AR-009-743

DSTO-TR-0361

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Adaptive Array Processing – Near Field Experiment

Dan Madurasinghe and Laysuan Teng

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Microwave Radar Division
Electronics and Surveillance Research Laboratory

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ABSTRACT

Adaptive array processing in airborne radars to overcome jamming is critical to obtaining satisfactory performance in a hostile environment. The usual application of a plane wave solution to this situation is inadequate when there is nearfield scattering. The proposed algorithm introduced in this paper is effective for wavefronts originating from all ranges, even though emphasis will be placed on the spherical wavefronts produced in the near field regions. A single snapshot based high-resolution adaptive algorithm employing the Least Square Technique is considered in two stages. First, the directions of arrival of all spherical or plane wavefronts impinging on a linear sensor array is estimated. Next, nulls are formed in the estimated directions of the interferers while receiving the desired signal. This optimal solution is undeniably computationally intensive. However, to be able to resolve near field specular reflections/jamming accurately is a breakthrough for all concerned. The effectiveness of this algorithm is demonstrated both experimentally and by simulation.

RELEASE LIMITATION

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Published by

DSTO Electronics and Surveillance Research Laboratory
PO Box 1500
Salisbury, South Australia 5108 Australia

Telephone: (08) 259 5555
Fax: (08) 259 5200

AR-009-743
July 1996

APPROVED FOR PUBLIC RELEASE

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EXECUTIVE SUMMARY

In this study adaptive array processing for a radar system refers to filtering algorithms that suppress interference/jamming signals while maximising the signal arriving from a desired direction. As a consequence, the probability of target detection in combined clutter and jamming environments is improved, and this is where the future of airborne radars lies.

However, the wavefronts arriving at the array have conventionally been assumed to be planar and this means they are assumed to be all far field sources. This is not generally true because specular reflections from the aircraft body itself are near field sources and thus the reflected wavefronts impinging on the array are non-planar. For this reason, a near field solution based on Least Squares is proposed in this paper. It is important to note that the near field solution can be used in far field problems but not vice-versa.

The proposed algorithm is verified using both simulated and measured data. In both cases, the model consists of two jammers and an L-Band, 15 element linear, equispaced array with half wavelength spacing between elements. The target is placed at 0° broadside at a distance of 17 m from the receiving array; i.e. in the near field region. The use of the 15-element array means the array has 14 degrees of freedom which in turn means it can resolve up to 14 interfering signals. One jammer is placed on each side of the target.

The results generated from both simulated and experimental data show the Least Squares near field solution to be more powerful than the conventional far field solution. However, all good things come at a cost and in this case it is at the expense of a greatly increased computational load.

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Dan Madurasinghe received the B.Sc(Physics) and B.Sc.Hon(Mathematics) degrees with first class honours from the University of Sri Lanka in 1978 and 1980 respectively, the M.Sc(Computer Science) degree from the Asian Institute of Technology(Bangkok) in 1982, the Ph.D (Applied Mathematics), the Grad. Cert. (Telecommunications) from the Adelaide University in 1986 and 1990 respectively. He joined DSTO's Communications Division as a Research Scientist in 1988 and later joined the Microwave Radar Division as a Senior Research Scientist. His current research interests cover Space Time Adaptive Processing, Non-Acoustic Detection and Numerical Techniques.

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Contents

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1. Introduction

In airborne radar applications, far field returns such as ground reflections degrade the quality of the recovered signal. As a consequence numerous clutter/noise suppression techniques have evolved. Near field specular reflections from the body of the aircraft themselves generate interference field that call for special attention. Application of the conventional far field plane wave solution to this situation is inadequate.

The objective of adaptive array processing is to compute a set of weights to optimally reconstruct the desired signal from a known direction of arrival when coherently summed[1]. These weights are adjusted from one snapshot to another to accommodate the changing environment to effectively null the jamming signals and atmospheric noise. This array processing method forms the basis of beamforming techniques such as the conventional direction finding technique where the directions of all arrivals are first estimated.

It is important for us to pay attention to the basic Direction of Arrival (DOA) estimation techniques and the associated complicating factors discussed in the literature. The presence of coherent signals, the estimation of direction of arrival of interference signals imposes great difficulties when high resolution methods, such as Eigendecomposition-based techniques, are used. A spatial smoothing [2] scheme is one of the preprocessing techniques recently proposed to circumvent these hurdles. However, the scheme reduces the effective degrees of freedom by half. Another optimal technique which operates in a fully coherent environment is a Maximum Likelihood based processor [3]. Full degrees of freedom are retained in resolving DOAs using this method. One dominant feature of most conventional beamforming/direction-finding methods is that the data model is assumed to be stationary and ergodic. The performance of the resulting adaptive algorithms, which in most cases minimises the mean output error of many successive snapshots, tends to display a transient behaviour. In some cases, they fail in the presence of fully coherent interferers. For this reason, in this study, we formulate a single snapshot based least squares estimator to construct a look direction constraint coherent interference nulling algorithm using a suitable single snapshot based DOA estimator.

The proposed null steering algorithm consists of two stages. First, the directions of arrival (DOA) of all the wavefronts, be it plane or spherical, impinging on the array are estimated using Least Squares Estimation. Having estimated the DOAs, a set of complex weights is computed to null the estimated jamming directions while producing a pole in the direction of arrival of the desired signal. The beamforming output is simply the coherent integration of the products of these weights with the measured data at each element. Algorithms such as these have been investigated in the past [1] based on the far field assumption that all arriving wavefronts are planar. In this paper, this problem is resolved at the array data model by assuming that the arriving wavefronts are a combination of both plane and spherical wavefronts.

To examine the validity of the theory, experimental data measured in the near field is processed using the proposed technique. The experimental model consists of an L-Band, 15 element linear, equispaced array with half wavelength spacing between elements at a distance 17 m from the target. The measured data shows the spherical nature of the wavefront detected at the receivers, as depicted in Figure 1.

Section 2 presents the formulation of the proposed algorithm in two steps. The performance of this algorithm is compared with the far field solution from a single step look direction constrained algorithm based on a single snapshot in Section 3. Section 4 shows the analysis of the experimental results using these two algorithms. Section 5 demonstrates the simulation results.

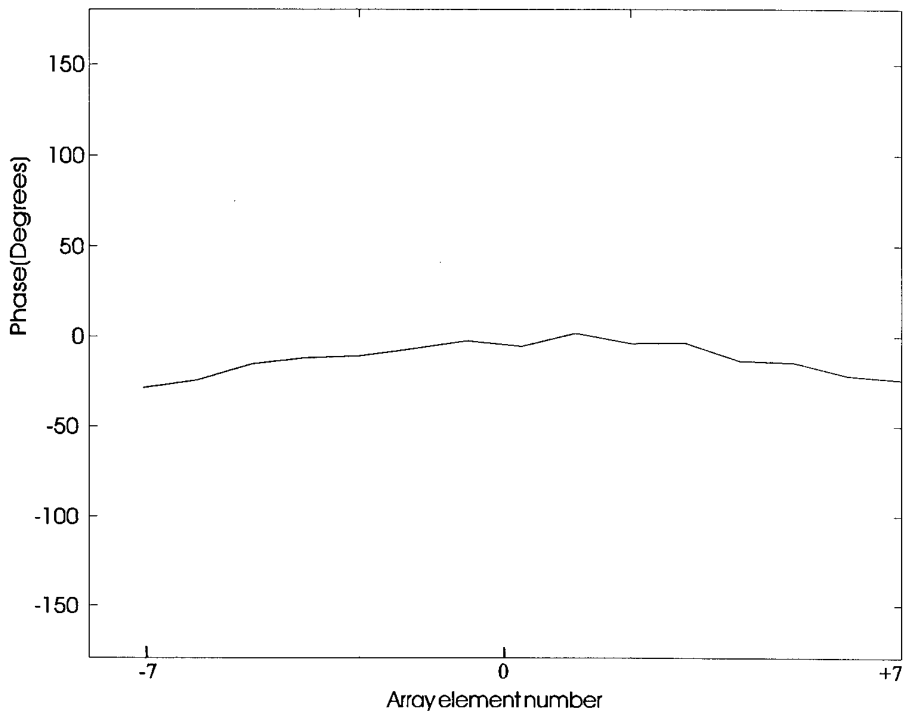

Figure 1: Measured phase variation across a 15 linear equispaced, L-Band (at 2GHz) array with half wavelength spacing for a signal arriving at broadsidet $\theta = 0°$. The centre element(8) is phase matched to the source at a distance 17 m.

2. Problem Formulation

Consider a linear array of $2N+1$ elements as shown in Figure 2. Let y be the distance to the nth element measured from the center element $(n = 0, \pm 1, \pm 2, \ldots, \pm N)$. Assume a spherical wavefront impinging on the array where θ is the direction of arrival measured from the array broadside, and r is the radius of curvature of the wavefront.

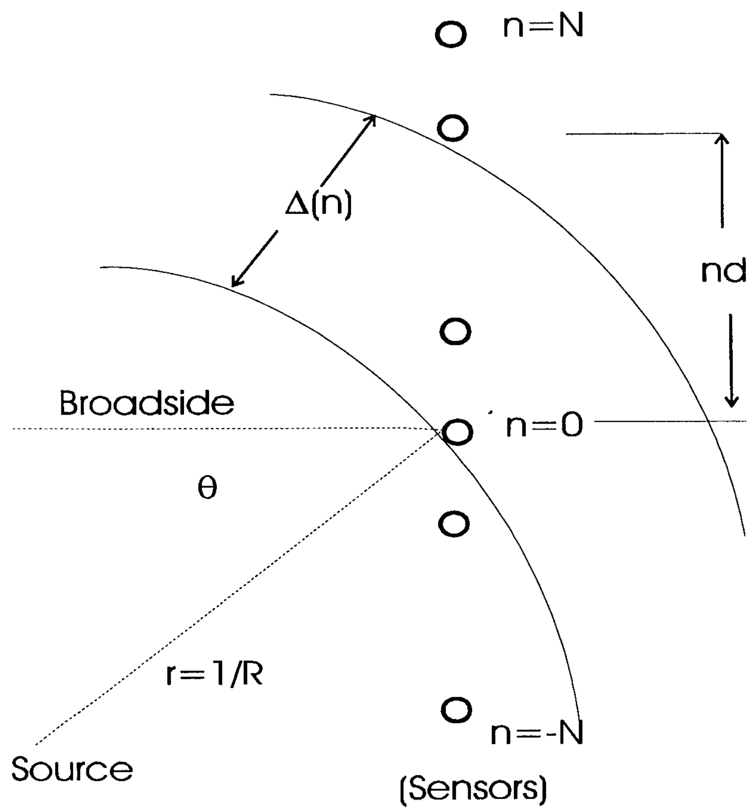

Figure 2: Spherical wavefront model with the source signal arriving broadside to the array. d is the interelement spacing and $\Delta(n)$ *is the phase lag across the array at element position* n.

The path delay at the nth element relative to the center element is given by

$$\Delta(n) = \sqrt{(r\cos\theta)^2 + (r\sin\theta + nd)^2} - r \tag{1}$$

where d is the element spacing.

By substituting $r = 1/R$ and simplifying Eq(1) using Taylor's expansion, we may write

$$\Delta(n) = nd\sin\theta + \frac{1}{2}Rn^2d^2 - \frac{R}{8}\left(2nd\sin\theta + Rn^2d^2\right)^2 \tag{2}$$

where $\frac{1}{2}Rn^2d^2$ is the first order correction, and

$\frac{R}{8}\left(2nd\sin\theta + Rn^2d^2\right)^2$ is the second order correction in the Taylor expansion.

Now we may write the $[(2N+1)\times 1]$ steering column called $\mathbf{S}(\theta, R)$, where

$$S(\theta,R) = \left(e^{-j2\pi\Delta(N)/\lambda}, e^{-j2\pi\Delta(N-1)/\lambda}, \ldots, e^{-j2\pi\Delta(1)/\lambda}, 1, e^{-j2\pi\Delta(1)/\lambda}, \ldots, e^{-j2\pi\Delta(N-1)/\lambda}, e^{-j2\pi\Delta(N)/\lambda}\right)^T, \tag{3}$$

$R = 0$ corresponds to incoming wavefronts in the far field which are plane,
T denotes the vector transpose, and
λ is the wavelength of the transmit signal.

Let $A_1(t), A_2(t), \ldots, A_M(t)$ be the magnitudes of M wavefronts impinging on the array at time t. Then the measured signal, $\mathbf{X}(t)_{[(2N+1)\times 1]}$, with narrowband carrier removed, is given by

$$\mathbf{X}(t) = \mathbf{B}\mathbf{A}(t) + \mathbf{E}(t) \tag{4}$$

where

$\mathbf{X}(t) = (x_1(t), x_2(t), \ldots x_{2N+1}(t))^T$,
$\mathbf{A}(t) = (A_1(t), A_2(t), \ldots, A_M(t))^T$,
$\mathbf{B} = (\mathbf{S}(\theta_1, R_1), \mathbf{S}(\theta_2, R_2), \ldots, \mathbf{S}(\theta_M, R_M))$, and
$\mathbf{E}(t)$ represents a random noise vector.

Our objective is to select a set of complex weights such that the desired signal, $A_1(t)$, arriving from a known direction θ_1 can be estimated as

$$\hat{\mathbf{A}}_1(t) = \mathbf{W}^T\mathbf{X}(t) \tag{5}$$

where $\hat{A}_1(t)$ is the estimate of $A_1(t)$.

For zero noise case (ie $\mathbf{E} = 0$), the measured signal $\mathbf{X}(t)$ is a linear combination of the columns of $\mathbf{B}$. However, noise causes it to wander away from this space. This error is given by

$$\mathbf{E}(t) = \mathbf{X}(t) - \mathbf{PX}(t) \tag{6}$$

where $\mathbf{P} = \mathbf{B}(\mathbf{B}^H\mathbf{B})^{-1}\mathbf{B}^H$ is the projection operator on the space spanned by columns of $\mathbf{B}$ and H denotes the Hermittian transpose[3].

We can minimise the error vector from this space to the measured vector $\mathbf{X}(t)$ to obtain an optimal set of values for $\theta_1,\theta_2,\ldots,\theta_M, R_1, R_2,\ldots, R_M$ at each snapshot by defining the following multidimensional minimization problem.

$$\min_{\Theta_R} \|\mathbf{E}\|^2 = \|\mathbf{X}(t) - \mathbf{PX}(t)\|^2 \tag{7}$$

where $\Theta_R = (\theta_1,\theta_2,\ldots,\theta_M, R_1, R_2,\ldots, R_M)$.

Optimisation techniques suitable for this type of minimisation problems have been discussed previously [3] for wavefronts originating in the far field using the method of alternating projection. Once the jammer directions and locations are estimated by minimising the total sum of squares of errors as defined in Eq(7), we have the best estimate for $\mathbf{A}(t)$ as

$$\hat{\mathbf{A}}(t) = \mathbf{VX}(t) \tag{8}$$

where $\mathbf{V} = (\mathbf{B}^H\mathbf{B})^{-1}\mathbf{B}^H$.

The estimate for $A_1(t)$ is given by Eq(5) where $\mathbf{W} = (v_{11}, v_{12}, \ldots, v_{1,2N+1})^T$ is the first row of $\mathbf{V}$.

3. Linear Beamforming

For plane wavefronts impinging on the array, we may write the steering vector, $\mathbf{S}_L(\theta)$, as follows :

$$\mathbf{S}_L(\theta) = (1, u_i, u_i^2, \ldots, u_i^{2N})^T \tag{9}$$

where $u_i = e^{-j2\pi d \sin\theta_i/\lambda}, (i = 1,2,\ldots,M)$ denotes the angles of arrival of coherent or incoherent plane wavefronts. The measured signal at the array is given by

$$\mathbf{X}(t) = A_1(t)\mathbf{S}_L(\theta_1) + \sum_{j=2}^{M} A_j(t)\mathbf{S}_L(\theta_j) + \mathbf{E} \tag{10}$$

where $A_1(t)$ is the desired signal and $A_j(t), (j = 2, \ldots, M)$ represents jamming signals.

In a single step look direction constraint architecture, we would like to find a set of complex weights $\mathbf{W} = (w_1, w_2, \ldots, w_{2N+1})$ to satisfy the following set of equations.

$$\mathbf{W}^T\mathbf{X}(t) = A_1(t)\mathbf{W}^T\mathbf{S}_L(\theta_1) + \sum_{j=2}^{M} A_j(t)\mathbf{W}^T\mathbf{S}_L(\theta_j) \tag{11}$$

where $\mathbf{W}^T\mathbf{S}_L(\theta_1) = 1$, and

$$\mathbf{W}^T\mathbf{S}_L(\theta_j) = 0, \text{ for } j = 2,3,\ldots,M.$$

In the absence of random noise for $M \le N + 1$, this problem has an exact solution [4] with $w_{M+1} = w_{M+2} = \ \ldots\ = w_N = 0$ and $(w_1, w_2, \ldots, w_m)^T = \mathbf{W}^{(1)}$ (say) given by $\mathbf{Q}\mathbf{W}^{(1)} = \mathbf{Y}$, where $\mathbf{Y}_{m\times 1} = (1, 0, 0 \ldots 0)^T$ and $\mathbf{Q}$ is a $M \times M$ matrix given by

$$
Q = \begin{bmatrix} 1 & u_1 & \cdots & u_1^{m-1} \\ x_1 - x_2/u_1 & x_2 - x_3/u_1 & \cdots & x_m - x_{m+1}/u_1 \\ x_2 - x_3/u_1 & x_3 - x_4/u_1 & \cdots & x_{m+1} - x_{m+2}/u_1 \\ \cdots & \cdots & \cdots & \\ x_{m-1} - x_m/u_1 & x_m - x_{m+1}/u_1 & \cdots & x_{2m-2} - x_{2m-1}/u_1 \end{bmatrix}_{[MxM]} \tag{12}
$$

For comparison we use this single snapshot based linear solution which is valid regardless of the source coherency.

4. Experiment

A simple, low cost experiment was carried out with the aim of reconstructing a CW signal in the presence of two coherent jammers using the above algorithms. The diagram of the setup is illustrated in Figure 3.

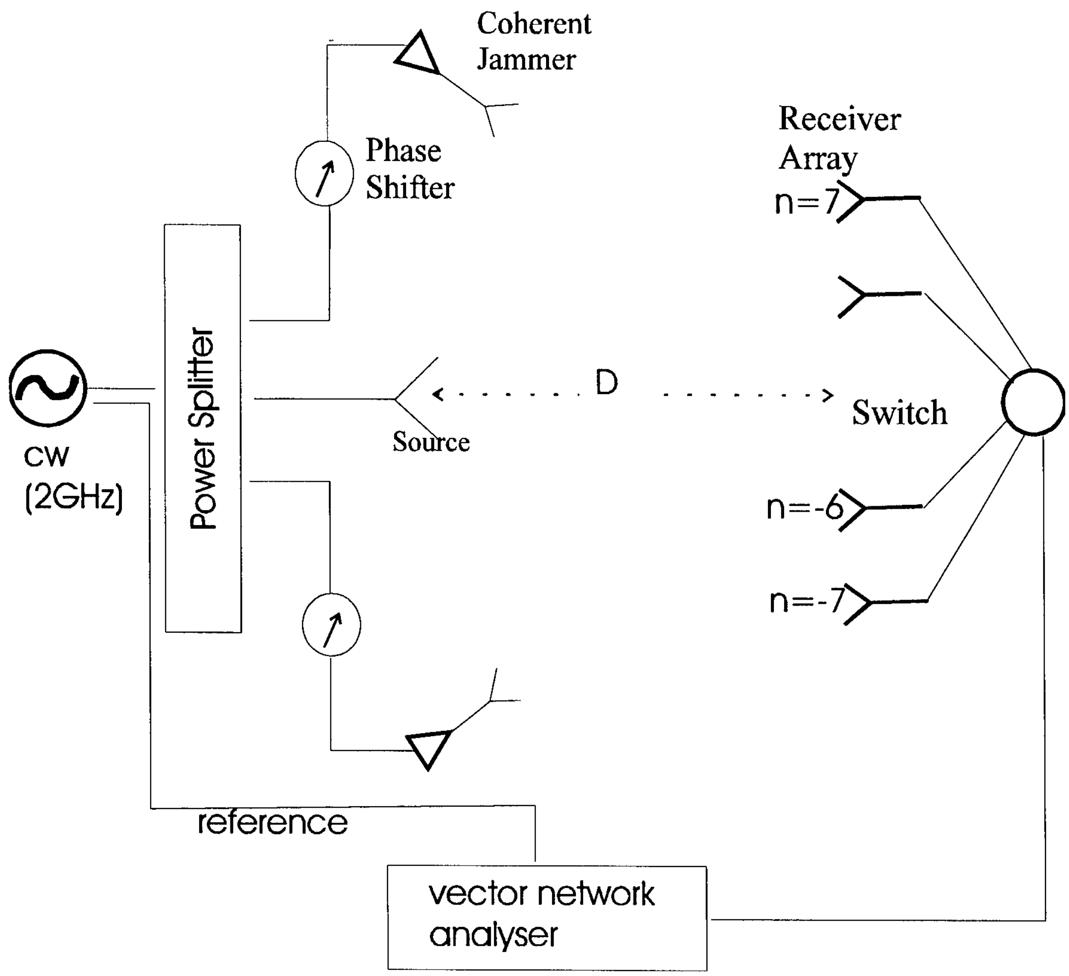

Figure 3: Setup of the adaptive interference nulling experiment with 15 receiver elements in the near field.

A two-port network analyser was used to measure the magnitude and phase of the received signal at discrete time interval (also refereed to as snapshots) to eliminate the need for an A/D converter. The setup consists of a 15 element linear equispaced receiving array with half wavelength spacing between elements operating at $2\,GHz$. The source, in this case, acts as the target at a distance 17m from the receivers at array broadside $(\theta = 0°)$. One jammer sits on each side of the source subtending equal angle to the center array element. The transmit signal is a 2GHz Continuos Wave(CW) signal and we record its phase versus voltage instead of time verses voltages. The source and the two jammers are synchronised using phase shifters to simulate a coherent jamming environment. For convenience, the transmit power from the two jammers were set to the same value.

The measured signal vector at the array in the presence of two coherent jammers, $\mathbf{X}(t)$, takes the following form :

$$\mathbf{X}(t) = (\hat{v}_{-7}e^{j\hat{\phi}_{-7}}, \hat{v}_{-6}e^{j\hat{\phi}_{-6}}, \ldots, \hat{v}_{7}e^{j\hat{\phi}_{7}})^T \tag{13}$$

where $\hat{v}_j$ is the magnitude of the received signal, and $\hat{\phi}_j$ is the phase of the received signal for each snapshot at sample nmber, t, for $j = -7,-6,\ldots,7$.

Without the jamming signals, the calibrated received signal at the center element gives the reference signal with which the reconstructed signal can be compared.

4.1. Results

Figure 4 illustrates the reconstructed signal in the presence of two coherent jammers at 10dB above the desired signal level (SNR of -10dB) impinging on the array at ±45° to the array broadside. The linear beamforming technique (as described in Section III) or the single step adaptive look direction constrained processor performs very poorly, but the two-step least squares approach performs exceptionally well. We have applied the least squares algorithm under two conditions, 1) far field and 2) near field. The plane wavefronts in the far field regions are enforced by setting $R_1 = R_2 = R_3 = 0$ in Eq(7) while the spherical wavefronts in the nearfield are estimated as given in Eqs(2)&(7). The number of degrees of freedom is assumed to be 7 (ie = M) for the cases presented in this paper. Three received signals at the array introduces six parameters in the optimisation procedure (three angles and radii of curvatures) and this demands at least six degrees of freedom. We have chosen seven degrees of freedom in orders to nullify any numerical errors.

Figure 5 represents the experimental results where the two jamming signals subtend the source at ±20° to broadside. Near mainlobe jamming where the DOAs of the jammers were placed at ±5° to broadside, was also performed and the results is illustrated in Figure 6. In this case, the analysis performed based on the far field assumption fails while the near field analysis recovers the transmit signal.

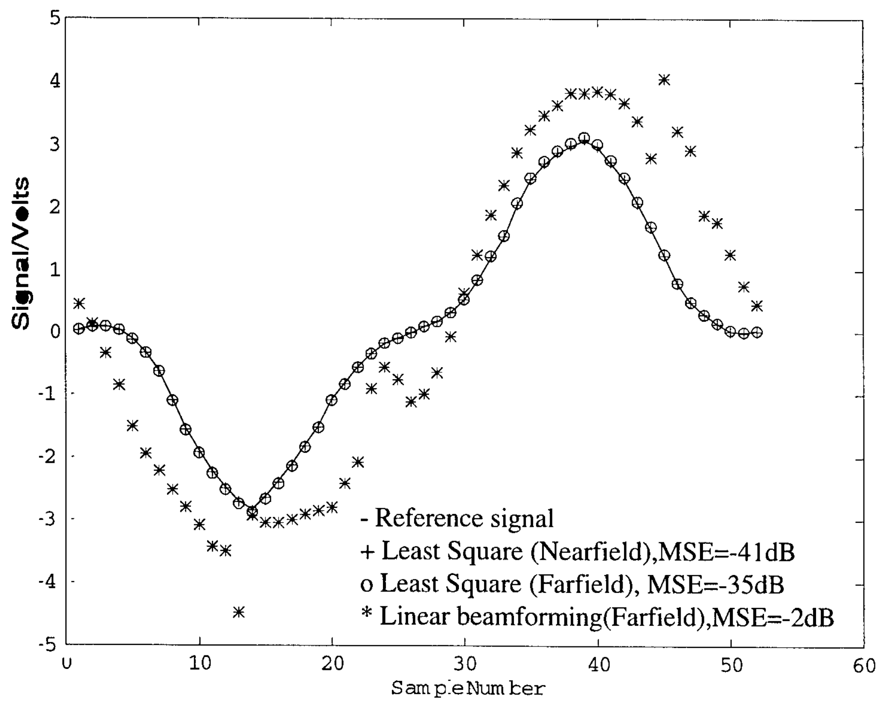

Figure 4: Reconstructed signal in the presence of two jammers(DOA= ±45^0, jammer/sig=10dB, reference signal DOA=0 degrees). MSE = mean squared output error/mean squared signal power.

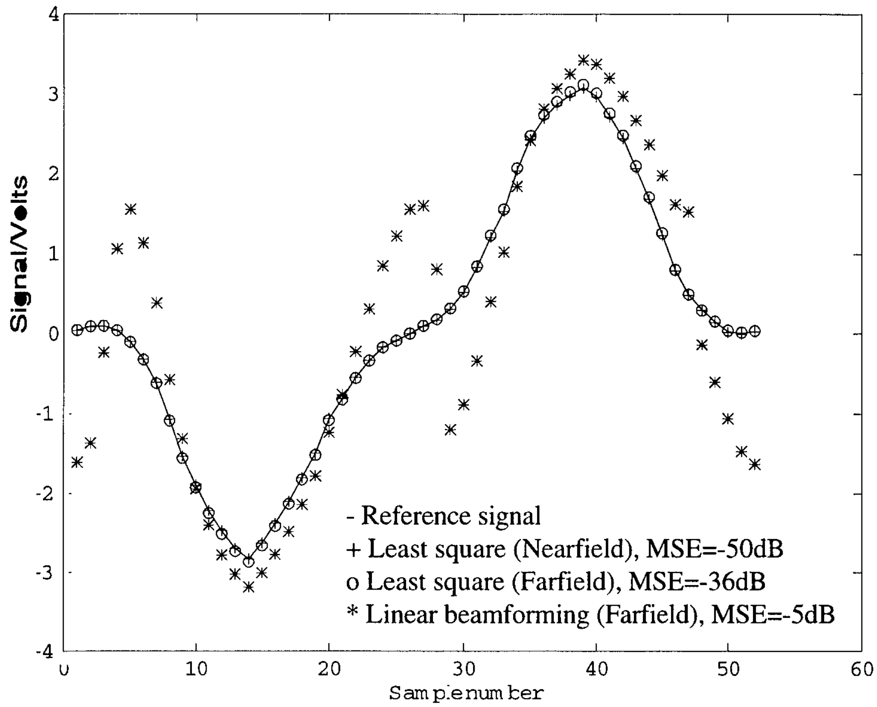

Figure 5: Reconstructed signal in the presence of two jammers(DOA=±20°, jammer/sig=20dB, reference signal DOA=0 degrees). MSE = mean squared output error/mean squared signal power.

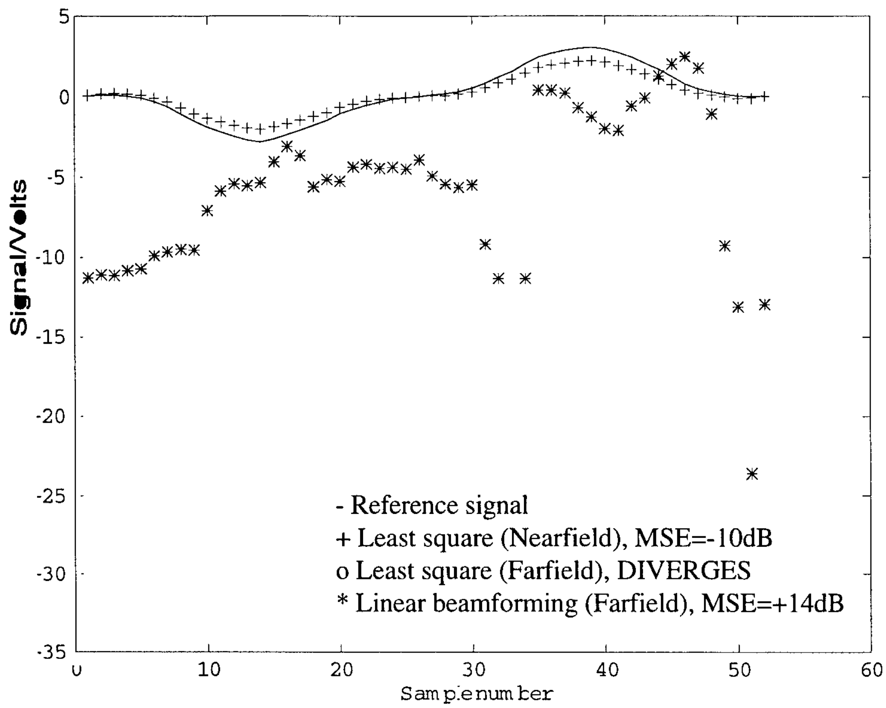

Figure 6: Reconstructed signal in the presence of two jammers(DOA= $\pm 5^0$*, jammer/sig=0dB, reference signal DOA=0 deg). MSE = mean squared output error/mean squared signal power.*

5. Simulation

To investigate the performance of the near field formulation based on the least squares approach, we have simulated a series of data sets using Eq(1) which assumes perfect spherical wavefronts at the array. The target is located the same distance away from the receiving array as the two jammers, as described in the Section 4. The parameters $R_1, R_2 \& R_3$ are the variables defined in terms of wavelength ie $R_1 = R_2 = R_3 = K\lambda$ where K is some integer.

Again, the signal source acts as the target positioned at 0° broadside to the array and the two jammers are placed at ±20° broadside with a signal to jamming ratio of -$20dB$. The jamming signals are simulated by generating a set of numbers at random and hence representing a constantly changing jamming environment but at fixed angles. In the absence of any system random noise (ie $E = 0$), as $K \to \infty$, it is expected that all three approaches discussed in Sections 2 and 3 will converge to a common error free solution. However, for small values of K this anticipated behaviour does not hold. In this case, the antenna pattern is determined by strong near field scatterers as illustrated in Figure 7.

The near field solution is represented in two cases :

1) with first order correction, and
2) with second order correction

as given in Eq(2).

The third solution presented is that for the far field by setting $R = 0$ where $R = 1/r$. The directions and distances of arrival of the jammers are the unknowns to be estimated in the minimisation procedure given in Section 2. The fourth being the far field solution based on the linear beamforming technique.

Figure 7 depicts the results generated from the above mentioned scenario. As predicted, near field solutions are better suited to describe the spherical wavefronts originating in the near field regions. The second order correction solution presents a more accurate result than that of the first order, as expected. For distances of arrival greater than 30λ, the mean square errors are below -30dB for both cases and hence the first order solution should suffice in order to save on computing cost. As for far field solutions, the linear technique surprisingly outperformed the least square estimation technique. The lack of injected channel noise (bearing in mind $E = 0$) could be one reason behind this unexpected behaviour. However, the comparison between far field solutions will not be pursued here. Average computational requirement to perform linear beamforming technique is about 4000 flops=4Kflops(floating points operations), but least squares method demands around 1000Kflops, and least squares with nearfield correction demands around 2000Kflops per run.

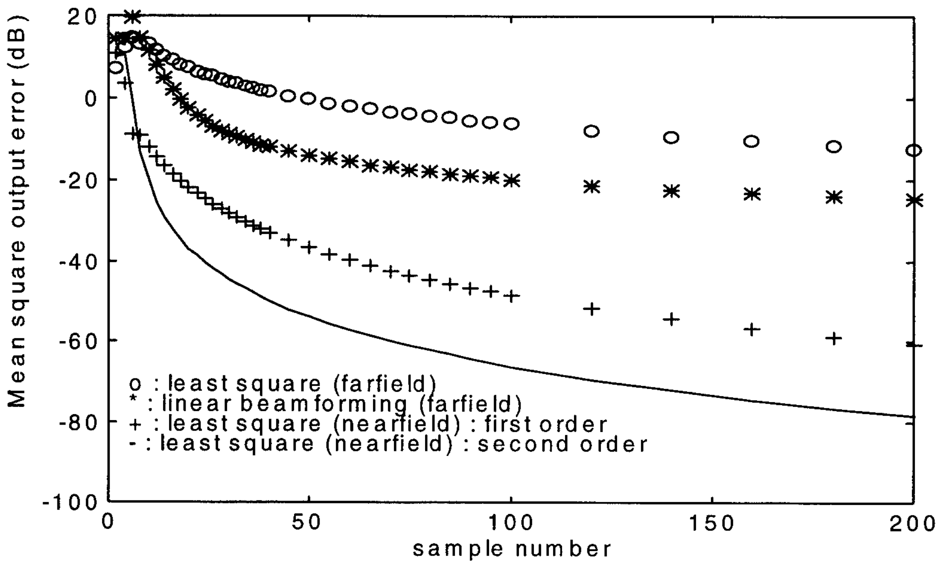

Figure 7: *Processor output error for the four techniques presented in this paper. Mean Square Output Error* = $10 * \log_{10}(E|\hat{A}(t) - A(t)|^2 / E|A(t)|^2)$ *where* $A(t)$ *is the transmit signal and* $\hat{A}(t)$ *is the reconstructed signal.*

6. Conclusion

The near field approach described here is shown to increase the accuracy of the processor output markedly. This algorithm allows spherical and plane wavefronts to be resolved simultaneously but at an increased computational load. The capability of the adaptive system in resolving near-mainlobe jamming is achieved where the conventional far field approach fails, as shown in Figure 6. Specular near field reflections are common occurrences in airborne radar and this form of array processing lends itself easily to such application.

As shown in Figure 7, first order correction is adequate, but the second order correction can describe the near field behaviour more accurately. Although several computationally intensive methods are available to determine the total number of arrivals at the receiving array using measured data, we find it useful to allow the system to operate at its maximum degree of freedom. This way, the system's random noise vector is split naturally into a number of phantom directions of arrivals with very weak signal values.

7. Acknowledgement

The authors are thankful to Jagmohan Jandu, Nick Lioutas, Layton Catford, Lloyd Clonan and Peter Drake, for their help in setting up the experiment and solving all the technical problems encountered. This work was sponsored by Defence Science and Technology Organisation of Australia.

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3a. AR Number 009-743	3b. Laboratory Number DSTO-TR-0361	3c. Type of Report Technical Report	4. Task Number DST94/060	
5. Document Date 16-5-95	6. Cost Code	7. Security Classification	8. No of Pages	17
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array processing, beamforming

18. DISCAT Subject Codes

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