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A NEW METHOD OF PATTERN SYNTHESIS WITH LOW SIDELOBES OVER A WIDE FREQUENCY RANGE

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1. Introduction

Wideband sidelobe suppression methods have been needed for TV ghost reduction in receiving antenna fields. This paper describes a new method to suppress sidelobes in over a octave frequency range. Sidelobe reduction, in this method, is made by synthesizing frequency-independent radiation nulls in (n-1) desired directions. It is known that radiation nulls of (n-1) points can be given with an equally-spaced n-element array antenna(1). However, no one has reported a feeding method for realizing these nulls over a wide frequency range. In this paper, a novel method of realizing wideband (n-1) null points by an n-element array is presented, under the condition of no mutual coupling. It is also shown that the method is applicable to sidelobe suppression by a practical endfire array in which mutual coupling usually exists.

2. Antenna Structure and Its Theory

This antenna is constructed as shown in Fig. 1. D_i and D_s represent power dividers with a high degree of isolation and $\Phi_{i,j}$ for transmission lines as phase shifters which have electrical length $L_{i,j}$. Let p_i be the number of divided ports for power divider D_i, whose common port is connected to #i element antenna. In this antenna, the number of ports p_i is chosen according to the binomial coefficient; that is

$$p_i = {}_{n-1}C_{i-1} \tag{1}$$

Consider the equally-spaced array of n identical elements, as shown in Fig. 2. When element #1 is placed at the origin, the array factor $D(\theta, \phi)$, using the polar coordinate system, is given by

$$D(\theta,\phi) = \sum_{i=1}^{n} \dot{I}_i Z^{i-1} \tag{2}$$

$$Z = \exp[\, j\beta d \sin\theta \cdot \cos\phi \,], \quad \beta = \frac{2\pi}{\lambda} \tag{3}$$

where $\dot{I}_i$ is exitation current of #i element antenna. Equation (2) has (n-1) zeros, because it is a complex coefficient polynomial of order (n-1). If these zeros are represented by $t_1, t_2, \cdots, t_{n-1}$, the following equation results from the relation between roots and coefficients:

$$\dot{I}_i = (-1)^{n-i}\, \dot{I}_n \sum_{k_i=1}^{i} \sum_{k_{i+1}=k_i+1}^{i+1} \sum_{k_{i+2}=k_{i+1}+1}^{i+2} \cdots \sum_{k_{n-1}=k_{n-2}+1}^{n-1} (t_{k_i} \cdot t_{k_{i+1}} \cdot t_{k_{i+2}} \cdots t_{k_{n-1}}) \tag{4}$$

where t_k ($k = 1 \sim n-1$) is given by Eq. (3) designating the null direction of (θ_k, ϕ_k).

Next, consider the feeding network for the array which has frequency-independent radiation nulls in desired directions. Figure 3 indicates a section which is fed to #i element antenna from the feeding network in Fig. 1. Voltages, currents, incident waves, reflected waves and others are denoted as shown in Fig. 3. Let $S^{(i)}_{k,s}$ be an S-matrix element between a common port ⓢ and a split port ⓚ on the power divider D_s. $S'_{i,k}$ is S-matrix element between a split port ⓚ and a common port ⓘ on the divider D_i. Then, as each port takes impedance matching, exitation current $\dot{I}_i$ of #i element antenna is obtained as

$$\dot{I}_i = (-1)^{n-i} \sum_{k=1}^{p_i} \{-S'_{i,k} \cdot S^{(i)}_{k,s} \cdot \exp(-j\beta L_{i,k})\} \cdot \frac{\dot{E}_s}{2Z_0} \qquad (i=1\sim n) \qquad (5)$$

The term $(-1)^{n-i}$ on the right side will be realized when the common ports of dividers D_i are connected to the element antennas by alternately reversing phase. In order to realize the exitation currents of Eq. (4), determined from desired null directions, this array antenna attains the purpose by the feeding network, as shown in Fig. 1, which combines the excitation currents of Eq. (5). Consequently, the following equation is obtained by letting Eq. (5) be the equivalent of Eq. (4):

$$\begin{aligned} &\sum_{k=1}^{p_i} S'_{i,k} \cdot S^{(i)}_{k,s} \cdot \exp\{-j\beta(L_{i,k} - L_{n,1})\} \\ &= S'_{n,1} \cdot S^{(n)}_{1,s} \sum_{k_i=1}^{i} \sum_{k_{i+1}=k_i+1}^{i+1} \sum_{k_{i+2}=k_{i+1}+1}^{i+2} \cdots \sum_{k_{n-1}=k_{n-2}+1}^{n-1} (t_{k_i} \cdot t_{k_{i+1}} \cdot t_{k_{i+2}} \cdots t_{k_{n-1}}) \\ &\qquad\qquad (i=1\sim n-1) \end{aligned} \qquad (6)$$

If one is to design $S'_{i,k}$, $S^{(i)}_{k,s}$ for power dividers and $L_{i,j}$ of phase shifters as identically satisfying Eq. (6), as to frequency variation, it is possible to realize an array antenna which has (n-1) nulls in specified constant directions over a wide frequency range. As a result, in order to independently design power dividers and phase shifters, it is sufficient to adopt the following conditions:

$$S'_{i,k} \cdot S^{(i)}_{k,s} = \text{const.} \qquad (i=1\sim n, k=1\sim p_i) \qquad (7)$$

$$\begin{aligned} L_{i,k} - L_{n,1} = -d\{&\sin\theta_{k_i} \cdot \cos\phi_{k_i} + \sin\theta_{k_{i+1}} \cdot \cos\phi_{k_{i+1}} + \sin\theta_{k_{i+2}} \cdot \cos\phi_{k_{i+2}} + \cdots \\ &+ \sin\theta_{k_{n-1}} \cdot \cos\phi_{k_{n-1}}\} \qquad (i=1\sim n-1, k=1\sim p_i) \end{aligned} \qquad (8)$$

However, phase shifts within power dividers are treated by including them into $L_{i,j}$. This antenna can be designed using Eq. (7) and Eq. (8) under the condition of no mutual coupling.

3. Numerical Results and Considerations

The low sidelobe radiation pattern of H-plane, synthesized by allocating six nulls in the 50°, 70°, 90°, 120°, 150° and 180° directions from the x axis, is shown as curve A in Fig. 4. The antenna dimensions are described in Fig. 4. Deep nulls in H-plane are shaped frequency-independently at both 100 MHz and 200 MHz at specified angles. It is observed that the nulls keep sufficiently low sidelobes over a wide frequency range.

In a practical array, mutual coupling must be taken into consideration. Input impedances of each element in the array may differ appreciably. Curve B in Fig. 4 shows calculated results with mutual coupling. The sidelobe suppression in H-plane is much better than that of a Yagi-array with the same number of elements over a wide frequency range. If spacing d is made larger than that in this case (d = 40 cm), the pattern in curve B becomes similar to that in curve A. Self and mutual impedances were calculated by Inagaki's theory(2).

4. Conclusion

In this paper, it is shown theoretically that a broadband sidelobe suppression is feasible with an array antenna which consists of power dividers, transmission lines and radiating elements. The described synthesis method gives radiation nulls in (n-1) desired directions with equally-spaced coupling-free n-element radiators. Numerical results of 7-element array show that the method is useful for sidelobe suppression for arrays with coupled elements over a wide frequency range.

Acknowledgement

The author would like to acknowledge the advice and the encouragement of Dr. K. Nagai and Mr. Y. Mikuni.

References

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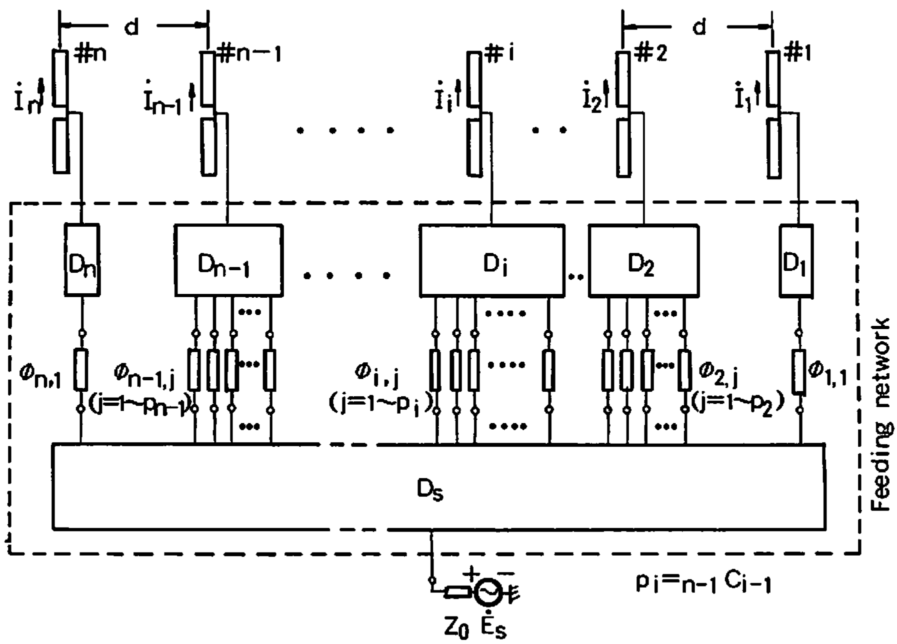

Fig. 1. Structure of array antenna with feeding network divided into binominal coefficient ports

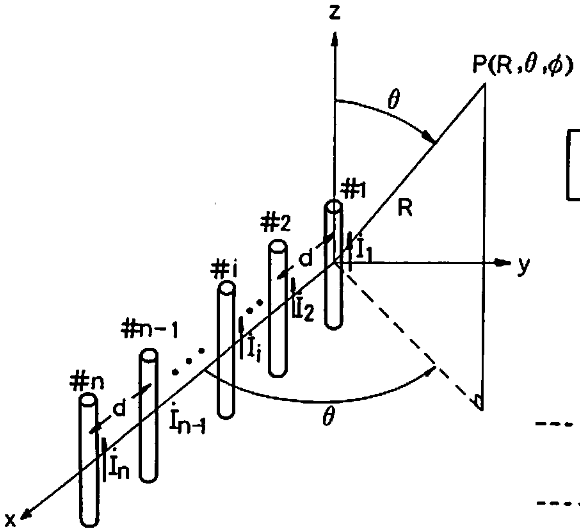

Fig. 2. N-element array antenna coordinate system

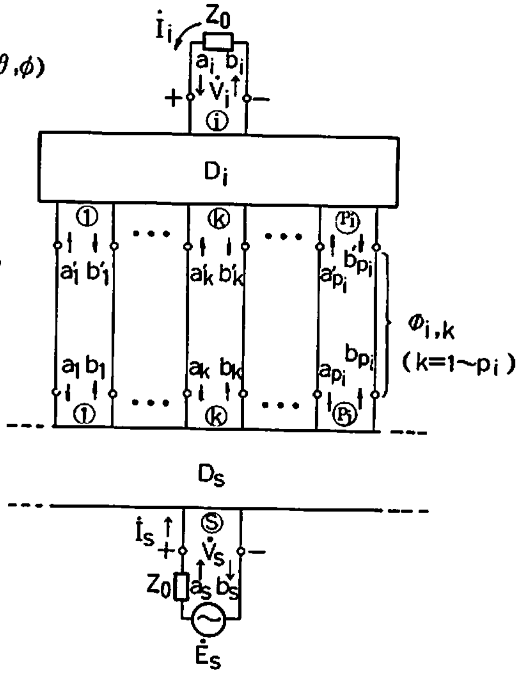

Fig. 3. Detailed section of feeding network fed to #i element antenna

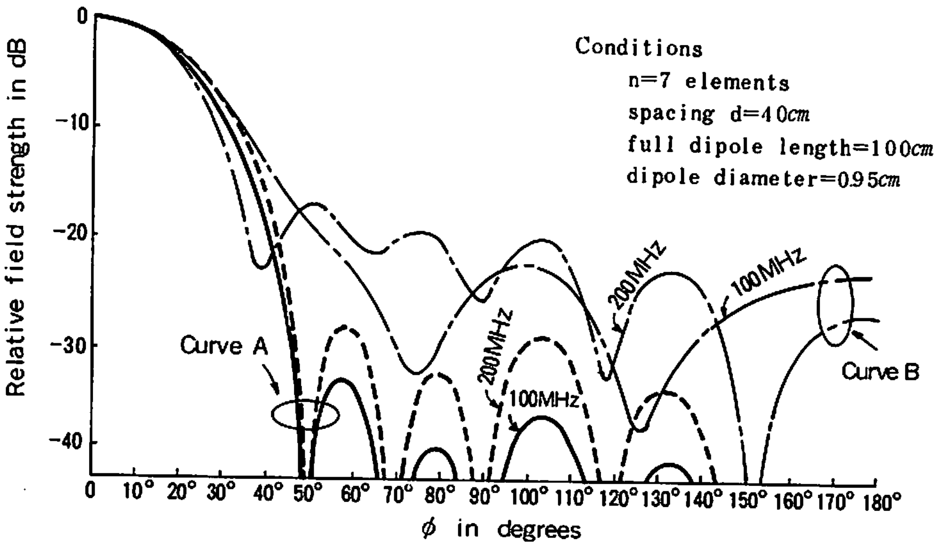

Fig. 4. An example of radiation patterns synthesized over a wideband (H-plane)