, a broadband BPF based on parallel-coupled microstrip line is proposed, and are achieved by placing L-shaped capacitive cross-coupling open stubs at the middle resonator [5]. However, the selectivity at low-edge of the passband has a large space to be increased. In [6], an UWB BPF with a ring resonator and two stepped-impedance stubs are loaded to provide good wideband filtering performance and sharp rejection skirts. To improve the selectivity, one folded half-wavelength resonator in the middle and two quarter-wavelength resonators are used to create five transmission zeros due to the cross coupling in Ref. [7], where the space between high-edge of the passband and the third transmission zero is still large. In [8], a tri-mode wideband BPF with controllable bandwidths and passband selectivity is proposed where shunted short- and open-stub loaded resonators are adopted.

In this article, based on the conventional structure of miniature open-loop half-wavelength resonator filter [9], a new compact wideband BPF with high selectivity and controllable bandwidth has been designed and reported. Compared with the conventional BPF as Figure 1(a), the proposed BPF adopts one inverted I-shaped resonator at the middle as Figure 1(b). By adjusting the width of upper or bottom coupling gaps ($g_3$ and $g_4$), which are located between inverted I-shaped resonator and two adjoining open loop resonators, the desired bandwidth in a large range and one new shifting transmission zero can be easily achieved, which effectively improves the selectivity. Similarly, unlike the majority of BPFs, the passband bandwidth is almost proportional to the width of upper or bottom coupling gaps. At last, Good results are obtained as demonstrated in simulation and experiment.

2. THE INVERTED I-SHAPED RESONATOR

Figure 2(a) shows the inverted I-shaped resonator, which is symmetric at horizontal direction. The resonator is composed of three microstrip lines with same line width, where the characteristic impedance is $Z_c$, and the electrical lengths are $\theta_1$, $\theta_2$, and $\theta_3$, respectively. Because of symmetric characteristic, even- and odd-mode excitation methods can be applied to analyses the inverted I-shaped resonator.

Even- and odd-mode equivalent circuits are shown in Figures 2(b) and 2(c), the even-mode has one pattern as Figure 2(b), from the condition $Z_{in} = \infty$, and the fundamental resonant frequency can be extracted: