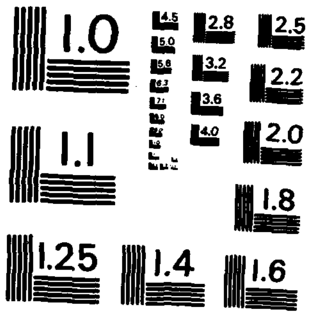
AD-A122 573 ADAPTIVE FEED CONTROL USING THE COMPLETELY OVERLAPPED SUBARRAY ANTENNA(U) ROME AIR DEVELOPMENT CENTER GRIFFISS AFB NY R E CLAPP JUN 82 RADC-TR-82-174

1 of 1

UNCLASSIFIED F/G 17/4 NL

END
DATE
FILMED
2 83
DTIC



MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

Unclassified

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE

READ INSTRUCTIONS BEFORE COMPLETING FORM

1. REPORT NUMBER: RADC-TR-82-174

2. GOVT ACCESSION NO.: A122578

3. RECIPIENT'S CATALOG NUMBER: 82-241

4. TITLE (and Subtitle): ADAPTIVE FEED CONTROL USING THE COMPLETELY OVERLAPPED SUBARRAY ANTENNA

5. TYPE OF REPORT & PERIOD COVERED: In-House Report

6. PERFORMING ORG. REPORT NUMBER:

7. AUTHOR(s): Randolph E. Clapp, 2 Lt, USAF

8. CONTRACT OR GRANT NUMBER(s):

9. PERFORMING ORGANIZATION NAME AND ADDRESS: Deputy for Electronic Technology (RADC/EEA) Hanscom AFB Massachusetts 01731

10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS: 62702F 46001401

11. CONTROLLING OFFICE NAME AND ADDRESS: Deputy for Electronic Technology (RADC/EEA) Hanscom AFB Massachusetts 01731

12. REPORT DATE: June 1982

13. NUMBER OF PAGES: 19

14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office):

15. SECURITY CLASS. (of this report): Unclassified

15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited.

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

18. SUPPLEMENTARY NOTES

Randolph E. Clapp, 2Lt, USAF
RADC/EEA

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Array antenna
Overlapped subarray antenna
Nulling
Jammer direction finding

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

The radiation pattern of the Completely Overlapped Subarray Antenna can be calculated. Knowledge of the phase and amplitude of the contribution from each element can be used to determine an appropriate weight for each element to achieve a desired pattern. The 16 feed elements on Face B of the antenna allow pattern control at 15 discrete points in the pattern. Success in using this method requires an accurate determination of the individual element contributions. Random phase and amplitude errors cannot be sensed, and

DD FORM 1 JAN 73 1473 EDITION OF 1 NOV 65 IS OBSOLETE

Unclassified

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

Unclassified
SECURITY CLASSIFICATION OF THIS PAGE(When Data Entered)

20. Abstract (Continued)

therefore cannot be compensated for. If the approximate angles of arrival of jamming signals are known, and the number of jammers is smaller than the number of elements in the locating array, a tap loop behind each element can determine the noise covariance matrix. A quiescent noise power level is assumed, and the adjoint of the matrix calculated. When discontinuities occur in the pattern, they indicate the directions of arrival for jamming signals. The pattern of the Overlapped Subarray Antenna may then be changed to place nulls in the directions of the jammers.

Unclassified
SECURITY CLASSIFICATION OF THIS PAGE(When Data Entered)

Contents

Illustrations

Adaptive Feed Control Using The Completely Overlapped Subarray Antenna

1. THEORY

The far-field radiation pattern of the linear array shown in Figure 1a is the $\left[\frac{\sin x}{x}\right]^2$ pattern of Figure 1b. This pattern is formed by the complex addition of the contributions from each of the n array elements. The contribution from the nth element (neglecting any mutual coupling between elements) in the far field is that of a single isotropic radiator.

When a phase and/or amplitude weighting taper is placed across the array, the resultant pattern can be found from the complex addition of the weighted element contributions. The element weights needed to produce a desired pattern can be calculated mathematically.

In a Completely Overlapped Subarray Antenna the situation is only slightly more complicated. Figure 2a shows one experimental version of this antenna.[1] Figure 2b shows the computer generated quiescent (unadapted) far-field pattern for this realization. The low sidelobes are achieved by placing a $\cos^2$-on-a-pedestal amplitude taper across the 16 channels of the power combiner (Face A). The use of face A for any pattern control weighting would upset this taper and raise the sidelobes. Also, the main radiating aperture (Face C) contains too many (60) elements for

(Received for publication 18 June 1982)

1. Southall, H. L. (1980) Completely-overlapped-subarray fed antenna for broadband, wide scan angle, low sidelobe radar applications, 1980 Antenna Applications Symposium, September 1980, University of Illinois, Monticello, Ill.

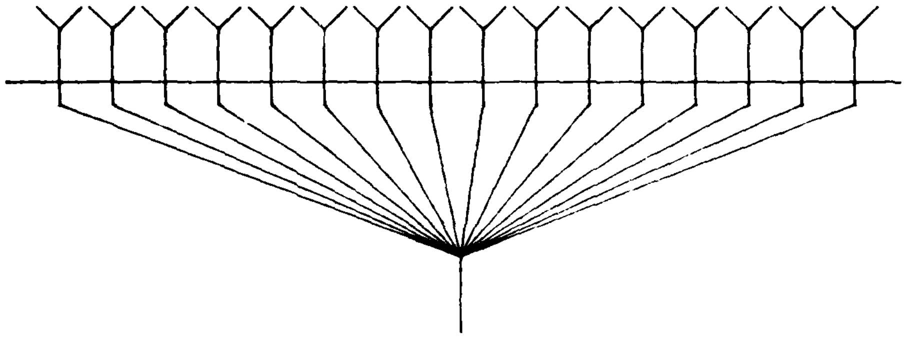


Figure 1a. Sixteen-Element Linear Array

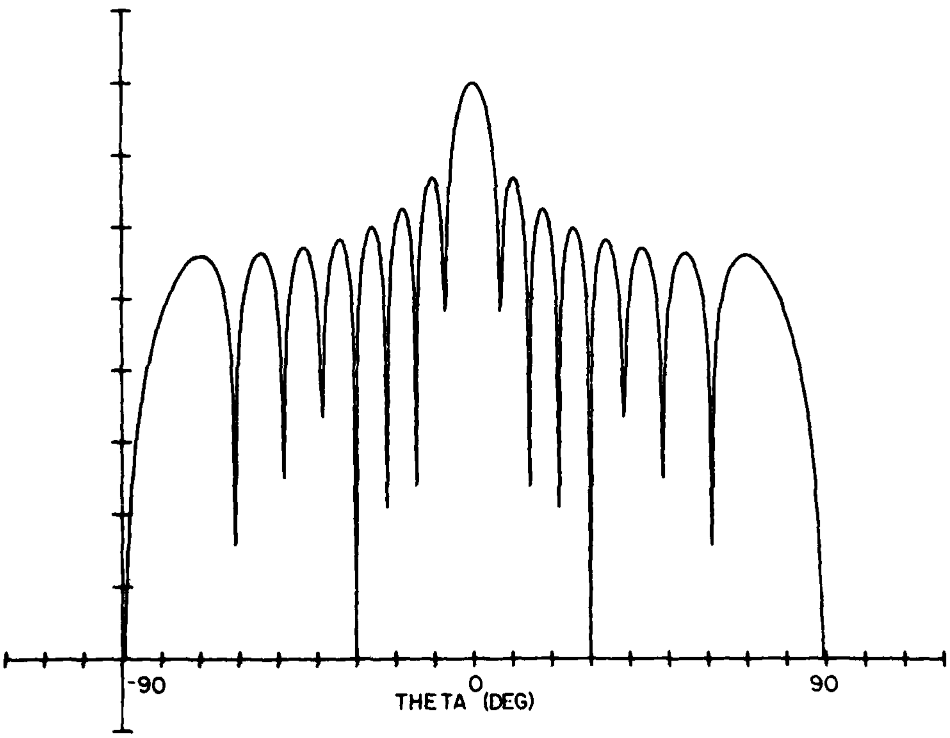


Figure 1b. Far-Field Pattern of 16-Element Linear Array

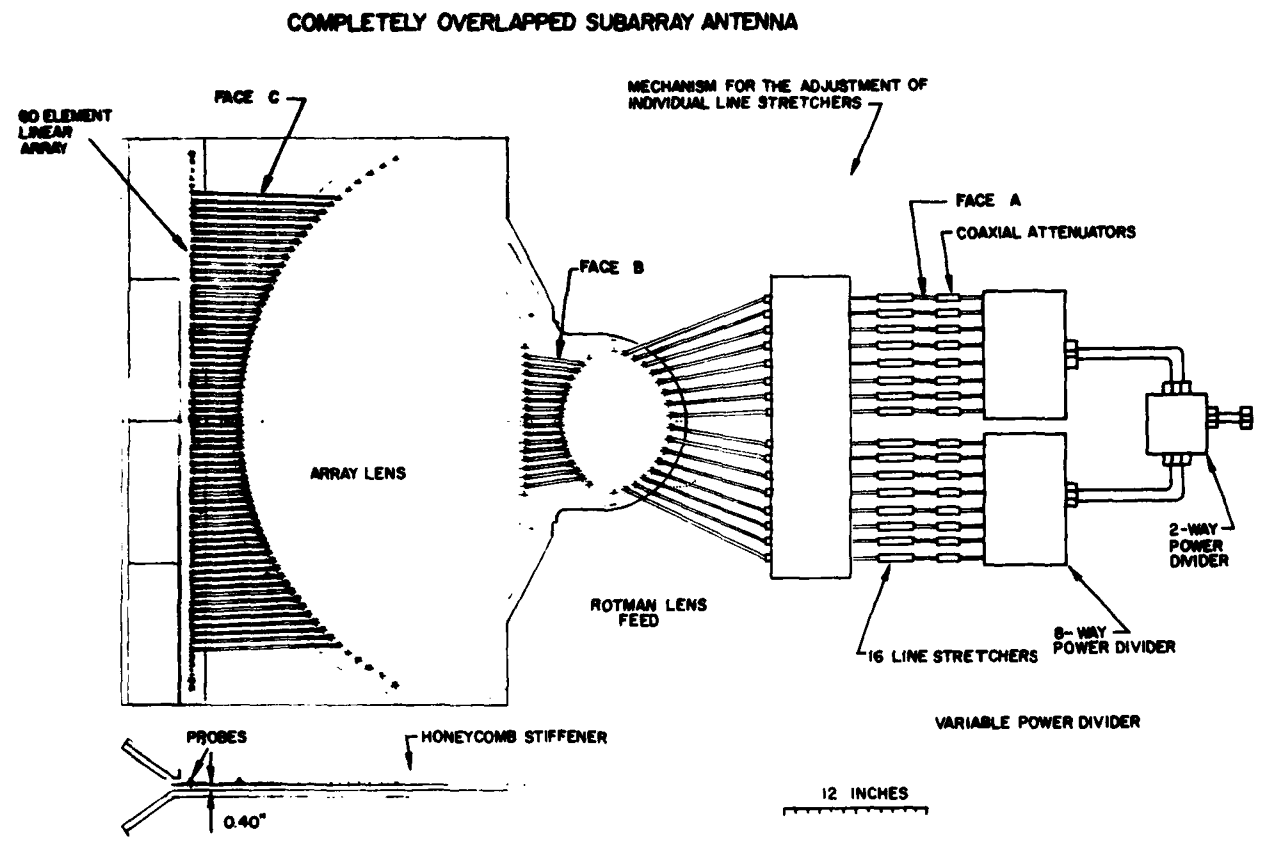


Figure 2a. Hardware Layout of the Experimental Completely Overlapped Subarray Antenna

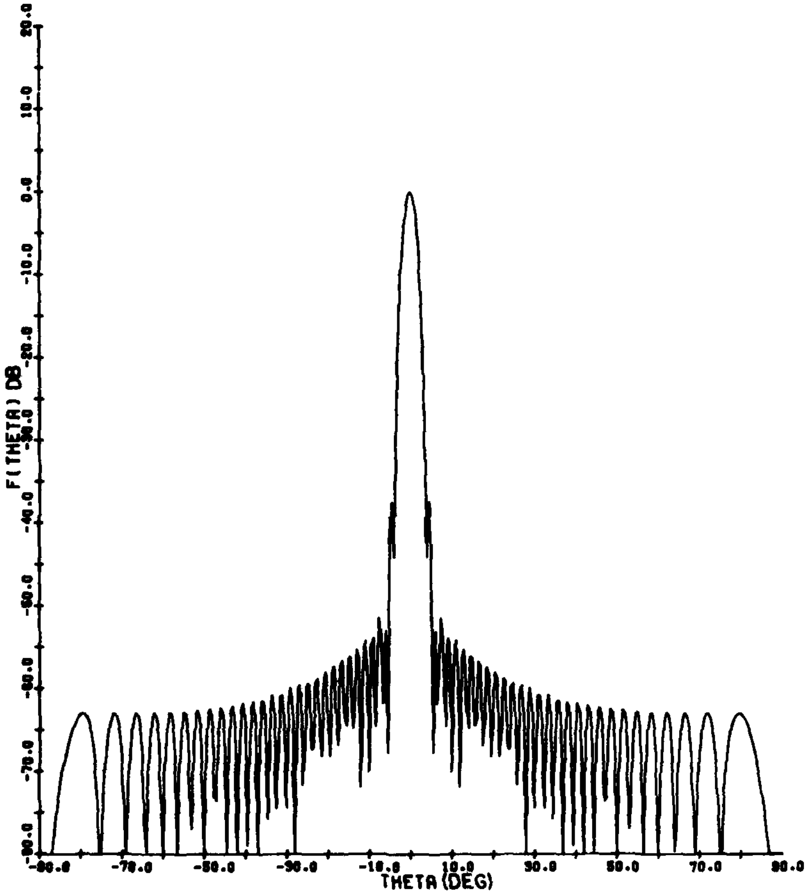


Figure 2b. Computer-Modeled Far-Field Pattern of the Completely Overlapped Subarray Antenna With Uniform Weighting. Patterns are cut off at -80.0 dB for plotting purposes

cost-effective pattern control. Thus, we must confine our control efforts to the face of the feed (Face B).

The only difference between this configuration and the previously described linear array is the addition of the main array lens. The contribution from each of the n feed elements is now no longer that of the isotropic radiator, but a $\frac{\sin x}{x}$ pattern steered to a specific angle by the action of the lens transform. An a priori knowledge of the phase and amplitude of the 16 element contributions at each angle will allow the determination of an appropriate weight for each element to achieve a desired pattern.

For example, if we examine only the contributions of the four feed elements in the center, the overall pattern value at an angle θ is given by:

$$F(\theta) = A_7C_7(\theta) + A_8C_8(\theta) + A_9C_9(\theta) + A_{10}C_{10}(\theta)$$

where,

A_n = The complex weight for the nth element

$C_n(\theta)$ = The complex pattern value of the nth element's contribution at the angle θ

For a given $F(\theta)$ the appropriate weight for one of the four elements can be written in terms of the other three element weights and the known $C_n(\theta)$, giving,

$$A_7 = \frac{F(\theta) - A_8C_8(\theta) - A_9C_9(\theta) - A_{10}C_{10}(\theta)}{C_7(\theta)}$$

This equation is simplified if A_8, A_9, and A_{10} are assumed to have unity weighting (constant amplitude, equal phase). The equation then reduces to,

$$A_7 = \frac{F(\theta) - C_8(\theta) - C_9(\theta) - C_{10}(\theta)}{C_7(\theta)}$$

The appropriate weights for the four elements are thus determined; the application of these weights to the four feed elements will result in the given pattern value at the angle θ. It should be noted that the choice of A_7 as the non-unity control variable was purely arbitrary. A similar calculation can be made for A_8, A_9, or A_{10}. Of course, the overall pattern will be different at angles other than θ, depending upon the chosen control weight.

The $F(\theta)$ value can be made to assume any value by the determination of appropriate element weights. In particular, if $F(\theta)$ is chosen to be zero, a null in the overall pattern will result. The pattern can be controlled and given many different shapes by appropriately weighting the feed elements.

Control at more than one angular position is achieved by solving a set of simultaneous equations of the form shown above. One equation is written for each desired control angle. Solution of the set by matrix algebra or any other valid method yields the appropriate element weights for the chosen control elements. For two-angle control an example set of equations would be:

$$F(\theta_1) = A_7 C_7(\theta_1) + A_8 C_8(\theta_1) + C_9(\theta_1) + C_{10}(\theta_1)$$

$$F(\theta_2) = A_7 C_7(\theta_2) + A_8 C_8(\theta_2) + C_9(\theta_2) + C_{10}(\theta_2)$$

Using elements 7 and 8 is again purely arbitrary. Any pair of elements can be used for control; however, it has been noted that the best overall pattern (in terms of rms sidelobe level) is achieved when the control weights are as close as possible to unity weighting.

2. COMPUTER-MODELED ADAPTED PATTERNS

The 16 feed elements on face B of the antenna allow pattern control at 15 discrete points in the pattern. Wide-band nulls can be achieved by clustering null angles close together. Figures 3a, 3b, and 3c show various positions of three pattern nulls. These computer-modeled patterns were generated by programming Eq. (1) for the far-field pattern of the Overlapped Subarray Antenna[2]

$$F(\theta, \theta_0) = \sum_{p=1}^{16} \sum_{n=1}^{16} I_p A_n \exp\left\{-J2\pi q_p (R-1) S_o - j2\pi \left(\frac{\Delta D}{\lambda F}\right) R q_n q_p\right\} \times \left[\frac{\sin 60(\pi h_n)}{\sin (\pi h_n)}\right] \qquad (1)$$

where,

$$h_n = R\left(\frac{\Delta\delta}{\lambda_o F}\right) q_n + \frac{\delta}{D}(RS - S_o)$$

θ = angle from boresight

θ_o = scan angle

2. Fante, R. L. (1979) Study of the Radiation Properties of Overlapped, Subarrayed Scanning Antennas, RADC-TR-79-293, AD A082401.

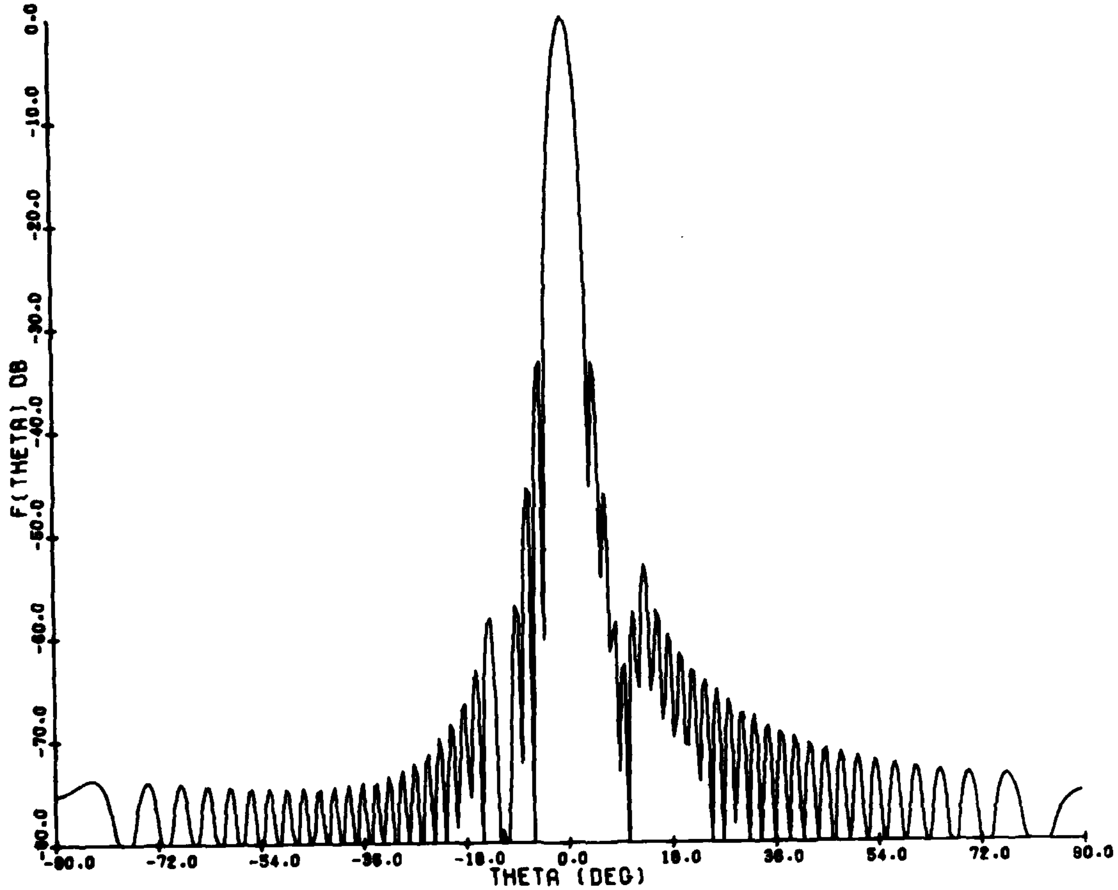


Figure 3a. Computer-Modeled Far-Field Pattern With Nulls Placed at $\theta_1 = -12.0^o$, $\theta_2 = -11.0^o$, and $\theta_3 = 36.0^o$

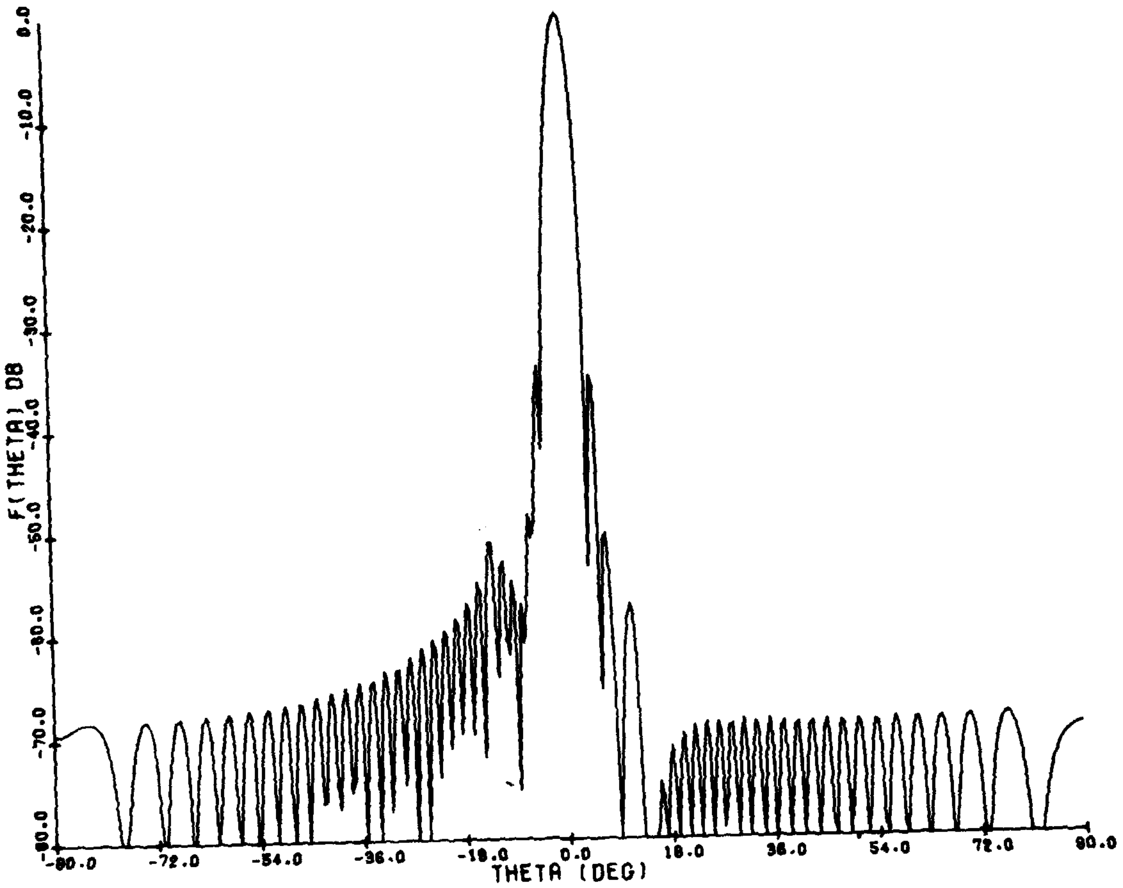


Figure 3b. Computer-Modeled Far-Field Pattern With Nulls Placed at $\theta_1 = 13.0^o$, $\theta_2 = 14.0^o$, and $\theta_3 = 15.0^o$

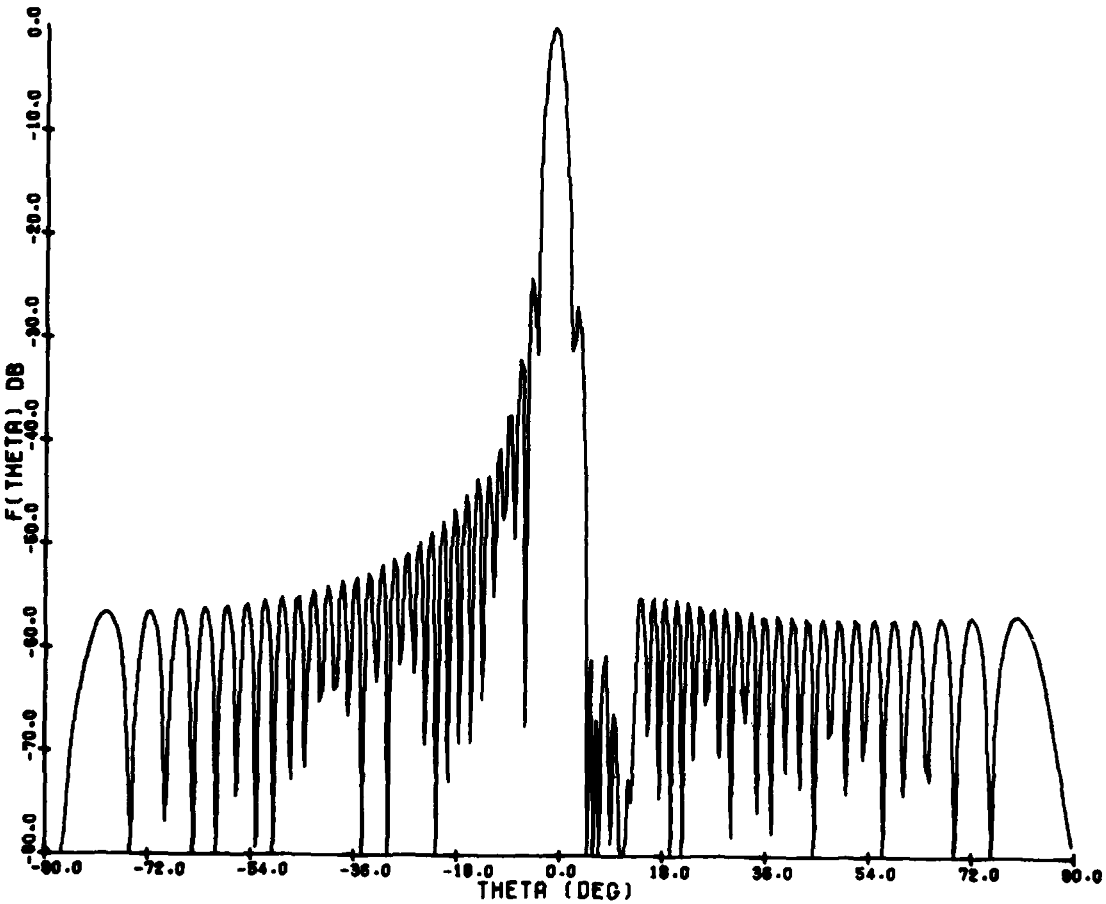


Figure 3c. Computer-Modeled Far-Field Pattern With Nulls Placed at $\theta_1 = 5.0^o$, $\theta_2 = 3.0^o$, and $\theta_3 = 7.0^o$. Note the higher near-in sidelobes caused by nulling close to the main beam

p = beam port number

n = feed element number

A_n = complex feed element weight

I_p = illumination amplitude taper

q_p = (p - 8.5)

q_n = (n - 8.5)

R = f/f_o

λ_o = wavelength at the center frequency, f_o

Δ = feed element spacing

δ = main array element spacing

D = subarray beam separation on the main array lens

S = $D/\lambda_o \sin\theta$

F = focal length of main array lens

S_o = $D/\lambda_o \sin\theta_o$

From Eq. (1), it can be seen that the far-field contribution from the nth feed element can be expressed as

$$F_n(\theta, \theta_0) = A_n \sum_{p=1}^{16} I_p \exp\left\{-j2\pi q_p (R-1) S_o - j2\pi \left(\frac{\Delta D}{\lambda_o F}\right) R q_n q_p\right\} \times \left[\frac{\sin 60 (\pi h_n)}{\sin (\pi h_n)}\right] \qquad (2)$$

The successful use of this method in practice rests upon the ability to accurately determine the individual element contributions. If random errors exist anywhere in the antenna, this accuracy is reduced. Also, as this method is inherently open-loop with no feedback of information to the control algorithm, these random phase and amplitude errors are not compensated for in the weight calculations. The method as a whole is highly sensitive to these errors and they tend to fill in any nulls achieved.

3. JAMMING SIGNAL DIRECTION FINDING

The practical application of this method to an actual jamming scenario requires the knowledge of the approximate angle of arrival of the jamming signals. These become the angles θ_n at which nulls are desired. This requirement can be satisfied by implementing the following locator in the form of an auxiliary array.

In the Applebaum adaptive algorithm[3] the information on the jammer angles is contained in the noise covariance matrix. There should be an operation which, when performed on the covariance matrix, will yield the correct jammer angles. For example, if we have two discrete jammers of arbitrary power and direction, the covariance matrix, M, for a three element array is given by,

$$M = \begin{bmatrix} P_1 + P_2 + P_q & P_1 e^{j\beta_1} + P_2 e^{j\beta_2} & P_1 e^{2j\beta_1} + P_2 e^{2j\beta_2} \\ P_1 e^{-j\beta_1} + P_2 e^{-j\beta_2} & P_1 + P_2 + P_q & P_1 e^{j\beta_1} + P_2 e^{j\beta_2} \\ P_1 e^{-2j\beta_1} + P_2 e^{-2j\beta_2} & P_1 e^{-j\beta_1} + P_2 e^{-j\beta_2} & P_1 + P_2 + P_q \end{bmatrix}$$

where P_1 and P_2 are the respective jammer powers and P_q is the quiescent noise power in each channel.

The Applebaum adaptive loop inverts this matrix and (if quiescent weighting) then sums the elements of each row to determine the appropriate weight vector W. The antenna pattern may be found by the dot product of this weight vector with a vector B of the form,

$$B = \begin{bmatrix} 1 & e^{j\beta} & e^{j2\beta} \end{bmatrix}$$

The result is a pattern with nulls at β_1 and β_2.

If instead of using the inverse of the covariance matrix in the above procedure, we first subtract the quiescent noise power from each of the diagonal elements and use the adjoint of this deleted covariance matrix, we find the following,

$$\text{Adj } M_{del} = P_1 P_2 \begin{bmatrix} (1) & (2) & (3) \\ (4) & (5) & (6) \\ (7) & (8) & (9) \end{bmatrix}$$

3. Applebaum, S.P. (1976) Adaptive arrays, IEEE Trans Antennas Propag. AP-24(No. 5):585-598.

where,

$$(1) = 2 - e^{j\beta_1} e^{-j\beta_2} - e^{-j\beta_1} e^{j\beta_2}$$

$$(2) = -e^{j\beta_1} - e^{j\beta_2} + e^{2j\beta_1} e^{-j\beta_2} + e^{-j\beta_1} e^{2j\beta_2}$$

$$(3) = 2e^{j\beta_1} e^{j\beta_2} - e^{2j\beta_1} - e^{2j\beta_2}$$

$$(4) = -e^{-j\beta_1} - e^{-j\beta_2} + e^{j\beta_1} e^{-2j\beta_2} + e^{-2j\beta_1} e^{j\beta_2}$$

$$(5) = 2 - e^{2j\beta_1} e^{-2j\beta_2} - e^{-2j\beta_1} e^{2j\beta_2}$$

$$(6) = -e^{j\beta_1} - e^{j\beta_2} + e^{2j\beta_1} e^{-j\beta_2} + e^{-j\beta_1} e^{2j\beta_2}$$

$$(7) = 2e^{-j\beta_1} e^{-j\beta_2} - e^{-2j\beta_1} - e^{-2j\beta_2}$$

$$(8) = -e^{-j\beta_1} - e^{-j\beta_2} + e^{j\beta_1} e^{-2j\beta_2} + e^{-2j\beta_1} e^{j\beta_2}$$

$$(9) = 2 - e^{j\beta_1} e^{-j\beta_2} - e^{-j\beta_1} e^{j\beta_2}$$

The weight vector is found by summing the elements of each row in this adjoint. When the dot product of this weight vector and the vector B is taken, we find that when β equals β_1 or β_2 the pattern value is near zero. When β is any other value the pattern value is relatively constant (see Figures 4a, b, c). Therefore the discontinuities in the resultant pattern indicate the directions of arrival for the two jamming signals. This has been shown to be true only when the number of jammers is one less than the number of elements in the locating array.

The implementation of the locating array is fairly simple. A linear array with a number of elements which is greater than the suspected number of jammers is mounted in the same orientation as the main array lens. A tap loop is established behind each element to determine the noise covariance matrix. An assumption is made as to the quiescent noise power level and the deleted covariance matrix is formed. The adjoint of this covariance matrix is then determined. If the severe discontinuities do not occur, then the tap loop from an end element is electronically deleted from the array and the computations are performed again. This process is continued until the discontinuities occur and the jammers may now be nulled from the pattern of the Overlapped Subarray Antenna.

A two-dimensional Overlapped Subarray Antenna might be made more compact and more maneuverable by using one of the horizontal rows of main array elements as the locating array.

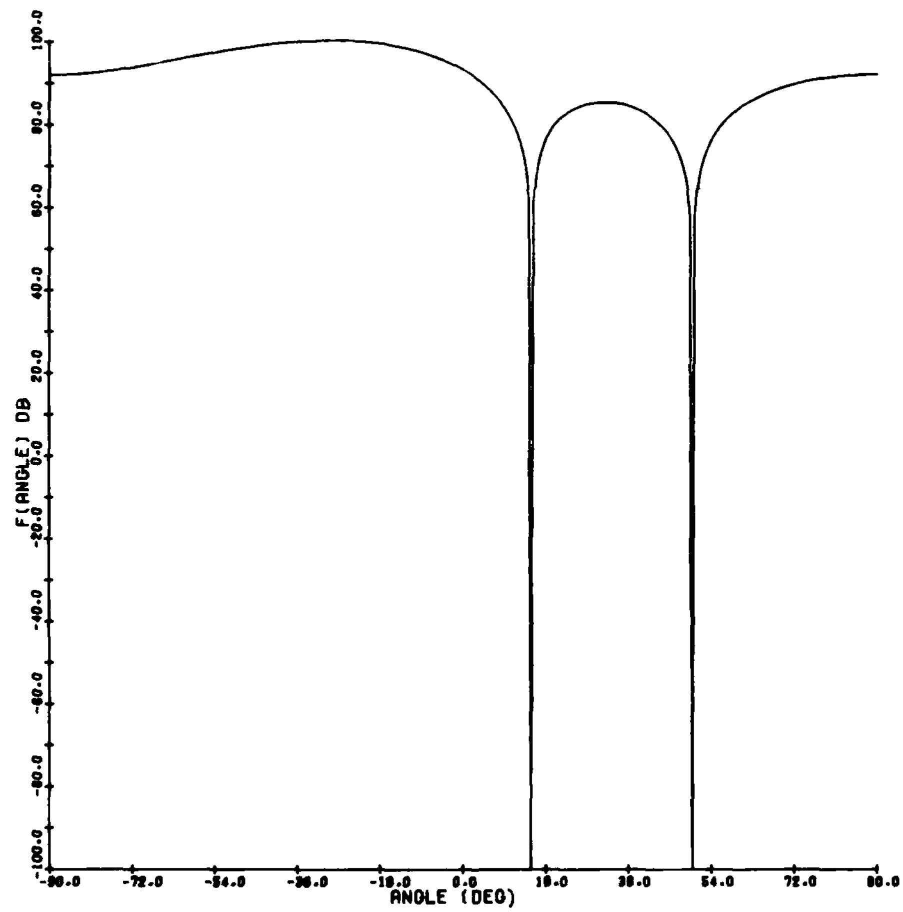


Figure 4a. Pattern From Locating Array Indicating Two Discrete Jamming Signals Where $\theta_1 = 15.0^o$, $\theta_2 = 50.0^o$, $P_1 = 100$ W, $P_2 = 100$ W, and $P_q = 1.0$ W

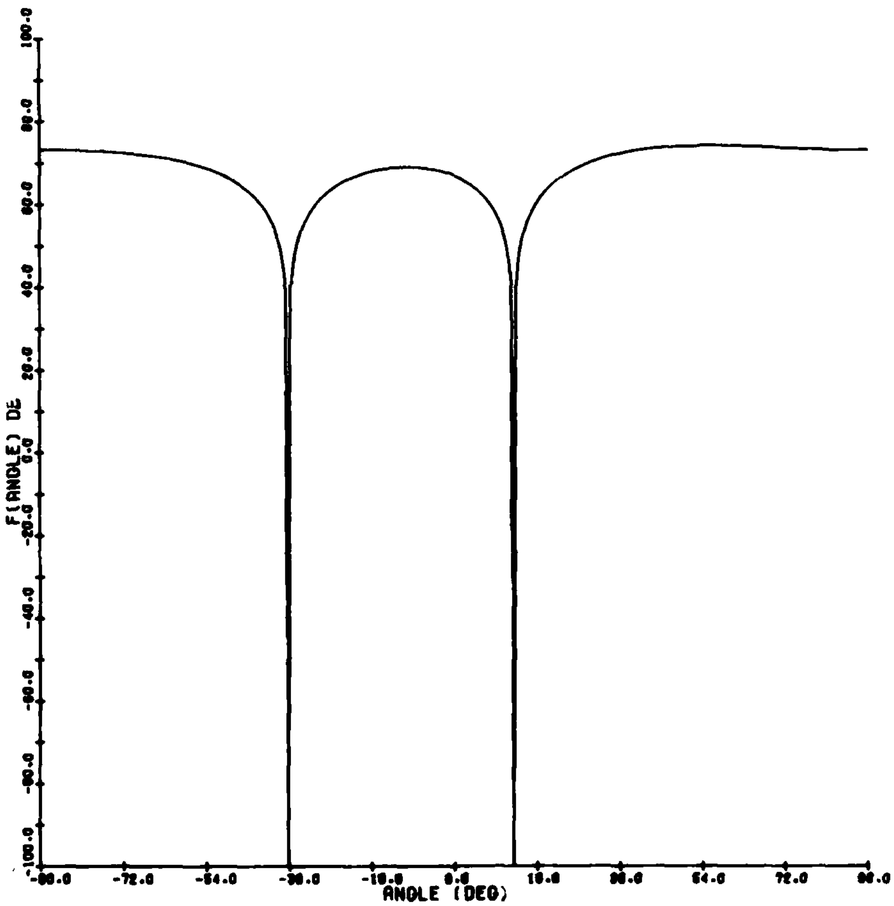


Figure 4b. Pattern From Locating Array Indicating Two Discrete Jamming Signals Where $\theta_1 = 36.0^\circ$, $\theta_2 = 13.0^\circ$, $P_1 = 1000$ W, $P_2 = 0.5$ W, and $P_q = 1.0$ W

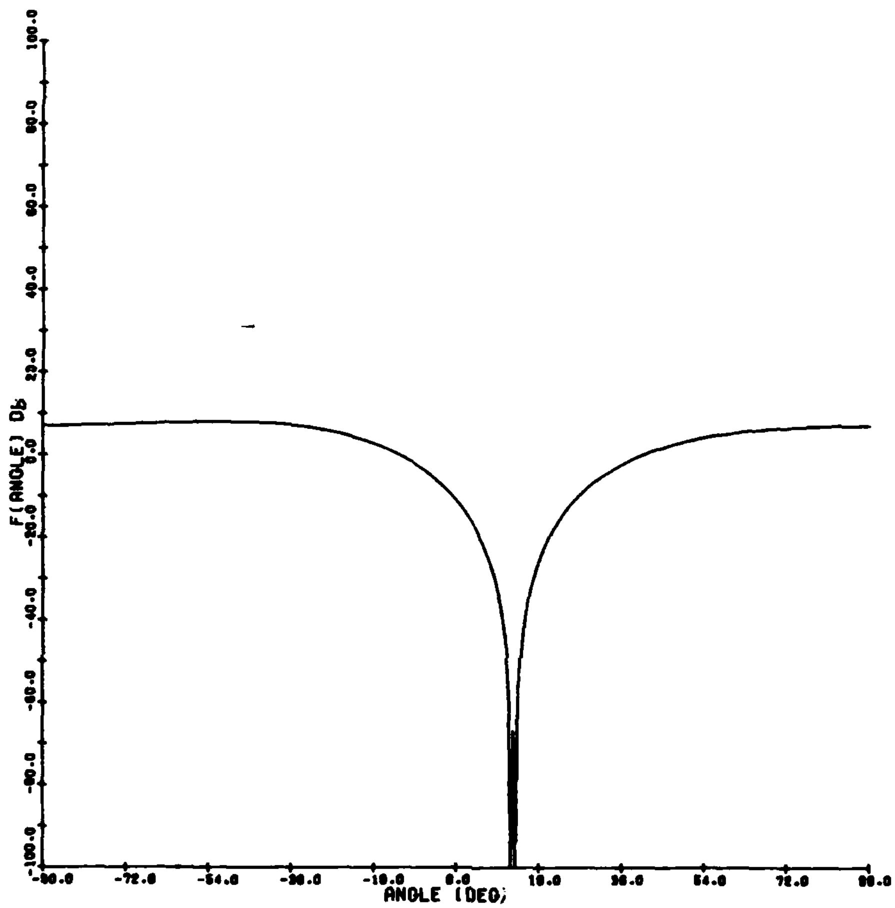


Figure 4c. Pattern From Locating Array Indicating Two Closely Spaced Jamming Signals Where $\theta_1 = 12.0^o$, $\theta_2 = 13.0^o$, $P_1 = 1000$ W, $P_2 = 0.5$ W, and $P_q = 1.0$ W

4. CONCLUSIONS

The procedure described above is effective in reducing susceptibility to jamming signals. The processing can be entirely electronically controlled. If the system errors can be kept to a minimum the performance will not be greatly degraded.

3