The Use of Symmetry to Simplify the Mixed-Potential Integral-Equation Method with Application to N-Way Radial Power Dividers/Combiners with Isolation Resistors

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Abstract—This paper presents a full-wave analysis for the scattering characteristics of N-way radial power combiners, especially the matching and isolation among the output ports. The analysis hybridizes the mixed-potential integral-equation (MPIE) formulation with the idea of discrete Fourier transform (DFT) to extract the scattering parameters accurately and efficiently. The technique is also generalized to analyze the power combiners in which the lumped-resistor elements are included to improve port isolation. The resultant additional equations necessary to accommodate the lumped elements into the hybrid DFT–MPIE technique are derived. Examples demonstrating the numerical efficiency and accuracy of the technique are given. The computed results are found to be in good agreement with the measured data.

Index Terms—DFT, MPIE, power combiner, power divider.

I. INTRODUCTION

Electromagnetic problems associated with planar structures of N-fold rotational symmetry structures are encountered in some microwave applications such as the radial power dividers/combiners [1]–[3]. Several approaches have been presented to analyze such problems, e.g., [2], [4], to name a few. In principle, they are based on the radial transmission-line theory [5] or the more general planar circuit approach [6]. To approximate fringing effects near the open boundary, the original structure is usually enclosed by lateral magnetic walls of a larger geometrical dimension with the substrate material of some effective dielectric constant [4], [7]. Nonetheless, the electromagnetic field in the structure is assumed to be composed of radial modes and of vertical polarization only. Such assumptions severely limit the accuracy of these approaches in extracting some characteristics of the power combiners, especially the isolation between the output ports. As a result, it is difficult to improve the output isolation by choosing proper positions and values of the isolation resistors.

In order to precisely extract the scattering parameters of the radial power divider and combiner, the full-wave analysis, which can take into account the effects due to fringing field, radiation, and surface wave leakage, should be resorted to. In this respect, the mixed-potential integral-equation (MPIE) formulation distinguishes itself in its flexibility for analyzing arbitrarily shaped planar circuits [8], [9]. However, the relatively heavy requirement in memory storage and central processing unit (CPU) time may easily exhaust the available computer resources when it is applied for more complicated structures. Consequently, the general multiport network analysis of practical multiway radial power dividers/combiners based on the MPIE method has not been yet reported in the literature.

In this paper, we employ a hybrid approach exploiting the structural symmetry to alleviate the computational load required in the MPIE solution. Section II applies the idea of discrete Fourier transform (DFT) to reduce the original problem of N-fold symmetry into N eigenproblems, over each of which the MPIE involves fewer unknowns. Section III derives the additional equations required for handling lumped elements in this hybrid DFT–MPIE technique. Numerical examples of several radial power dividers/combiners are presented in Section IV, including validation of the technique through comparison between computed results and measured data. Finally, brief conclusions are drawn in Section V.

II. HYBRID DFT–MPIE TECHNIQUE

A. DFT

The N-fold rotational symmetry in the structure can be exploited to simplify the integral-equation solution by the idea of DFT [10], [11]. Consider a typical N-way radial power divider/combiner, shown in Fig. 1. In characterizing the multiport network scattering parameters, it is necessary to consider the excitation from one of the output ports. Due to the symmetry among the output ports, the single-element excitation can be expressed in terms of a finite discrete spectrum of angular harmonics. Let $X_n$ denote a certain variable $X$ over the $n$th element. Its Fourier series representation has the form [12]

$$X_n = \sum_{k=0}^{N-1} \tilde{X}_k e^{j(2\pi(n-1)/N)k}, \quad n = 1, 2, \cdots, N \quad (1)$$
where the Fourier series coefficients $\tilde{X}_k$ are

$$\tilde{X}_k = \frac{1}{N} \sum_{n=1}^{N} X_n e^{-j(2\pi(n-1)/N)k}, \quad k = 0, 1, \ldots, N-1.$$  \hfill (2)

Due to the application of the DFT, the original problem with an arbitrary excitation at the output ports can be divided into $N$ eigenproblems, shown in Fig. 2, for which the excitation vectors over the $N$ ports are

- prob. 0 : $\{1, 1, \ldots, 1\}$
- prob. 1 : $\{1, e^{j(2\pi/N)}, \ldots, e^{j((N-1)2\pi/N)}\}$
- prob. 2 : $\{1, e^{j(3\pi/N)}, \ldots, e^{j((N-1)4\pi/N)}\}$
- prob. $N-1$ : $\{1, e^{j(2\pi/N)(N-1)}, e^{j(3\pi/N)(N-1)}, \ldots, e^{j((N-1)2\pi(N-1)/N)}\}$. \hfill (3)

For example, the problem of a unit excitation of the first element, i.e., port 2, can be expanded into the superposition of the $N$ eigenproblems with expansion coefficients $\tilde{a}_0 = \tilde{a}_2 = \cdots = \tilde{a}_{N-1} = 1/N$.

In each eigenproblem, the current density on the $N$ elements is identical, except for a phase shift introduced by the complex exponentials expressed in (3). Given the current distribution $\tilde{J}_k$ in each eigenproblem ($k = 0, 1, \ldots, N-1$), the current distribution in any element $n$ of the original problem can be expressed by

$$\tilde{J}_n = \sum_{k=0}^{N-1} \tilde{J}_k e^{j((n-1)2\pi/N)k}, \quad n = 1, 2, \ldots, N.$$ \hfill (4)

**B. MPIE Formulation and Method-of-Moments (MoM) Solution**

The MPIE formulation relates the known tangential electric field incident on the conductor in the presence of the dielectric layer to the unknown surface current on all the conductors. The tangential electric field is expressed in terms of the magnetic vector and electric scalar potentials $\vec{A}$ and $\Phi$, which are obtained from convolution of the current density $\tilde{J}$ and charge density $\rho$, respectively, with the Green’s function $G_A$ and $G_V$. Noting that the charge density $\rho$ can be written from $\tilde{J}$ through the continuity equation, $\tilde{J}$ is the only unknown of interest.

It is a well-known procedure to solve MPIE for the unknown $\tilde{J}$ by applying the MoM [13]. The conductor is subdivided into small cells $\Omega_p$ with subdomain basis functions $\tilde{f}_n$ approximating the current flow over the cells. To increase the modeling flexibility, we employ the roof-top basis functions [8] for the regular region and, if necessary, the triangular-domain basis function [9] for the region near the curved boundary. Each subdomain basis function $\tilde{f}_n$ models the current flow over two adjacent cells sharing a common branch. Let $\tilde{B}_p(\vec{r})$ correspond to the current distribution on the cell $\Omega_p$, which flows out of the $i$th side of the cell. The subdomain basis function $\tilde{f}_n$ can be written in the form of

$$\tilde{f}_n = \sum_{p} T_{np} \tilde{B}_p(\vec{r})$$ \hfill (5)

where

$$T_{np} = \begin{cases} 
1, & \text{if the } n\text{th branch current flows out of the } i\text{th side of the } p\text{th cell} \\
-1, & \text{if the } n\text{th branch current flows into the } i\text{th side of the } p\text{th cell} \\
0, & \text{otherwise} \end{cases}$$
The current distribution \( \mathbf{J} \) is then expanded into these functions by
\[
\mathbf{J} = \sum_n \mathbf{f}_n I_n = \sum_n \sum_p \mathbf{B}_{p,n} T_{p,n} I_n
\]
in which \( I_n \) is the unknown current across the \( n \)th branch.

The expansion functions \( \mathbf{B}_{p,n} \) are also used as the weighting functions, producing a matrix equation between voltage vector \( \mathbf{V} \), impedance matrix \( \mathbf{Z} \) and the current coefficients \( I_n \), i.e.,
\[
\mathbf{V}_m = \sum_n Z_{mn} I_n = \sum_n \left\{ \sum_p \sum_q T_{mq,ij} Z_{q,ij} T_{p,ij} \right\} I_n
\]
The matrix elements \( Z_{q,j,p} \) can be further decomposed into three parts [14, p. 186] and written in the following form:
\[
Z_{q,j,p} = j\omega L_{q,j,p} + \frac{1}{j\omega C_{q,p}} + R_{q,j,p}
\]
where the terms \( L_{q,j,p}, C_{q,p}, \) and \( R_{q,j,p} \) are given by
\[
L_{q,j,p} = \int_{\Omega_q} \int_{\Omega_p} \mathbf{B}_{q}(\mathbf{r}) \cdot \mathbf{\bar{G}}_{A}(\mathbf{r}) \cdot \mathbf{\bar{B}}_{p}(\mathbf{r}) d\mathbf{r} d\Omega
\]
\[
C_{q,p} = \left\{ \int_{\Omega_q} \int_{\Omega_p} \Pi_{p}(\mathbf{r})G_{V}(\mathbf{r})\Pi_{p}(\mathbf{r}) d\Omega d\mathbf{r} \right\}^{-1}
\]
\[
R_{q,j,p} = \delta_{pq} Z_s = \int_{\Omega_q} \mathbf{\bar{B}}_{q}(\mathbf{r}) \cdot \mathbf{\bar{B}}_{p}(\mathbf{r}) d\mathbf{r}
\]
in which the scalar function \( \Pi_{p} = \nabla \cdot \mathbf{B}_{p} = 1/(\text{area of } \Omega_p) \), \( \delta_{pq} \) takes the value of one for \( p = q \) and zero for \( p \neq q \), \( \mu \) represents the permeability in free space, and \( \sigma \) is the conductivity of the conductor. The surface impedance \( Z_s \) in (8) can approximate the effects of the finite conductivity \( \sigma \) [14, p. 137]. The approximation is appropriate only when the conductor thickness is much smaller than the substrate height and much larger than the skin depth.

C. Computation of Matrix Elements in Eigenproblems

For each eigenproblem, say the \( k \)th one, the contributions from the current distribution on all other elements can be added into the first element by using the relationship of the rotational symmetry. Hence, one need only solve the current distribution on the first element, while using (4) to reconstruct those on all the other elements. Consequently, the dimension of the impedance matrix in (7) will be substantially reduced.

For a microstrip structure including isotropic dielectric layers only, the dyadic Green’s function \( \mathbf{\bar{G}}_{A} \) in the tangential plane is a scalar function \( G_{A} \). Both the potentials \( G_{V} \) and \( G_{A} \) of an \( x \)-directed horizontal electric dipole (HED) are given in the space domain by [14]. The integral in (9) should be rewritten to include all contributions coming from other elements, which are in the following form:
\[
L_{q,j,p} = \sum_{n=0}^{N-1} \int_{\Omega_q} \int_{\Omega_p} G_{A}(x, y | x', y') I_n
\]
\[
\cdot \left\{ B_{q,j}^{x}(x, y) \left[ B_{p}^{x}(x', y') \cos(n\alpha) - B_{p,n}^{x}(x', y') \sin(n\alpha) \right]
\right. \\
\left. + B_{q,j}^{y}(x, y) \left[ B_{p}^{y}(x', y') \sin(n\alpha) + B_{p,n}^{y}(x', y') \cos(n\alpha) \right] \right\} e^{j\alpha k n} d\mathbf{r} d\Omega
\]
\[
C_{q,p} = \left\{ \int_{\Omega_q} \int_{\Omega_p} \Pi_{p}(x, y) \left[ \sum_{n=0}^{N-1} G_{V}(x, y | x', y') I_n \right] \right\}^{-1}
\]
\[
R_{q,j,p} = \delta_{pq} Z_s = \int_{\Omega_p} \mathbf{B}_{q}(x, y) \cdot \mathbf{\bar{B}}_{p}(x', y') d\Omega
\]
where \( \alpha = 2\pi/N \). Note that \( R_{q,j,p} \) stays unchanged since its value is zero when cell \( p \) is different from cell \( q \).

D. Reconstruction of Scattering Matrix

In general, \( N \) independent excitation schemes are required to evaluate an \( N \)-port network. For example, the two-port network for prob. 0 in Fig. 2 can be characterized by solving the current on the both ports, while imposing excitation on ports 1 and 2 sequentially. For the excitation, we apply a unit voltage to the driven port while all other ports grounded. The current distribution is then found by solving (7), which yields the self admittance seen from the driven port and the mutual admittance between the driven port and all other ports.

Repeating the excitation on all the ports, the admittance matrix of the \( N \)-port network can be obtained and, in turn, converted to the desired scattering matrix representation. Here, all the ports are assumed to be terminated to 50 \( \Omega \) when transforming the admittance matrix into the scattering matrix.

From the scattering parameters of the \( N \) eigenproblems, the desired multiport network scattering parameters \( S_{(j+1),j} \) \((i, j = 1, 2, \ldots, N+1)\) of the \( N + 1 \) input/output ports in the original problem, shown in Fig. 1, can be reconstructed. To be more specific, consider the scattering parameters in each eigenproblem. For prob. 0, shown in Fig. 2, all the output ports are equipotential and the scattering matrix can be written as
\[
[S_0] = \begin{bmatrix} S_{1,1} & S_{1,2} \\ S_{2,1} & S_{2,2} \end{bmatrix}
\]
where
\[
S_0 = \sum_{n=1}^{N} S_{2,(n+1)}
\]
denotes the reflection coefficient at port 2 when all the output ports are subject to the same excitation of unit amplitude. Note

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that $S_{11}$ and $S_{12} = S_{21}$ are the same as those of the original problem.

For prob. $k (k = 1, 2, \cdots, N-1)$, the incident port (port 1) can be assumed to be grounded since the voltage should be zero there. The reflection coefficient at the output port (port 2) can be written as

$$
\frac{v_2}{v_1} = \frac{Z_{21}}{Z_{11}} \frac{Z_{22} + R}{Z_{11} + R} \cdots \frac{Z_{2n}}{Z_{11} + R}.
$$

Due to the application of the DFT, the scattering parameters in the original problem can be obtained from the reflection coefficients of the eigenproblems. The parameters when the device is operated as a power divider, i.e., $S_{11}$, $S_{1,j} = S_{j,1} = S_{1,2} (j = 2, 3, \cdots, N+1)$, have already been solved in prob. 0. All the other parameters that are necessary for power-combiner operation can be found from (12) and (13) by

$$
S_{k,j} = \frac{1}{N} \sum_{i=0}^{N-1} S_{k,i} e^{-j(2\pi i/j)/N}, \quad k = 1, 2, \cdots, N-1.
$$

III. MANAGEMENT OF LUMPED-CIRCUIT ELEMENTS

Typically, the output isolation of a divider can be improved by introducing proper isolation resistors among the output ports. Recently, it has been demonstrated how the isolation resistor in a Wilkinson power combiner can be included in the conventional MPIE analysis [15]. Here, the idea will be generalized to handle such lumped elements in the hybrid DFT–MPIE technique. As described in Section II-A, the original problem can be divided into $N$ eigenproblems. Consider the $k$th eigenproblem shown in Fig. 3(a). Let $R$ denote the value of isolation resistor and $\phi = (2\pi k)/N$, the phase difference of the adjacent element.

To help the derivation of the equations due to the insertion of the lumped elements, the isolation resistor $R$ is separated into two identical resistors $R/2$ in series, as shown in Fig. 3(b). Based on the Kirchhoff current law and the phase relation, the currents flowing along the resistors satisfy

$$
i_1 = i_2 e^{j\phi}.
$$

The virtual voltages at the center points of the resistors satisfy

$$v_1 = v_2 e^{j\phi}.
$$

Now, the MoM solution can be applied to the eigenproblem shown in Fig. 3(c). The MPIE will be reduced into a set of simultaneous equations. The insertion of the cascaded resistors results in the term $R/2$ added to the corresponding diagonal elements of the impedance matrix [14]. Consequently, the modified matrix equation becomes

$$
\begin{bmatrix}
v_1 \\
v_2 \\
0 \\
\vdots \\
0
\end{bmatrix} = 
\begin{bmatrix}
Z_{11} + \frac{R}{2} & Z_{12} & \cdots & Z_{1n} \\
Z_{21} & Z_{22} + \frac{R}{2} & \cdots & Z_{2n} \\
\vdots & \vdots & \ddots & \vdots \\
Z_{n1} & Z_{n2} & \cdots & Z_{nn}
\end{bmatrix}
\begin{bmatrix}
v_1 \\
v_2 \\
0 \\
\vdots \\
0
\end{bmatrix}.
$$

The voltages $v_1$, $v_2$ and currents $i_1$, $i_2$ are related to each other by (15) and (16). Treating them as unknowns and rearranging the matrix equation in (17), the final matrix to be solved is in the following form:

$$
\begin{bmatrix}
0 \\
0 \\
0 \\
\vdots \\
0
\end{bmatrix} = 
\begin{bmatrix}
1 - e^{j\phi} & 0 & 0 & \cdots & 0 \\
0 & 0 & 1 & \cdots & 0 \\
-1 & 0 & Z_{11} + \frac{R}{2} & Z_{12} & \cdots & Z_{1n} \\
0 & -1 & Z_{21} & Z_{22} + \frac{R}{2} & \cdots & Z_{2n} \\
\vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & 0 & Z_{n1} & Z_{n2} & \cdots & Z_{nn}
\end{bmatrix}
\begin{bmatrix}
v_1 \\
v_2 \\
i_1 \\
\vdots \\
i_n
\end{bmatrix}.
$$

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All the branch currents and, in addition, the unknowns $v_1$, $v_2$, $\dot{v}_1$, $\dot{v}_2$ can then be solved from (18).

It is worth mentioning that the approach can be applicable even when the isolation resistors cannot be modeled by a simple resistance. In those cases, the resistance $R$ in (18) should be replaced by the equivalent impedance of the isolation resistors, which can usually be obtained by measurement at the frequency of interest.

IV. NUMERICAL EXAMPLES

A. Four-Way Radial Power Divider

To demonstrate the validity of the hybrid DFT-MPIE technique, a four-way radial power divider is analyzed and compared with experimental results. The discretization of the four-way radial power divider is shown in Fig. 4, along with other relevant parameters. The power divider is designed and measured on a microstrip structure with the substrate dielectric constant 4.33 and height 1.5 mm. The output ports are 50-Ω transmission lines, which are realized by microstrip lines of width 2.89 mm.

The scattering parameters of the four-way power divider with the outer radius 26.6 mm and the sector angle 27.4° are shown in Fig. 5. For this problem, it is still possible to solve the original problem without applying the DFT simplification. The results, although not shown here, are found to be the same as those obtained by the present approach. The experimental data are also shown in the figure. They are in good agreement with the computed results.

As shown in Fig. 4(b), the number of cells of one element in the discretization is 47, and the number of branches between cells is 64. If we directly solve the original four-way radial power divider, the number of the cells will increase up to 188, and the number of the branches between cells will also increase up to 256. It has been found that the required core memory is 1.7 Mb versus 8.3 Mb, while the CPU time is 22.3 s versus 85.2 s per frequency using a Sparc-10 workstation. The overall improvement is about 4.9 times in the memory storage and 3.8 times in CPU time.

B. 14-Way Radial Power Divider

The second example considers a 14-way radial power divider. The configuration and discretization of the 14-way radial power divider are shown in Fig. 6, along with other relevant parameters. The input signal of the central port is radially divided into seven ports and the sectorial component divides each port into two ports. Although the power divider has 14 output ports, the periodicity is only seven. The power divider is designed and measured on the same substrate as the four-way radial power divider.

The scattering parameters of the 14-way radial power divider are shown in Fig. 7. The calculated scattering parameters are found to be in good agreement with the measured data. As shown in Fig. 6(b), there are 89 discretized cells and 108 branches in one element. Using the same division, the number of the cells and branches of the original 14-way radial power divider will be 623 and 756, respectively. It has been found that the required core memory is 4.5 Mb versus 32.9 Mb, while the CPU time is 165.6 s versus 2495.6 s per frequency using a Sparc-10 workstation. The overall improvement is about 7.3
Fig. 6. A 14-way radial power divider. (a) Geometry. (b) Discretization.

Fig. 7. Scattering parameters of the 14-way radial power divider depicted in Fig. 6. (a) Return loss $S_{11}$, $S_{21}$, $S_{22}$. (b) Isolation parameters $S_{2,1}$, $S_{2,3}$, $S_{2,4}, \ldots, S_{2,9}$.

Fig. 8. A 14-way radial power divider with isolation resistors. (a) Geometry. (b) Discretization.

C. 14-Way Radial Power Divider/Combiner with Isolation Resistors

As shown in Fig. 7, the isolation between ports 2 and 3 of the 14-way power divider exhibits poor performance in the desired frequency band of 3–3.5 GHz. It is beneficial to introduce proper isolation resistors for output match and isolation. In this example, resistors $R_1$ are placed at a distance from the center-fed probe and resistors $R_2$ are located at a distance from the circular edge of the sector, as shown in Fig. 8(a). The configuration and discretization of this power divider/combiner are also shown in Fig. 8(b). The isolation resistors $R_1$ and $R_2$ are chip resistors of 30 and 100 $\Omega$, respectively.

Fig. 9 shows the computed reflection coefficients $S_{1,1}$, $S_{2,2}$, and coupling coefficients $S_{2,j}$ ($j = 1, 3, 4, \ldots, 9$). A significant improvement in the isolation over that in Fig. 7 can be noticed in the operating frequency band. The measured data for this power divider/combiner are also shown in Fig. 9. It is found that the computed and measured results are in reasonable agreement. Here, the isolation resistors $R_1$ and $R_2$ are assumed to be of fixed pure resistance values versus the frequency in the numerical analysis. This may result in a slight discrepancy between the simulated and measured results.

The dependence of the return loss $S_{2,2}$ and isolation $S_{2,3}$ on isolation resistors $R_1$ and $R_2$ has been studied by numer-
Fig. 9. Scattering parameters of the 14-way radial power divider depicted in Fig. 8. (a) Return loss $S_{1,1}$, $S_{2,2}$. (b) Isolation parameters $S_{2,1}$, $S_{2,3}$, $S_{2,4}$, $\ldots$, $S_{2,9}$.

dical simulations, and the results are presented in Fig. 10. In Fig. 10(a), the results indicate that both $S_{2,2}$ and $S_{2,3}$ improve monotonically when the isolation resistor $R_2$ is varied from 50 to 10 $\Omega$, while $R_1$ is fixed at 100 $\Omega$. In Fig. 10(b), the return loss $S_{2,2}$ improves slightly and the isolation $S_{2,3}$ improves monotonically when the isolation resistor $R_1$ varies from 120 to 80 $\Omega$, while $R_2$ is fixed at 30 $\Omega$. In view of the above observations, one may expect that the return loss $S_{2,2}$ can be improved by adjusting the isolation resistor $R_1$ and the isolation $S_{2,3}$ can be approximately controlled by both $R_1$ and $R_2$.

D. Improvement of Matrix Size and Speed Ratio

For an $N$-way radial power divider/combiner, it assumes that $\Delta$ is the number of the branches between cells for one element. The matrix size will be $(\Delta N)^2$ and $N \Delta^2$ for the original problem and eigenproblems, respectively. As a result, the improvement factor in memory requirement will be

$$A = \frac{(\Delta N)^2}{N \Delta^2} = N.$$  \hspace{1cm} (19)

In case of large $\Delta$, the CPU time is dominated by the solution of the matrix equation, which is proportional to the cube of the number of unknowns. The speed ratio improvement $S$ by the present hybrid DFT–MPIE approach is asymptotically given by

$$S = \frac{(\Delta N)^3}{N \cdot \Delta^3} = N^2.$$  \hspace{1cm} (20)

Fig. 10. Effects on the return loss $S_{2,2}$ and isolation $S_{2,3}$ due to the isolation resistors $R_1$ and $R_2$ of the 14-way power divider/combiner shown in Fig. 8.

V. CONCLUSIONS

In order to precisely and efficiently extract the scattering parameters of the radial power divider/combiner, especially the isolation between the output ports, a hybrid DFT–MPIE technique is applied to account for fringing effects, conductor loss, radiation, and surface wave leakage. The size reduction of the matrix elements is achieved by using a DFT. The original problem is divided into $N$ eigenproblems such that the number of unknowns is reduced by a factor of $N$ and the overall computational load by $N^2$. The $N$-way radial power dividers/combiners with isolation resistors are used as illustrative examples. Numerical results demonstrate the numerical efficiency and accuracy of the present approach in the analysis and design of $N$-way radial power dividers/combiners.

REFERENCES

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