Mutual Coupling Compensation in Transmitting Arrays of Thin Wire Antennas

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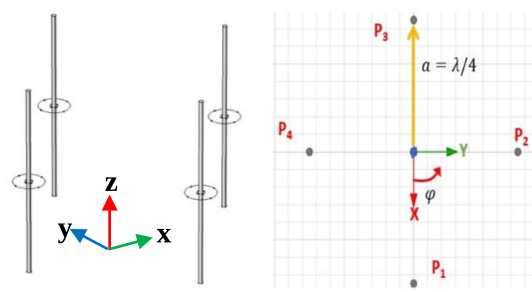
Abstract **─** This paper presents a numerical technique for the compensation of mutual coupling in transmitting arrays of thin wire antennas. The Method of Moments is used to compute the scattering parameters of the array. Using these parameters and the original excitations, a set of new excitation values are computed. The new excitation values compensate the effect of the mutual coupling. Computed results are given for linear, circular, and 3-Dimensional arrays. Uniform and binomial excitations are studied. Results obtained are compared with other available data and very good agreement is observed.

Index Terms **─** 3-Dimensional arrays, circular arrays, dipole antenna arrays, linear arrays, Method of Moments, mutual coupling compensation, planar arrays, transmitting arrays.

I. INTRODUCTION

In a simplistic array design approach the classical pattern multiplication method is used. In this method the radiation pattern of an array of identical antenna elements is the product of an element pattern and an array factor. The element pattern is the pattern of an isolated element with its center at the origin. This element is assumed to be excited by a unit voltage. The array factor is a sum of fields from isotropic point sources located at the center of each array element and is found from the elements excitation (amplitude and phase) and their locations [1]. Therefore, in the classical array theory, it is assumed that all the elements of the array have equal radiation pattern, in other words, the coupling between individual elements is ignored. For a practical array, this is not true since mutual coupling causes each element to see a different environment and consequently has a different radiation pattern from its neighboring elements.

Mutual coupling between the antenna elements in an antenna array is a classical problem, which is responsible for the degradation of array performance [2–4]. The purpose of this work is to include the effect of mutual coupling and still produce the same pattern as the array factor method.



(a) (b)

Fig. 1. (a) Circular array of four thin wire antennas with length λ/2. (b) Top view of the system.

However, this requires replacing the original excitations of the array with a new set of excitations. These new excitation values can be found if the scattering matrix of the array is measured or computed.

Here, the Method of Moments [5] is used to compute the scattering parameters of the array. For example, consider the circular array of four thin halfwave dipoles as shown in Fig. 1 (a). The radius of the circle is λ/4 and the radii of the dipoles are λ/200. Assume that the antennas are excited with the same magnitude and with phase progression of 30^{o}. We call these excitations as the original excitations. The resulting pattern computed using these original excitations and classical array theory is shown in Fig. 2. We will call this result as the desired pattern. It is incorrect because the mutual coupling effects are not included in the classical approach. When mutual coupling effects are included, the pattern shown in Fig. 3 is obtained. We call this pattern the practical pattern and it is quite different than the desired pattern given in Fig. 2.

Using the method described in Section II, a set of new excitations for the array is computed. The true practical pattern of the array computed using these new excitations and by including mutual coupling effects is shown in Fig. 4. It is quite similar to the desired pattern

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of Fig. 2. In the next section, we will summarize a numerical method for computing the new excitations.

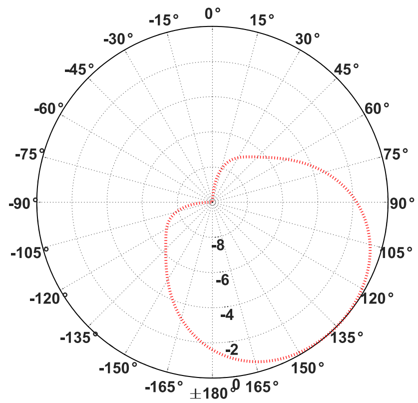

Fig. 2. The *desired pattern* (theoretical pattern using array factor) of circular array of four halfwave dipole antennas excited with uniform phase progression.

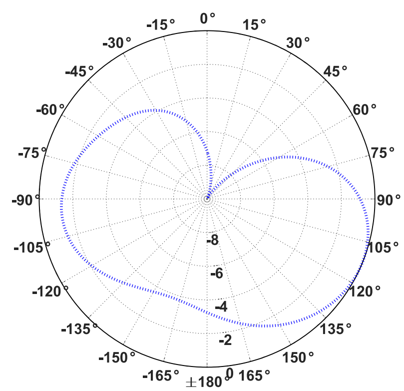

Fig. 3. The *practical pattern* of circular array of four halfwave dipole antennas excited with uniform phase progression.

Several techniques to reduce mutual coupling and improve the isolation between antennas have been investigated [6–11]. Some of these include decoupling networks (DN) which use lumped elements and hybrid couplers to reduce mutual coupling. Another common technique uses the defected ground plane structure (DGS) in which the ground plane is modified by introducing slits of different shapes.

Inductance and capacitance are introduced using electromagnetic band gap (EBG) structures to create forbidden band of frequencies which helps in isolation of the antennas. Metamaterials are also used due to existence of band gaps in their frequency responses. Neutralization lines are also used to feed reverse current to lessen the amount of coupled current. All the techniques mentioned above increase complexity of the network in one way or another. Physical modification of the structure is required, e.g., introduction of lumped elements, parasitic elements, and sometimes special materials may be required to reduce the mutual coupling.

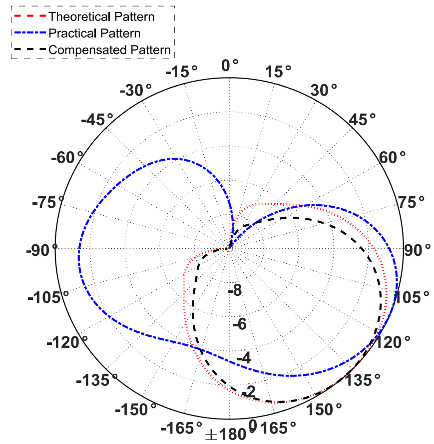

Fig. 4. Theoretical pattern obtained using the classical pattern multiplication method and the original sources (*solid*), the true practical pattern obtained using the original sources (*dashed-dotted*), and the compensated pattern obtained using the compensated voltages (*dashed*).

This paper uses a non-invasive method in which the physical structure or design of the antenna is not changed. Only the excitation voltages of the antennas are changed and these new excitation voltages which are called *compensated voltages* produce a radiation pattern which is similar to the pattern obtained from pattern multiplication method with the original voltages. The pattern multiplication method uses the original excitations and assumes no mutual coupling between the elements. The resulting *theoretical desired pattern* is usually much different than the actual practical pattern as shown in the above example. On the other hand, the new pattern, obtained by using the compensated voltages and by assuming the existence of the mutual coupling does agree with the *theoretical desired pattern*. Only thin wire dipole antennas of lengths less than or equal to $\lambda/2$ are considered. Such antennas are usually called as single mode antennas. The individual elements in the array may or may not be identical in dimensions (length, wire radius). A number of similar methods have been suggested for the compensation of mutual coupling [12–14]. This paper is an extension of the work published in [15]. Here, in addition to simple linear arrays we consider circular and 3-Dimensional arrays. We also consider linear arrays with non-uniform excitations.

II. COMPUTATION OF COMPENSATED VOLTAGES

Consider a transmitting array of N identical elements. Each element is assumed to be driven by a voltage source, V_{gk}, for $k = 1,2,...,N$. All sources are assumed to have an identical internal impedance of Z_o. When mutual coupling is ignored, the input impedance, Z_{in}, of all antennas are the same. Then, the current entering the k^{th} element is simply,

$$I_k = \frac{V_{g_k}}{Z_o+Z_{in}}. \tag{1}$$

The pattern of the k^{th} element is basically fixed by this I_k. The array multiplication method assumes this pattern for the k^{th} element. This is the ideal desired pattern.

In a practical array, because of the existence of mutual coupling, the actual input impedance of the k^{th} element, $Z'_{k_{in}}$, is not anymore equal to Z_{in}. Hence, the actual current entering the k^{th} element is now given by,

$$I'_k = \frac{V_{g_k}}{Z_o+Z'_{k_{in}}}. \tag{2}$$

Since this current is different than I_k, the pattern of this antenna will be different. However, if instead of V_{gk} we use a new voltage,

$$V'_{g_k} = \frac{V_{g_k}(Z_o+Z'_{k_{in}})}{Z_o+Z_{in}}, \tag{3}$$

then, the current entering the k^{th} element will be equal to I_k and hence the pattern of the k^{th} element will be the same as the desired pattern. It is not easy to measure $Z'_{k_{in}}$ directly. However, the scattering matrix of the array can be computed or measured easily. As shown in Appendix, the new compensated voltages can be obtained in terms of the original voltages and the S-matrix as follows,

$$[V'_g] = \frac{2Z_0}{Z_o+Z_{in}}\{U-S\}^{-1}[V_g]. \tag{4}$$

Where U represents an $N \times N$ unit matrix, S is the $N \times N$ matrix of the scattering parameters of the array, $[V_g]$ and are $[V'_g]$ column vectors showing the original and compensated voltages, respectively.

In this work the scattering matrix of the array is computed using the Method of Moments. The feeds are modeled by magnetic frill currents. The unknown currents are approximated by piece-wise sinusoidal functions and Galerkin's method is used for matching [16]. Using the moment matrix and its inverse, one can compute the open circuit parameter matrix, short circuit parameter matrix and hence the scattering parameter matrix of the array. The details are available in [17–19].

III. NUMERICAL RESULTS

Results are given for three different arrays: (i) Circular array of four wire antennas, (ii) Linear array of five wire antennas, and (iii) 3-Dimensional array of seven wire antennas. For each case three different patterns are computed. The *theoretical pattern* is computed using pattern multiplication method which uses the original voltages and assumes no mutual coupling. The *practical pattern* is computed using the original voltages but including the effect of mutual coupling. Finally, the *compensated pattern* is computed using the newly computed voltages and assuming the presence of mutual coupling. Note that all patterns computed in this work are normalized. The results are verified using COMSOL Multiphysics® [20].

A. Circular array of four antennas

In this section, four halfwave dipole antennas have been arranged in a circular array of radius $a = \lambda/4$. The elements are identical with radius $\lambda/200$. The number of expansion functions used for all the MoM simulations are 63 per antenna. In order to find the array factor for the circular array of four dipole antennas, the antennas are replaced by four isotropic sources at their centers, as shown in Fig. 1 (b). The radiation pattern of such an array can be found easily using pattern multiplication method. The element pattern of a halfwave dipole antenna in the H-plane is a unit circle. Since the total pattern is the product of the element pattern and array factor, then the total pattern, in this case, is equal to the array factor. The array factor is given by,

$$AF = \sum_{n=1}^{4} V_n e^{j\beta(x_n \sin\theta\cos\varphi + y_n \sin\theta\sin\varphi)}, \tag{5}$$

where β is the wave number and V_n is the complex input voltage at n^{th} antenna.

A.1. Uniform excitation of circular array

In the first case we excite the array uniformly with $V = 1$ V. The compensated voltages were computed to be equal to $V' = 1.75\angle 45^o$ for each antenna element. Figure 5 shows, the three computed patterns in the H-plane. At first sight, it might be surprising to see that the theoretical and the practical patterns are identical, as if there was no mutual coupling between the elements. Of course there is mutual coupling as indicated by the compensated voltage being different than the original one, however, due to the symmetry of the structure and uniform excitations the effect of mutual coupling is automatically compensated. This can be seen by examining the current distributions on the antennas as shown in Fig. 6. Note that although the magnitude of the currents is different, their distribution and hence their radiation patterns are identical.

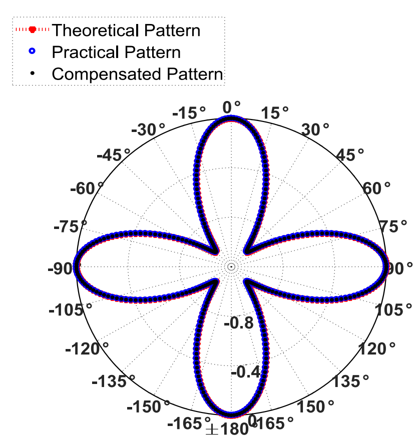

Fig. 5. Radiation pattern (H-plane) for four element circular array of halfwave dipole antennas with uniform excitation.

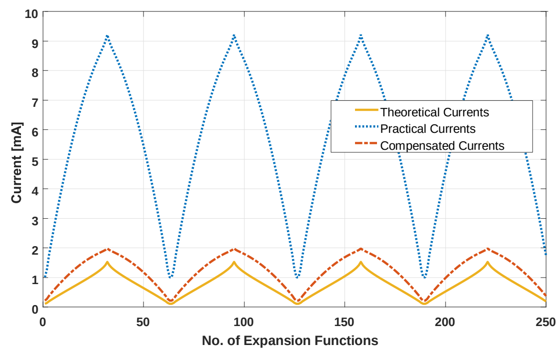

Fig. 6. Current distribution for uniformly excited four halfwave dipole antennas arranged in a circular array.

A.2. Non-uniform excitation

In this case the elements (P_1, P_2, P_3, and P_4) in Fig. 1 (b) are assumed to be excited with voltages $V_1 = 1\angle 0^o$, $V_2 = 1\angle 30^o$, $V_3 = 1\angle 60^o$, and $V_4 = 16\angle 90^o$, respectively. The computed compensated voltage values are $V_1' = 3.06\angle 78^o$, $V_2' = 2.29\angle 82^o$, $V_3' = 1.32\angle 134^o$, and $V_4' = 0.29\angle -92^o$. The three computed patterns are shown in Fig. 4. The current distribution on the antennas are shown in Fig. 7. It can be seen that the currents are not symmetric which explains the significant difference in the patterns of Fig. 4. Figure 8 compares our computed practical patterns with those of COMSOL and excellent agreement is observed. Figure 9 compares three patterns computed by (i) using (5), (ii) using MoM with compensated voltages, and (iii) using COMSOL with compensated voltages.

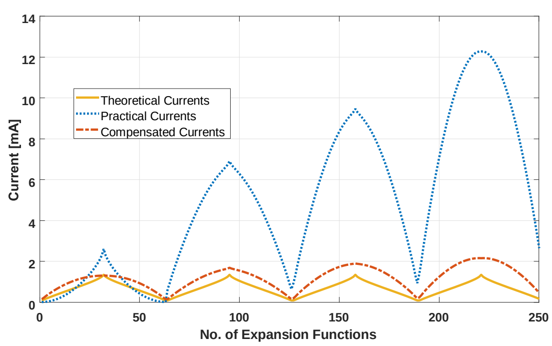

Fig. 7. Current distribution of circular array of four halfwave dipole antennas with a progressive phase excitation.

IV. LINEAR ARRAY OF FIVE ANTENNAS

Here we consider a linear array of five identical halfwave dipoles as shown in Fig. 10 with inter-element spacing of 0.1λ. A binomial excitation of the linear array with $V_1 = 0.25\angle 180^o$, $V_2 = 1\angle 90^o$, $V_3 = 1.5\angle 0^o$, $V_4 = 1\angle 90^o$, and $V_5 = 0.25\angle 180^o$ was used. Figure 11 compares the theoretical and the practical patterns for this array.

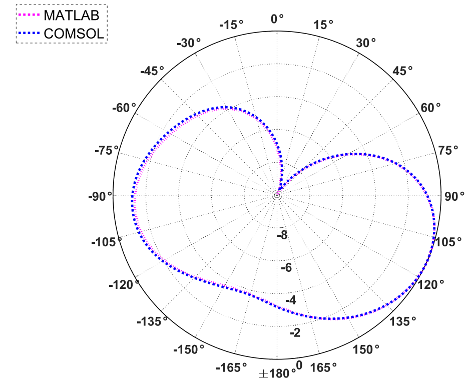

Fig. 8. Comparison of practical pattern for circular array in H-plane.

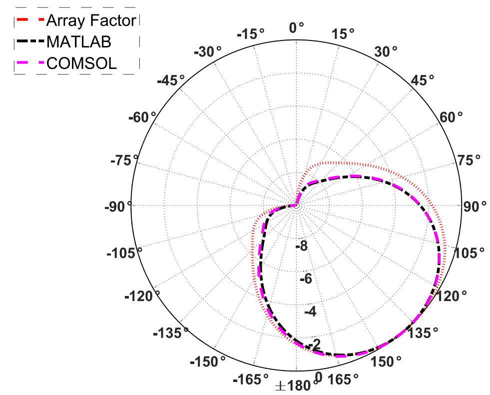

Fig. 9. Comparison of compensated pattern for four element circular array of halfwave dipole antennas (H-plane).

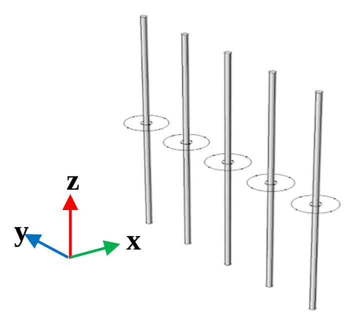

Fig. 10. Linear array of five halfwave dipole antennas uniformly spaced with a separation of 0.1λ.

It is seen that because of mutual coupling the practical pattern is much different than theoretical pattern although both methods use the same original excitations. The computed compensated voltage values are $V_1' = 2.05\angle 84^o$, $V_2' = 2.47\angle 126^o$, $V_3' = 3.86\angle 135^o$, $V_4' = 2.47\angle 126^o$, and $V_5' = 2.05\angle 84^o$. Figure 12 compares the compensated pattern, computed using these new voltage values, with the theoretical pattern computed using array factor method with original voltages. It is observed that the use of the new voltages has compensated the effect of mutual coupling to a considerable extent.

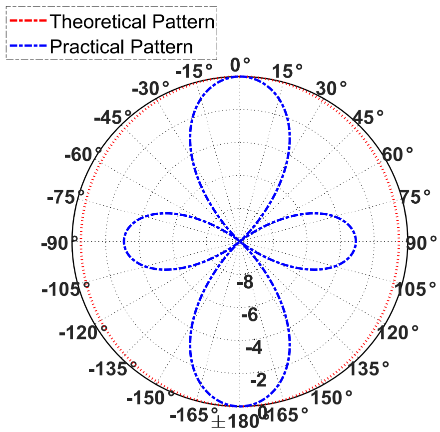

Fig. 11. Comparison of theoretical pattern and practical pattern for the five element linear dipole array using binomial excitation (H-plane).

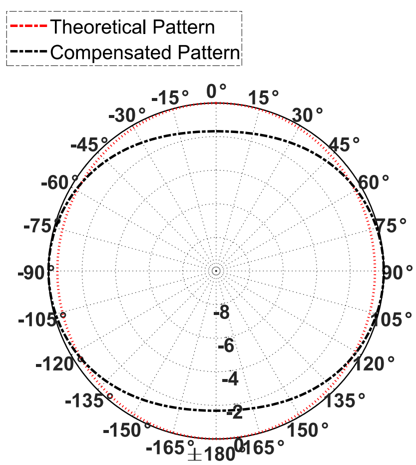

Fig. 12. Comparison of theoretical pattern and compensated pattern for the five element linear dipole array (Fig. 10) using binomial excitation (H-plane).

V. COMPENSATION OF 3-DIMENSIONAL ARRAYS

In this section, a 3-Dimensional array is studied. As shown in Fig. 13 (a), seven dipoles are arranged in two circles.

Four dipoles are placed along a circle in the lower ring and three dipoles in the upper circular ring which is displaced along z-axis. The dipoles are identical, having length L=λ/2 and radius a=λ/200. The centers of the dipoles in the lower ring are at z=0 plane whereas the centers of the upper dipoles are at z=λ/4 plane. When the elements of the array shown in Fig. 13 (b) are excited by $V_1 = 1\angle 0^o$, $V_2 = 0.5\angle 180^o$, $V_3 = 1.5\angle 0^o$, $V_4 = 0.75\angle 180^o$, $V_5 = 1\angle 0^o$, $V_6 = 0.5\angle 0^o$, and $V_7 = 0.75\angle 0^o$, then the computed compensated voltages are given by $V_1' = 4.9\angle 89.6^o$, $V_2' = 5.8\angle -94^o$, $V_3' = 6.3\angle 87.8^o$, $V_4' = 5.12\angle -130^o$, $V_5' = 3.7\angle 82.5^o$, $V_6' = 4.56\angle -128.6^o$, and $V_7' = 2.48\angle 100.4^o$. The patterns in the three principle planes can be seen in Figs. 14, 15, and 16. It is seen that the newly computed voltages compensate the effect of mutual coupling to a large extent.

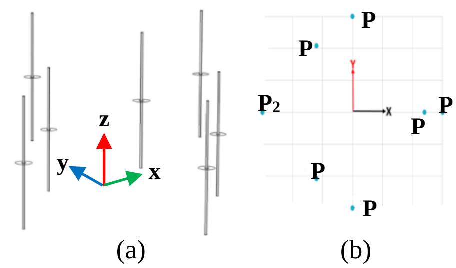

Fig. 13. (a) Three-dimensional dipole array with seven elements. Four dipoles are placed along a circle of radius 0.5λ and three dipoles along a radius of 0.4λ which are moved in the z-direction by λ/4. (b) Top view of the system.

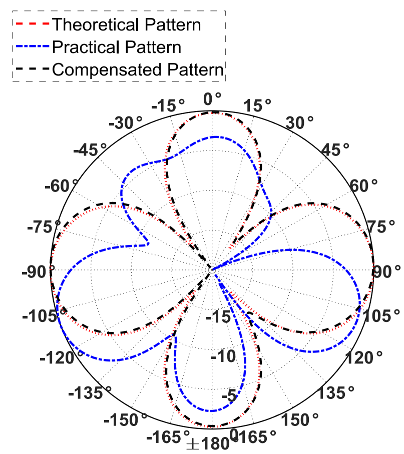

Fig. 14. XY-plane pattern for 3-dimensional array.

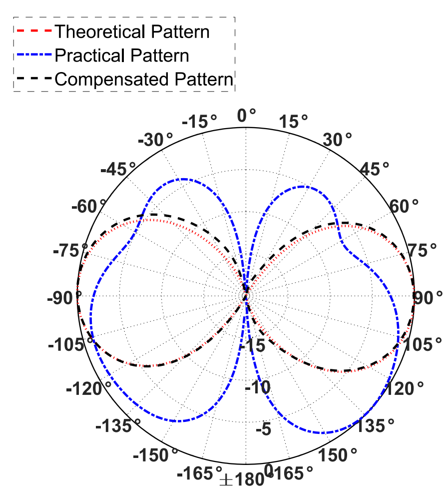

Fig. 15. YZ-plane pattern for 3-dimensional array.

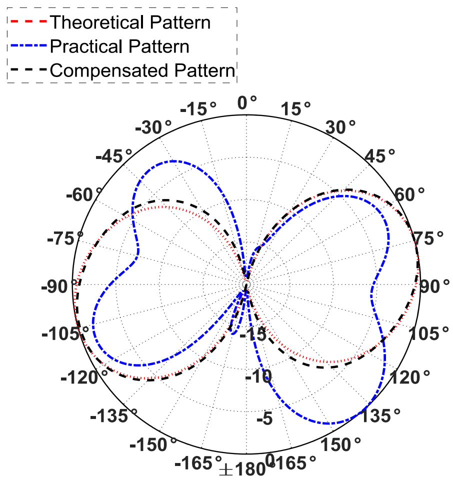

Fig. 16. XZ-plane pattern for 3-dimensional array.

VI. CONCLUSION

In this work the effect of mutual coupling in transmitting arrays of thin wire antennas has been effectively compensated. This is established without changing the antenna structure but only modifying the excitation voltages. These modified excitation voltages were computed using the scattering matrix of the array which was computed using a simple moment method technique. Three different arrays and different methods of excitations were considered. In some cases, the effect of mutual coupling was drastic compared to others. The effect of mutual coupling was successfully reduced in all cases. The patterns computed were verified with COMSOL. The limitation of this method is that it only applies to transmitting arrays. The extension of this simple method to receiving antenna array is not straight forward. In the transmitting case the excited ports are actual physical ports, whereas in the receiving case the excitation source is far away shown by the $(N+1)^{th}$ port which brings some complications. We are in the process of developing a simple method for the receiving antenna array.

ACKNOWLEDGMENT

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A. APPENDIX

Using transmission line theory, we know that the current entering the k^{th} antenna can be expressed as,

$$I_k = \frac{V_k^+ - V_k^-}{Z_0}, \quad \text{(A.1)}$$

where V_k^+ is the forward (incident) voltage entering the k^{th} antenna of the N-port network defined by,

$$V_k^+ = \frac{V'_{gk}}{2}, \quad \text{(A.2)}$$

and V_k^- is the reflected voltage from the k^{th} antenna It can be written as,

$$\begin{aligned} V_k^- &= S_{k1}V_1^+ + S_{k2}V_2^+ + \cdots S_{kN}V_N^+, \\ &= (S_{k1}V'_{g1} + S_{k2}V'_{g2} + \cdots S_{kN}V'_{gN})/2. \end{aligned} \quad \text{(A.3)}$$

Where, S_{ij} is the element of the scattering matrix for the system. Substituting the values of (2) and (3) in (1) we get,

$$\begin{aligned} I_1 &= (V_1^+ - V_1^-)/Z_0, \\ &= \frac{V'_{g1}}{(2Z_0)} - (1/Z_0)\{S_{11}V_1^+ + S_{12}V_2^+ \ldots S_{1N}V_N^+\}, \\ &= \frac{1}{(2Z_0)}\{V'_{g1} - S_{11}V'_{g1} - S_{12}V'_{g2} \ldots S_{1N}V'_{gN}\}, \quad \text{(A.4)} \end{aligned}$$

$$I_2 = \frac{1}{(2Z_0)}\{V'_{g2} - S_{21}V'_{g1} - S_{22}V'_{g2} \ldots S_{2N}V'_{gN}\}, \quad \text{(A.5)}$$

$$I_N = \frac{1}{(2Z_0)}\{V'_{gN} - S_{N1}V'_{g1} - S_{N2}V'_{g2} \ldots S_{NN}V'_{gN}\}. \quad \text{(A.6)}$$

The above equations can be written in matrix form as,

$$\begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix} = \frac{1}{2Z_0}\begin{bmatrix} V'_{g1} \\ V'_{g2} \\ \vdots \\ V'_{gN} \end{bmatrix} - \frac{1}{2Z_0}\begin{bmatrix} S_{11} & S_{12} & \ldots & S_{1N} \\ S_{21} & S_{22} & \ldots & S_{2N} \\ \vdots & \vdots & \vdots & \vdots \\ S_{N1} & S_{N2} & \ldots & S_{NN} \end{bmatrix}\begin{bmatrix} V'_{g1} \\ V'_{g2} \\ \vdots \\ V'_{gN} \end{bmatrix},$$

or in short hand notation as,

$$[I] = \frac{1}{2Z_0}\{U - S\}[V'_g]. \quad \text{(A.7)}$$

Here, $[I]$ is the N×1 column vector of the desired input currents. U is N×N unit matrix and $[V'_g]$ is the N×1 column vector of the desired compensated source voltages feeding the antennas. Then, the desired compensated source voltages (in the presence of mutual coupling) are given by,

$$[V'_g] = (2Z_0)\{U - S\}^{-1}[I], \quad \text{(A.8)}$$

or

$$[V'_g] = (2Z_0)\{U - S\}^{-1}\begin{bmatrix} \frac{V_{g1}}{Z_0+Z_1} \\ \frac{V_{g2}}{Z_0+Z_2} \\ \vdots \\ \frac{V_{gN}}{Z_0+Z_N} \end{bmatrix}.$$

Where, Z_i, is the input impedance of the i^{th} element. When the elements are identical (A.8) reduces to (4).

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