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PAPER

Analysis of Power Inversion Adaptive Array Performance by Moment Method

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SUMMARY A rigorous and systematical analysis of the performance of the power inversion adaptive array is shown where the method of moments is applied to analyze the antenna system and the electric field intensity is used as the descriptive parameter of the incident signals. This method can be easily extended to analyze adaptive arrays with arbitrary array structure and with wire antenna elements other than monopoles and dipoles. To show the adavantages of the method, the performance of adaptive arrays with mutual coupling effect among the array elements is considered and the inverted-F antennas (IFA) are used as the elements. The steering vector in the presence of mutual coupling which differs from that in the absence of the mutual coupling is derived. The output signal-to-interference-plus-noise ratio (SINR) of two kinds of adaptive arrays composed of IFA elements is computed and the importance of properly selecting steering vector is demostrated by means of the quiescent array patterns and the output SINR performance.

1. Introduction

One of the primary interests of current studies in the adaptive arrays is the interference suppression[1]. For this purpose, the power inversion adaptive array (PIAA)[2],[3] is a useful antenna system, particularly when the detailed information about the structure and the arrival direction of the desired signal are not available.

The purpose of this paper is to show an analysis of the performance of a PIAA by using the method of moments[4]. The application of the method of moments is more flexible than the use of the EMF method and has the following features: (a) it may allow more rigorous analysis for the arrays with the effect of mutual coupling; (b) the adaptive array consisting of arbitrarily shaped wire elements can be treated rigorously; and (c) the analysis is also valid for arrays with very small interelement spacing.

Another distinguishing feature of this paper to be stressed is the use of the incident electric field intensity as the descriptive parameter for the incoming signal instead of the input power which was used in most previous papers. In some cases where the mutual coupling effect is not taken into account or when only the processor performance is considered, the use of the input power is certainly a simple way to express the parameter of the arrival signals. However, since the input power is an intermediate value in analyzing the performance of an array system composed of both practical antenna elements and the processor, the use of the power of the input signal does not describe clearly the array element characteristics. In addition, in the presence of mutual coupling, it is no longer exact to express the incident signal in terms of the input power, since the actual input power of each array element is different from each other due to the effect of mutual coupling. Therefore, the incident electric field intensity is a more reasonable parameter to describe the strength of an incoming signal when a rigorous analysis of the adaptive arrays is desired.

The effect of mutual coupling on the performance of adaptive arrays has been studied previously by Leviatan et al. for an adaptive Yagi array[5], by Gupta and Ksienski for the least mean square error (LMS) and the Applebaum adaptive arrays[6], and by Dinger for a reactively steered adaptive array (RESAA)[7]. The paper written by Leviatan et al.[5] mainly discussed the frequency-dependent characteristics of the effect of mutual coupling of an adaptive Yagi array which limits the achievable cancellation performance. Their antenna configuration and the purpose are different from what this paper is concerned with. In Dinger's work[7] microstrip patches are used as the array elements and the mutual impedance values from experimental antenna measurements are used in his analysis. In Gupta and Ksienski's paper[6] the induced electromotive force (EMF) method[8] is used and the effect of mutual coupling between the open-circuited array elements was not taken into account so that some discrepancy may result between their analysis and the actual performance of the adaptive arrays. The effect of the mutual coupling between open-circuited elements has been given by Zhang et al.[9] which numerically shows the significance of the effect of the current on the open-circuited parasitic elements.

Previous analyses of the adaptive array perfor-

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mance in the presence of mutual coupling were mostly made by using the EMF method. However, the EMF method is only valid for arrays consisting of antenna elements of simple structure, such as monopoles and dipoles. On the other hand, computations by using the method of moments will lead to a simple procedure and a rigorous solution, even for an array consisting of antenna elements of a complex structure and the inter-element spacing is small. This is in contrast with the previous approximate treatments by which it is difficult to analyze an adaptive array with antenna elements of a complex structure.

As an example of antenna elements having a complex structure, the inverted-F antenna (IFA)[(10)-(12)] is used in this analysis. The practical importance of the IFA makes it to receive increasing attentions especially in mobile communications because of its low profile structure and feasibility of excellent impedance matching[(11)]. In addition, the cross-polarization performance of the IFA may contribute to improve degradation of signal reception due to the multipath fading which may be often encoutered in urban environment communications[(10),(12)]. Therefore, the IFA is useful to the mobile communications. This is the reason for taking the IFA into consideration here.

Section 2 discusses the analysis by applying the method of moments. The received voltage vector and its self-correlation matrix in the presence of mutual coupling are obtained, which become the base to compute the output parameters such as the powers and the signal-to-interference-plus-noise ratio (SINR). Section 3 states the selection of the steering vector with the consideration of mutual coupling. Section 4 gives examples of the input signal power at the antenna terminal, the quiescent array pattern and the output SINR in the absence and in the presence of mutual coupling effect. The results show the significance of mutual coupling effect and the importance of properly selecting the steering vector when mutual coupling effect is taken into account.

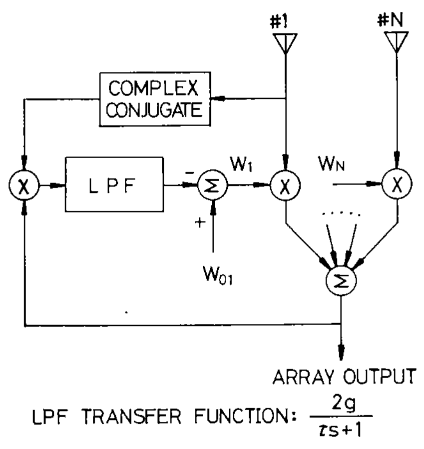
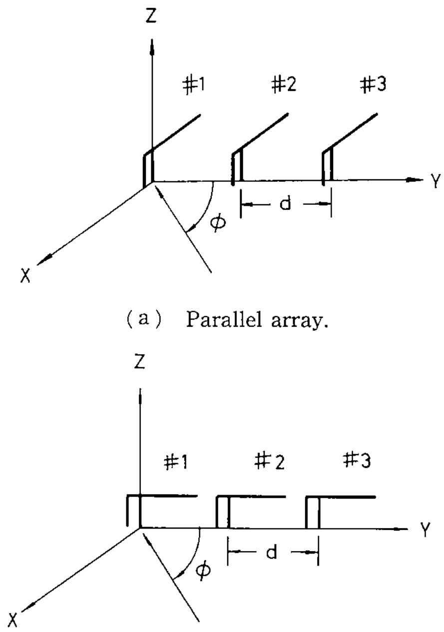


Fig. 1 A diagram of PIAA.

2. Analysis

This section shows an analysis of the performance of an adaptive array by using the method of moments. To show the usefulness of the analysis, the array system consists of antenna elements other than dipoles and monopoles is considered. For simplicity, it is assumed that the incident electric fields are vertically polarized plane waves and come from the horizontal direction.

A diagram of the PIAA system is shown in Fig. 1 and the coordinate system of the receiving array is shown in Fig. 2, where an N element linear array with an equal spacing d and the azimuthal angle ϕ taken from an endfire direction are assumed. The IFA shown in Fig. 3 is considered here as the antenna element which is located on the infinite ground plane (X-Y plane). The image of the IFA under the infinite ground plane is considered for the calculation of the IFA performance. In the calculation, the method of moments is used with piecewise sinusoidal subsectional expansion and weighting functions for the current flowing on the antenna elements. The current distribution $\bar{J}$ on the antenna elements induced by the incident electric field is expres-



(a) Parallel array.

(b) Collinear array.

Fig. 2 The coordinate system.

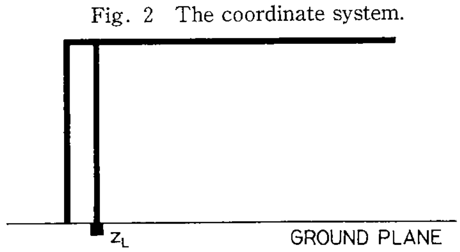


Fig. 3 An IFA element.

sed as,

$$\bar{J}=\sum_{p=1}^{M} i'_p \bar{J}_p \tag{1}$$

where M is the total number of subsectional functions defined on the antenna elements, i'_p is the current expansion coefficient to be determined, and $\bar{J}_p$ is the expansion function on the p-th subsection. The boundary-value problem for the current $\bar{J}$ can be reduced to a set of linear equations in $i'_p(p=1, 2,\cdots, M)$. The result can be written in the matrix form as Ref. (13)

$$[Z'][I']=[V']+[V'_L] \tag{2}$$

where $[Z']$ is the $M\times M$ mutual impedance matrix among the subsections, $[I']$ is the current expressed by a column vector with elements $(i'_1, i'_2,\cdots, i'_M)$, and $[V']$ is the induced voltage column vector $(v'_1, v'_2,\cdots, v'_M)$ which describes the voltages on the subsections induced by the incident electric field $\bar{E}$. The M dimensional vector $[V'_L]$ in Eq. (2) denotes the load voltage and the non-zero elements of $[V'_L]$ are the voltages at the load ports. A voltage v'_p at the p-th subsection is expressed by

$$v'_p=\int \bar{J}_p\cdot \bar{E}_p dl \tag{3}$$

where $\bar{E}_p$ is the incident electric field $\bar{E}$ at the p-th subsection, and the integral extends over the p-th subsection. Note that the integral is zero for the horizontal part of the IFA element since the vertically polarized incident electric field coming from the horizontal direction is assumed. By using the relation at the load ports,

$$[V'_L]=-[Z'_L][I'] \tag{4}$$

the current vector $[I']$ is obtained from Eq. (2) as

$$[I']=[Y'][V'] \tag{5}$$

where

$$[Y']=\{[Z']+[Z'_L]\}^{-1} \tag{6}$$

and $[Z'_L]$ is an $M\times M$ diagonal matrix where the non-zero elements correspond to the loaded impedance at the load ports.

From Eq. (5), an N-port load current vector $[I_L]$ is obtained by extracting the elements of $[I']$ corresponding to the load ports of the antenna elements. Then the voltage $[V_L]$ at the load ports can be obtained as

$$[V_L]=-[Z_L][I_L] \tag{7}$$

with

$$[Z_L]=\mathrm{diag}\{z_{L1}, z_{L2},\cdots, z_{LN}\}. \tag{8}$$

Particularly, when $z_{L1}=z_{L2}=\cdots=z_{LN}=z_L$, Eq. (7) can be simplified to be

$$[V_L]=-z_L[I_L]. \tag{9}$$

As can be understood from the above procedure, only the currents flowing at the load ports are required in the analysis. Therefore, the computation can be simplified to an N dimensional process as

$$[I_L]=[Y_0][V'] \tag{10 a}$$

or

$$[V_L]=-[Z_L][Y_0][V'] \tag{10 b}$$

where $[Y_0]$ is a $N\times M$ dimensional matrix where the rows are those of $[Y']$ corresponding to the load ports.

The above analysis can be even more simplified since the separation between the two vertical parts of an IFA is usually so close that the phase difference between the induced voltages of the two parts can be ignored. Therefore, the elements of $[V']$ become identical at the subsections defined on the vertical parts of the same IFA, and the multiplication in Eq. (10) can be condensed to an $N\times N$ dimensional calculation by adding the elements of $[Y_0]$ corresponding to the elements of $[V']$ of the vertical element of the same antenna. The simplified form of (10 b) will be

$$[V_L]=-[Z_L][Y][V] \tag{11}$$

where $[Y]$ is an $N\times N$ admittance matrix and $[V]$ is an N dimensional induced voltage vector reduced from $[V']$.

If we consider the thermal noise voltage $[V_N]$ at the antenna output, the total output voltage vector $[X]$ becomes

$$\begin{aligned}[X]&=[V_L]+[V_N]\\&=[V_{LD}]+[V_{LI}]+[V_N]\end{aligned} \tag{12}$$

where $[V_{LD}]$ and $[V_{LI}]$ are the desired and interference component of $[V_L]$, respectively.

We assume here that the thermal noise voltages of the N antenna elements have a statistically independent zero-mean Gaussian distribution with variance σ_N^2, and each of the desired and undesired signals is a narrow-band signal with zero mean. The signals and the thermal noise are statistically independent of each other. The covariance matrix $[\Phi]$ of the output voltage vector then becomes

$$\begin{aligned}[\Phi]&=E\{[X]^*[X]^T\}\\&=E\{[V_L]^*[V_L]^T\}+E\{[V_N]^*[V_N]^T\}\\&=E\{[V_{LD}]^*[V_{LD}]^T\}\\&\quad+E\{[V_{LI}]^*[V_{LI}]^T\}+\sigma_N^2[U]\end{aligned} \tag{13}$$

where $E\{\cdot\}$ denotes expectation and $[U]$ is a unit matrix.

Therefore, the steady state weight vector $[W]$ can be obtained from the control equation of the PIAA,

$$[W]=\{[U]+g[\Phi]\}^{-1}[W_0] \tag{14}$$

where g is the loop gain and $[W_0]$ is the steering vector, which will be discussed in detail in the next section. By assuming that the load impedances of all antenna elements are equal to z_L, the output powers of the desired, interference signals and noise, respectively, are given as

$$P_D=\frac{1}{2z_L}E\{|[W]^T[V_{LD}]|^2\}$$
$$=\frac{1}{2z_L}[W]^T E\{[V_{LD}]^*[V_{LD}]^T\}[W]^*,$$
$$P_I=\frac{1}{2z_L}E\{|[W]^T[V_{LI}]|^2\}$$
$$=\frac{1}{2z_L}[W]^T\{[V_{LI}]^*[V_{LI}]^T\}[W]^* \quad (15)$$

and

$$P_N=\frac{1}{2z_L}[W^T][W^*]\sigma_N{}^2$$

and the output SINR is

$$\mathrm{SINR}=\frac{P_D}{P_I+P_N}. \quad (16)$$

3. Selection of Steering Vector

Since the power inversion algorithm is usually applied in the conditions where the arrival angle of the desired signal is a priori unknown, an omni-directional quiescent receiving pattern in the horizontal plane is desired. How to achieve this in the presence of mutual coupling is discussed here.

When mutual coupling exists, two steps could be taken to form a desired receiving pattern. The first step is to determine the proper weights for the desired receiving pattern in the absence of mutual coupling and the next step is to compensate the mutual coupling effect.

When each of the antenna elements has an omni-directional pattern in the horizontal plane, one choice of the steering vector in the absence of mutual coupling may be

$$[W_{00}]=[1, 0,\cdots, 0]^T \quad (17)$$

to get an omni-directional array pattern in the horizontal plane. This idea implies that the arrival signals of the antenna elements except the first one are not received by the processor of the adaptive array system so that the array pattern is the same as that of the first antenna element.

When mutual coupling among antenna elements exists, $[W_{00}]$ in Eq. (17) will no longer provide an omni-directional quiescent pattern since the antenna elements will interact each other. Hence, in the presence of mutual coupling, a proper factor to cancel the effect of mutual coupling should be found and no output voltage is contributed by the antenna elements other than the first one.

Gupta and Ksienski[6] have presented a way to compensate the effect of mutual coupling in determining the weight vector. However, their treatment is not exact because the effect of mutual coupling among open-circuited elements is not taken into account. Particularly, for an array composed of elements other than the monopoles and dipoles, the current peaks may appear at some points other than the feed point so that the effect of mutual coupling among open-circuited elements becomes more significant and it requires to apply a more rigorous treatment.

By applying the method of moments, a perfect omni-directional quiescent pattern in the presence of mutual coupling can be obtained by ultilizing the results shown in the previous section. From Eq. (11), the array output s_0 in the quiescent condition is

$$s_0=[W_0]^T[V_L]$$
$$=-[W_0]^T[Z_L][Y][V] \quad (18)$$

If we select the steering vector as

$$[W_0]=[Z_L]^{-1}[Z][W_{00}] \quad (19)$$

Equation (18) will lead to

$$s_0=-[W_{00}]^T[V]=-v_1 \quad (20)$$

where v_1 is the first element of $[V]$. In Eq. (19), $[Z]$ is the inversion of $[Y]$, and in deriving Eq. (20), the symmetric property of $[Y]$ and $[Z_L]$ is used. The result of Eq. (20) shows that the quiescent array pattern is the same as that of the first antenna element which is omni-directional.

If a non-omnidirectional quiescent pattern is desired, for instance, in a case where an Applebaum adaptive array[14] is used, the steering vector in the presence of mutual coupling may be found by just replacing the vector $[W_{00}]$ in Eq. (19) by the one which forms the desired non-omnidirectional quiescent pattern in the absence of mutual coupling.

4. Numerical Results

In this section, a numerical example for two kinds of three-element linear arrays consisting of IFA elements operating at 974 MHz is shown to demostrate the usefulness of the method introduced in this paper. An adaptive array composed of IFA elements is difficult to

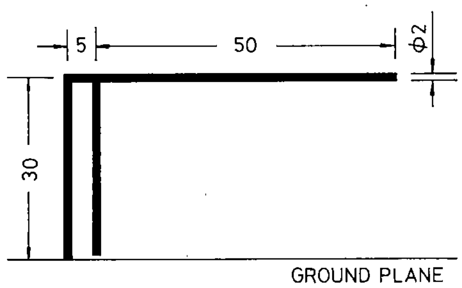


Fig. 4 The dimension of the IFA (unit : mm).

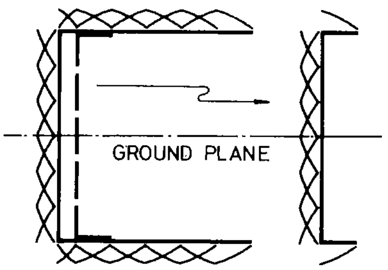
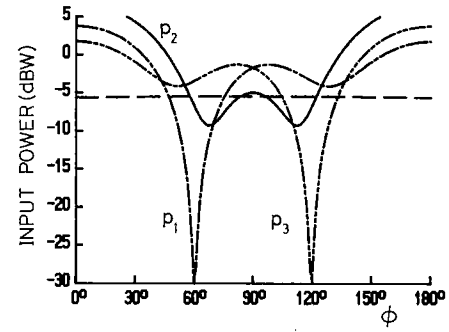
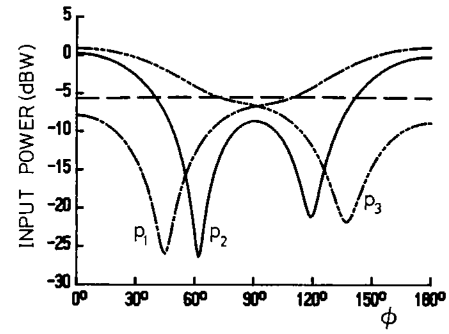


Fig. 5 Expansion functions defined on an IFA element.



(a) Parallel array.



(b) Collinear array.

Fig. 6 Input power vs. signal arrival direction (E=40 dBV/meter).

be analyzed by using the EMF method.

Figure 4 shows the dimension of the IFA element used in the computation. Each of the array elements is terminated by a 50 Ω load since the resonant IFA element at 974 MHz has the input impedance of 50 Ω. Two kinds of linear arrays, i. e., the horizontal parts of the array elements are (a) parallel and (b) collinear, are used in the computation where the interelement spacing is chosen 0.25 wavelength.

Figure 5 shows the subsections of an IFA element used in the computation. An IFA element, including its image under the infinite ground plane, is divided into 26 subsections, and 24 piecewise sinusoidal expansion functions are defined.

Figure 6 shows the input signal powers at the antenna terminals versus the direction of an incoming signal, where the incident electric field magnitude of the input signal is E=40 dB V/meter. Although the input powers are constant in the absence of mutual coupling,

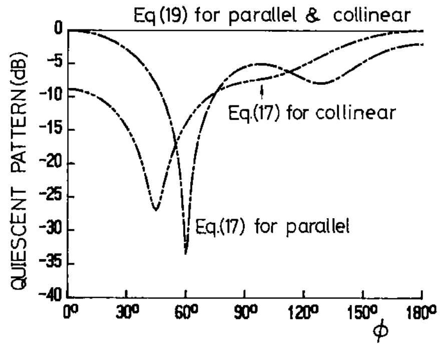
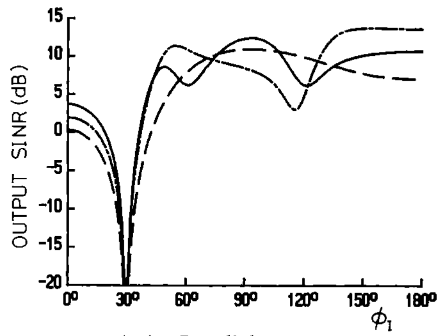
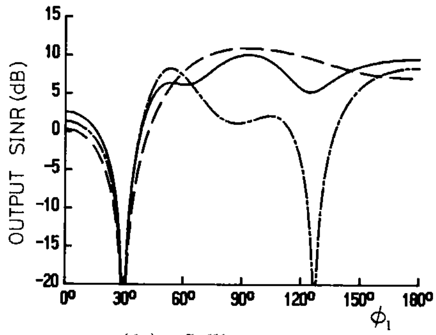


Fig. 7 Normalized quiescent array pattern by using different steering vectors in the presence of mutual coupling.



(a) Parallel array.



(b) Collinear array.

Fig. 8 Output SINR performance of PIAA using IFA elements (E_D/σ_N=40 dB/meter, ϕ_D=30°, E_I/σ_N=60 dB/meter, G=0.01, d=0.25 wavelength, ······ no mutual coupling, ——with mutual coupling using Eq. (19) as steering vector,——with mutual coupling using Eq. (17) as steering vector).

they vary greatly with respect to the signal arrival direction ϕ when the mutual coupling is taken into account. Therefore, using the input power is not exact to represent the strength of an input signal when the effect of mutual coupling is involved. In Fig. 6(b), the asymmetrical input power characteristic, which is noticeable only around the null rigions, is due to the slightly asymmetrical configuration of the vertical parts of the antenna elements.

The normalized quiescent array pattern in the presence of mutual coupling is shown in Fig. 7. The steering vector given by Eq. (19) gives an omni-directional array pattern for both the two kinds of arrays. However, when Eq. (17) is used as the steering vector, the array pattern is not omnidirectional, and the pattern actually is the same as the input power of the first antenna element, since the weights of the other two antena elements are zero. With Eq. (17) it is noticeable that the parallel and the collinear array pattern have an significant null at about 60° and 45°, respectively.

Figure 8 gives the output SINR performance of the two arrays, where one desired and one interference signal are assumed to arrive. The incident electric field magnitudes of the desired and the interference signal are $E_D/\sigma_N=40$ dB/meter and $E_I/\sigma_N=60$ dB/meter, respectively. The desired signal comes from the direction of $\phi_D=30°$ and the arrival angle of the interference signal is varied from $\phi_I=0°$ to 180°. Furthermore, the normalized loop gain ($G=g\times\sigma_N^2$) is assumed as 0.01.

When the effect of mutual coupling is ignored, the two arrays provide the same output SINR as shown in Fig. 8 since the horizontal parts of the antenna elements do not contribute to the signal receiving.

For case (a), i. e., when the antenna elements are parallel, Fig. 8(a) shows that the output SINR in the presence of mutual coupling varies around that in the absence of mutual coupling. When Eq. (19) is used as the steering vector, the maximum difference in the output SINR due to mutual coupling effect is 6.4 dB at $\phi_I=41°$, except that in the null direction of $\phi_I=30°$. The output SINR using Eq. (17) in the presence of mutual coupling varies somehow more greatly and the maximum difference of 7.2 dB in the output SINR is observed at $\phi_I=115°$ in this case.

For case (b), i. e., when the antenna elements are collinear, similar results of the output SINR as shown in Fig. 8(b) are obtained in the presence of mutual coupling when Eq. (19) is used as the steering vector. However, a deep null is formed at about $\phi_I=125°$ when Eq. (17) is used as the steering vector. The reason for the degradation of the output SINR is considered as that, when the null direction of the quiescent array pattern formed by Eq. (17) is near the direction of the desired signal, the null may shift to the direction of the desired signal when a strong interference signal arrives. Actually, it occours when $\phi_I=125°$. This fact shows the importance of properly selecting the steering vector. In the presence of mutual coupling, the steering vector should be chosen to compensate the effect of mutual coupling properly.

In the presence of mutual coupling, the two kinds of arrays provides different output SINR. When Eq. (19) is used as the steering vector, the maximum difference between the output SINR of the two arrays exceeds 2 dB. This difference is caused by the mutual coupling effect of the horizontal parts of the IFA elements. This is one of the features different from those of the monopoles and dipoles.

5. Conclusion

The performance of a power inversion adaptive array was analyzed by using the method of moments. Inverted-F antenna elements were used as an example of antenna elements having a complex configuration (not simply linear wires such as monopoles and dipoles). It was shown that the application of the method of moments along with the incident electric field intensity as the descriptive parameter of the incident signals provides a general and rigorous way to analyse the performance of an adaptive array with mutual coupling effects. It was also shown that the method is very effective to analyze the array composed of arbitarily shaped wire antenna elements. In addition, a general method to determine the steering vector in the presence of mutual coupling was obtained. Numerical results showed the significance of mutual coupling effect and the importance of properly chosing the steering vector.

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References

(1) R. A. Monzingo and T. W. Miller: "Introduction to Adaptive Arrays", John Wiley, New York (1980).

(2) C. L. Zahm: "Application of adaptive arrays to supress strong jammers in the presence of weak signals", IEEE Trans. Aerosp. & Electron. Syst., **AES-9**, 2, pp. 260-271 (March 1973).

(3) R. T. Compton, Jr.: "The power-inversion adaptive array: concept and performance", IEEE Trans. Aerosp. & Electron. Syst., **AES-15**, 6, pp. 803-814 (Nov. 1979).

(4) R. F. Harrington: "Field Computation by Moment Methods", Macmillan, New York (1968).

(5) Y. Leviatan, A. T. Adams, P. H. Stockmann and D. R. Miedaner: "Cancellation performance degradation of a fully adaptive Yagi array due to inner-element coupling", Electron. Lett., **19** (March 1983).

(6) I. J. Gupta and A. A. Ksienski: "Effect of mutual coupling on the performance of adaptive arrays", IEEE Trans. Antennas & Propag., **AP-31**, 5, pp. 785-791 (Sept. 1983).

(7) R. J. Dinger: "Reactively steered adaptive array using microstrip patch elements at 4 GHz", IEEE Trans. Antennas & Propag., **AP-32**, 8, pp. 848-856 (Aug. 1984).

(8) S. A. Schelkunoff and H. T. Friis: "Antenna Theory and Practice", Dover, New York (1964).

(9) Y. Zhang, K. Hirasawa and K. Fujimoto: "Opened parasitic elements nearby a driven dipole", IEEE Trans. Antennas & Propag., **AP-34**, 5, pp. 711-713 (May 1986).

(10) Y. Tanaka, H. Kuboyama, K. Fujimoto and K. Hirasawa: "Performance of antenna systems having cross polarization in mobile communication", 1985 Natl. Conv. Rec., IECE

Japan, S5-6.
(11) K. Fujimoto, A. Henderson, K. Hirasawa and J. R. James: "Small Antennas", Research Studies Press (1987).
(12) H. Kuboyama: "Studies on antennas having cross polarization applied for mobile communications", Ph. D. dissertation, Univ. of Tsukuba (1986).
(13) K. Hirasawa, K. Fujimoto, T. Uchikura, S. Hirafuku and H. Naito: "Power focusing characteristics of ellipsoidal reflector", IEEE Trans. Antennas & Propag., **AP-32**, 10, pp. 1033-1039 (Oct. 1984).
(14) S. P. Applebaum: "Adaptive arrays", IEEE Trans. Antennas & Propag., **AP-24**, 5, pp. 585-598 (Sept. 1976).

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