TIME AND SPACE DOMAIN FILTERING FOR IMPROVED HF COMMUNICATION

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SUMMARY

High speed data transfer has until recently been generally unattainable over HF channels because of the highly disturbed nature of the channel. The development of real-time signal processors has done much to mitigate effects of channel disturbances, making it possible to substantially increase throughput. Data rates as high as 9600 bits per second can be achieved by implementing signal processors as adaptive channel equalizers. Real-time processors are also being used to adapt antenna array systems to provide improvement in spatially disturbed environments.

This recent increase in data rates has rendered feasible HF spread spectrum communication systems. Spread spectrum systems are attractive in many applications because they are resistant to narrowband interference. However, channel equalizers necessary for high data rates are susceptible to narrowband interference. Spatial or temporal prefiltering has commonly been employed to prewhiten the received data, providing improved equalizer performance. It now appears possible to realize a further improvement in system performance by combining spatial and temporal techniques into a single multidimensional prefiltering process.

This paper review some of the more popular one-dimensional techniques for temporal and spatial prewhitening of HF communication signals. Also, combined space-time prewhitening techniques area proposed. Algorithms for designing and implementing these two-dimensional whitening filters are presented. Advantages and disadvantages of the two design strategies are discussed. Particular attention is focused on system performance, computational requirements, and cost. Computer simulated results for these signal processing algorithms are presented.

I. INTRODUCTION

In the past several years there has been a marked increase in communication systems that operate in the HF range (3-30 MHz). This increase has been spawned by advances in the understanding of the HF communication channel and by the development of effective channel equalization processors (see figure 1, 2). Data rates of as high as 9600 bits/sec have been achieved in systems incorporating these channel equalizers 1 . Adaptive receiver antenna arrays are also being used to provide improved performance in spatially disturbed environments.

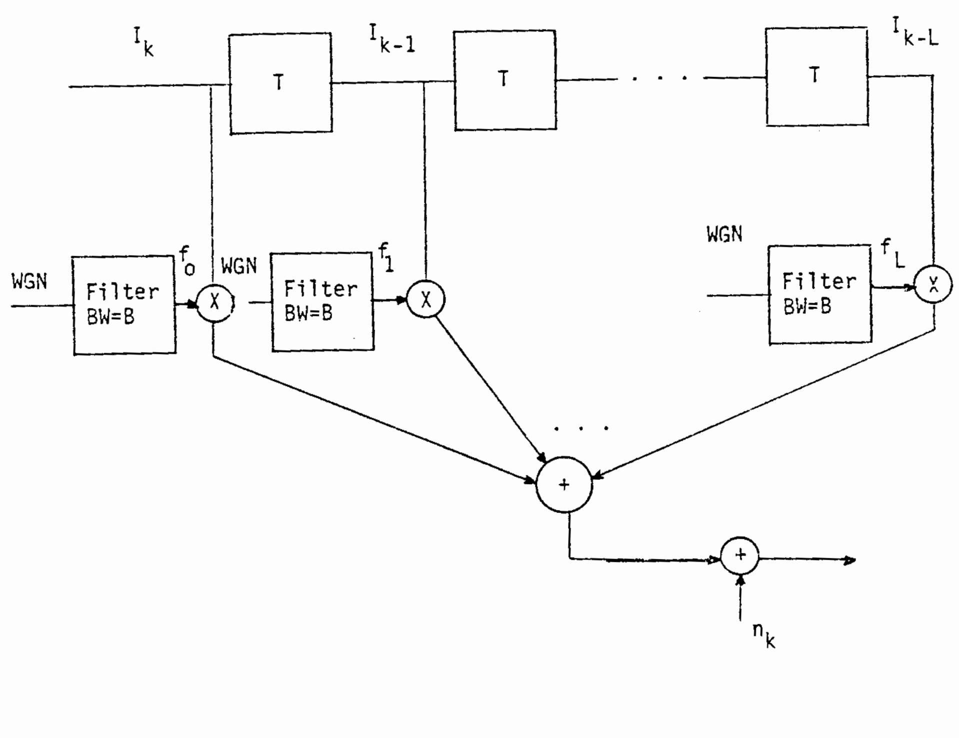


FIGURE 1. A DISCRETE CHANNEL MODEL

transmission schemes can be seriously degraded when spectral nulls in the channel fall on or near one of the tone frequencies used. 16

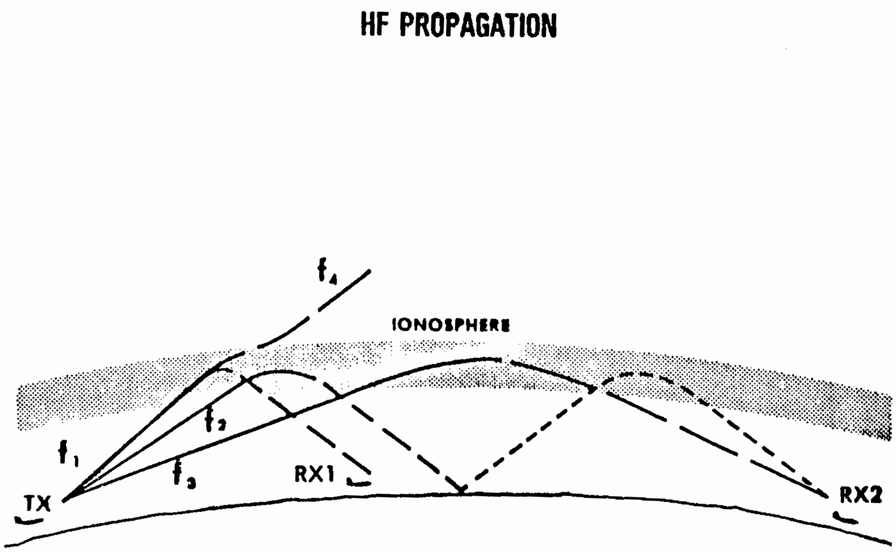


FIGURE 3. HF PROPAGATION

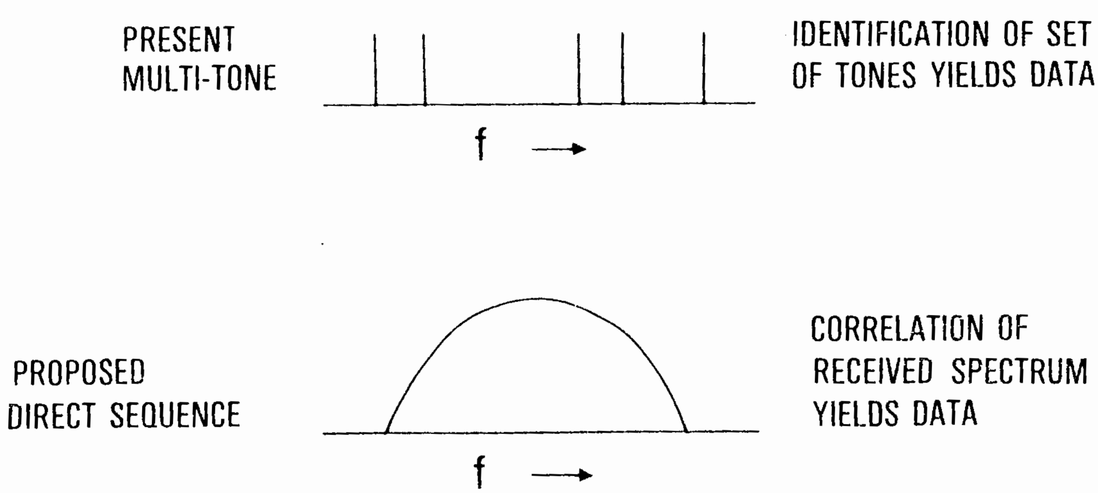


FIGURE 4. HF NARROWBAND PROCESSING

A second configuration for an HF modem uses single tone M-ary PSK modulation and a decision feedback equalizer (DFE) employing some form of recursive least-squares estimation for adjusting its tap weights [17], [18], [19], (see figure 5, and 6). In contrast to the multitone case, a single tone modem in principle suffers from none of the above consequences. Both higher data rates and low (nominally unity) peak-to-average ratio can be achieved by a single tone modem. Nevertheless, there are several drawbacks related to the DFE techniques. One disadvantage of the DFE is that, the equalizer tap weights must change more quickly than the channel does in order for the equalizer to track the channel. This implies that a fast converging algorithm for updating the DFE is needed even if the channel varies very slowly in time. The computational complexity of a fast converging algorithm such as the square-root Kalman algorithm for updating the qualizer tap weights is proportional to N^2 per symbol, where N is

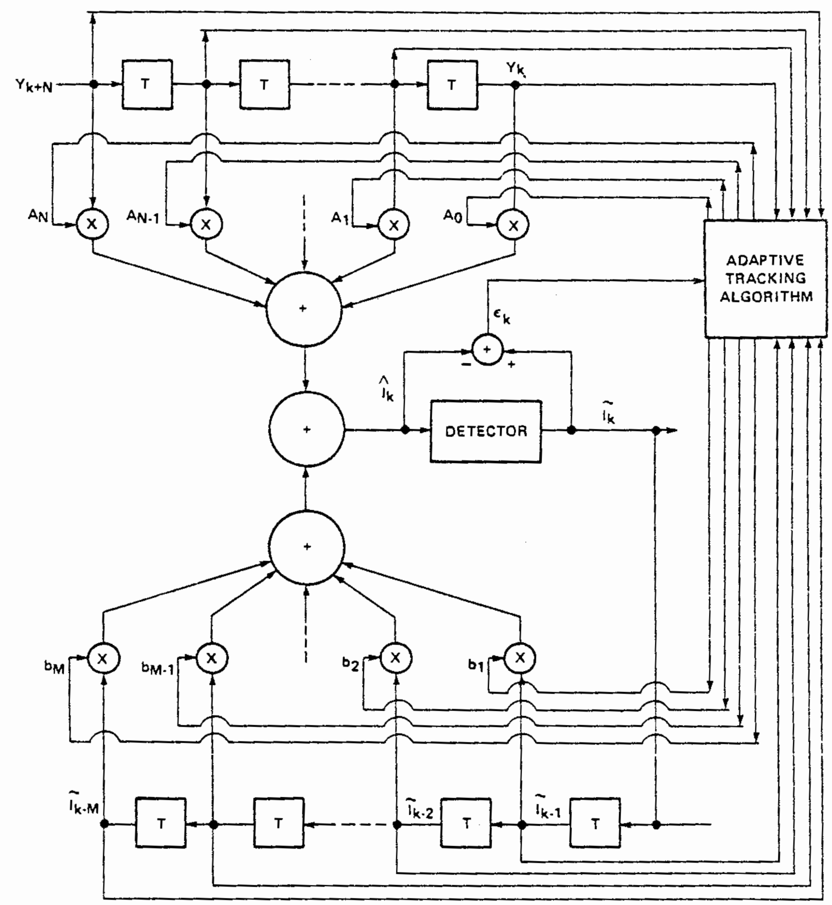


FIGURE 6. DECISION FEEDBACK EQUALIZER FOR T SECOND SPACING

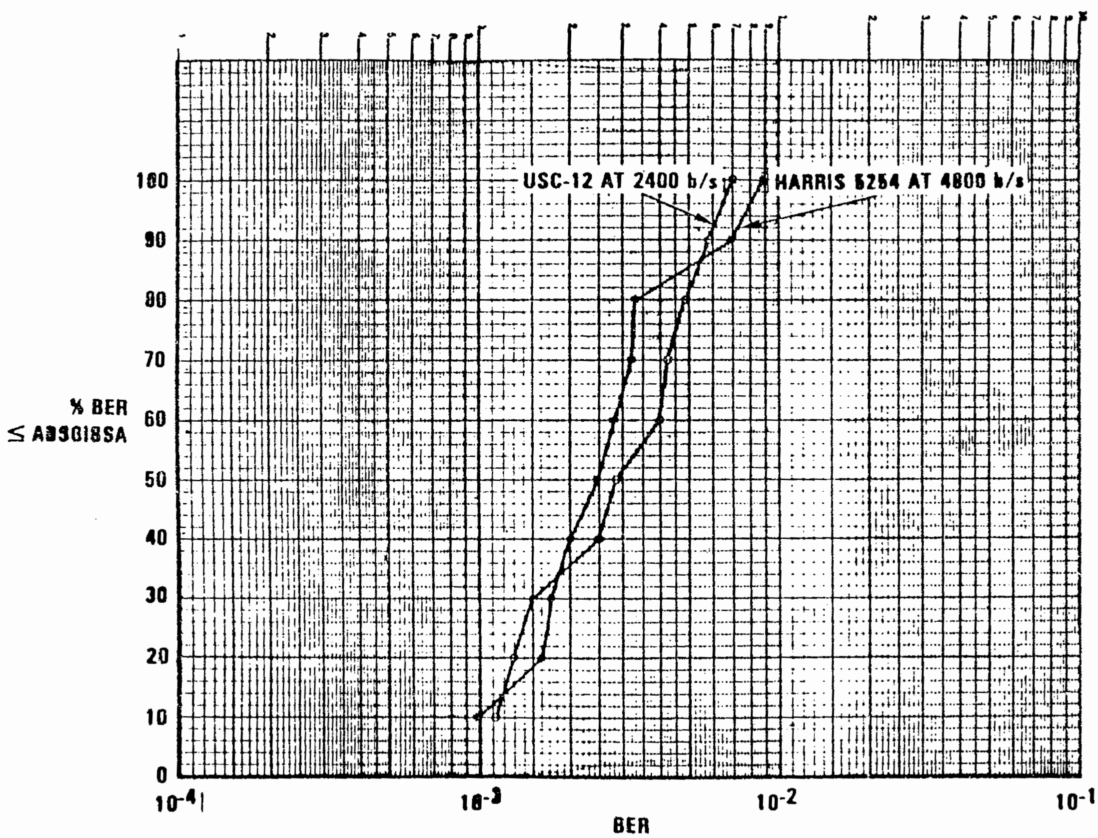


FIGURE 6A. HARRIS/USC-12 COMPARISON ASCENSION ISLAND TO CAPE CANAVERAL

TABLE 1

Modulation Alphabets, Block Parameters, Data Rates, Percentage of Training, Throughput and Maximum Allowable Multipath Spread for HF Modem Using the DDE Techniques with 2,400 keying Rate in 3 kHz Bandwidth.

Modulation Alphabets	Block Parameters	Data Rates (bps)	Percentage of Training	Through-put	Maximum Allowable Multipath Spread
4ϕ psk	M=20, N=20	2,400	50%	50%	8.3 ms
	M-32, N=16	3,200	33-1/3%	66-2/3%	6.6 ms
	M=36, N=12	3,600	25%	75%	5.0 ms
8ϕ psk	M=20, N=20	3,600	50%	50%	8.3 ms
	M=32, N=16	4,800	33-1/3%	66-2/3%	6.6 ms
	M=36, N=12	5,400	25%	75% n	5.0
16ϕ psk	M=20, N=20	4,800	50%	50%	8.3 ms
	M-32, N=16	6,400	33-1/3%	66-2/3%	6.6 ms
	M-36, N=12	7,200	25%	75%	5.0 ms

With linear DDE, the source data block ($b_0, b_1..., b_{M-1}$) is estimated by a Levinson recursive algorithm. With non-linear DDE, only $\hat{b}_0$ and $\hat{b}_{m-1}$ obtained from a Levison algorithm are kept and $\hat{b}$ to b_{m-2} discarded in the first step of estimation. The estimates, $\hat{b}_0$ and $\hat{b}_{m-1}$ are then quantized to the nearest symbols, b_0 and b_m, respectively. The decisions, $\hat{b}_0$ and $\hat{b}_m$, are substracted from the simultaneous equations. The M-2 symbols (b_1, and b_2, ..., b_{m-2}) left are estimated by a Levinson algorithm again, and so forth. The procedure described here is used recursively until all the unknown symbols are obtained. In both linear and nonlinear DDE techniques, the channel is estimated by a steepest-decent algorithm or a pseudo-inverse algorithm.

As a linear equalizer, the linear DDE lacks the ability to cope with the severe fading dispersive HF radio channels. However, the nonlinear DDE works extremely well under the same severe channel conditions. The nonlinear DDE can achieve performance similar to the DFE or better.

The nonlinear DDE (NDDE) can achieve data rates up to 7,200 bps in a 3 kHz bandwidth with wide multipath spread (e.g., 5 ms).

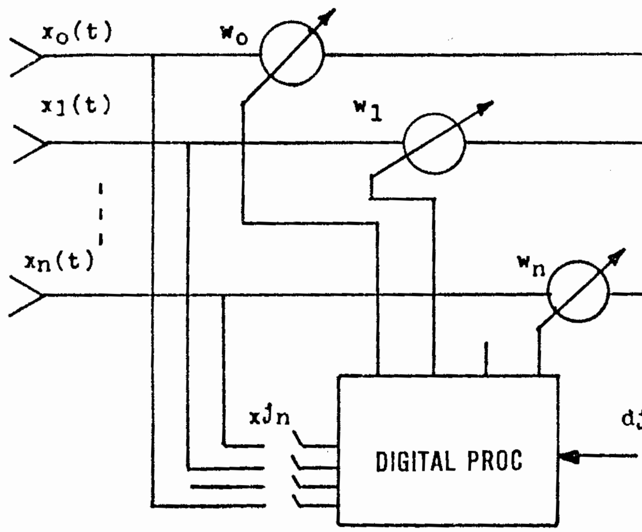
2. Adaptive Whitening

Spread spectrum, direct sequence or pseudo-noise (PN) modulation is employed in digital communication systems to reduce the effects of interference. When the interference is narrow band the cross-correlation of the received signal with the replica of the PN code sequence reduces the level of the interference by spreading it across the frequency band occupied by the PN signal. Thus, the interference is rendered equivalent to a lower level noise with a relatively flat spectrum. Simultaneously, the cross-correlation operation collapses the desired signal to the bandwidth occupied by the information signal prior to spreading.

The interference immunity of a PN spread spectrum communication system corrupted by narrow band interference can be further improved by filtering the signal prior to cross-correlation, where the objective is to reduce the level of the interference at the expense of introducing some distorton on the desired signal. This filtering can be accomplished by exploiting the wideband spectral characteristics of the interference. Since the spectrum of the PN signal is relatively flat across the signal frequency band, the presence of a strong narrow band interference is easily recognized. Then the interference can be suppressed by means of an appropriately designed linear filter (see figure 7).

Hsu and Giordano [12] considered the problem of narrow band interference estimation and suppression by means of two linear prediction algorithms, the Burg algorithm [13], [14], and the Levinson algorithm [13], [15]. The channel through which the PN spread spectrum signal is transmitting was assumed to be nondispersive. Results were presented on the effectiveness of the linear prediction filter in suppressing the interference. Performance was measured in terms of signal-to-noise ratio at the output of the PN correlator.

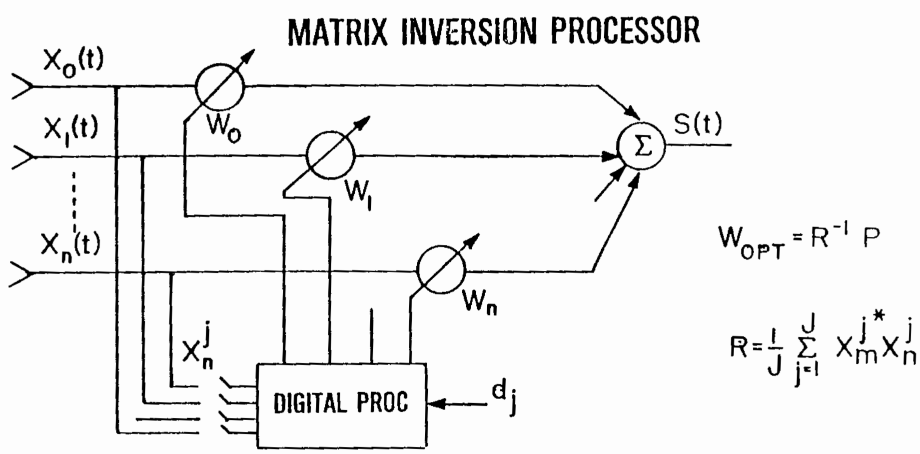
Proakis and Ketchum extended the results obtained by Hsu and Giordano on filter requirements and characteristics in single and multiple frequency band interference. In addition to a nondispersive channel, they consider the transmission of the PN spread spectrum signal over a channel characterized by fading and multipath (time dispersion). This serves as a model for radio channels such as HF. The existance of time dispersion in the received signal necessitates some means for dealing with this type of distortion at the receiver. We have considered the use of an adaptive decision-feedback equalizer preceding the PN correlator for mitigating the effects of time dispersion due to multipath and the linear, interference suppression filter (see figure 8).



$W_{opt} = R^{-1} P$

1. Sample $X_n(t)$
2. Form $R = \frac{1}{J} \sum_{j=1}^{J} X_m^{j*} X_n^j$
3. Form $P = d_j \; X_j^T$
4. Invert $\underline{R}$
5. Form Product $R^{-1}P$
6. Apply Weights

FIGURE 10. MATRIX INVERSION PROCESSOR



NUMBER OF MULTIPLES TO COMPUTE R^{-1}

ALGORITHM	PER UPDATE	PER BLOCK (2n SAMPLES)
BRUTE FORCE	$\frac{1}{2}n^3+\frac{3}{2}n^2+\frac{5}{2}n$	$\frac{3}{2}n^3+2n^2+2n$
CHOLESKY	$\frac{1}{6}n^3+2n^2+\frac{5}{6}n$	$\frac{7}{6}n^3+\frac{5}{2}n^2+\frac{2}{3}n$
SQUARE ROOT	$\frac{5}{2}n^2+\frac{9}{2}n$	$\frac{5}{2}n^3+8n^2+\frac{3}{2}n$
FACTORED SQUARE ROOT	$2n^2+6n$	$\frac{5}{3}n^3+7n^2-\frac{5}{3}n$
FACTORED INVERSE	$\frac{5}{2}n^2+\frac{7}{2}n$	$3n^3+8n^2$

FIGURE 11.

$$S_x(Z_1, Z_2) = \left| \frac{B(Z_1, Z_2)}{A(Z_1, Z_2)} \right|^2 \qquad (2)$$

where $B(z_1, z_2)$ is a polynomial of degree q_1 in z_1, and q_2 in z_2, and $A(z_1, z_2)$ is a polynomial of degree p_1 in z_1, and p_2 in z_2.

It is apparent that interference signals and some nose appear as areas of high power density in the 2-D frequency plane. An appropriate interference rejection filter, then, is one that spectrally whitens the data, as this flattens such peaks. If the signal power is below the background noise power (which is often the case in practice), a whitening filter has little effect on the desired signal, so SIR and SNR gains are realized at the filter output. If the signal power is substantially above the noise power level, a whitening filter attempts to attenuate the signal along with the interference. This effect is unavoidable when the signal incident angle is unknown, as the whitening filter cannot distinguish between signal and interference. However, even in this case SIR and SNR gains may be realized at the filter output.

The optimum whitening filter for ,x(m,n). has transfer function from each input to each output of the form

$$H(e^{jw_1}, e^{jw_2}) = \frac{A(e^{jw_1}, e^{jw_2})}{B(e^{jw_1}, e^{jw_2})} \qquad (3)$$

where A and B are given in equation (2). Since H contains a denominator terms, it is an infinite impulse respone (IIR) filter. IIR filters are undersirable because they pose a stability problem, (especially when the filters are time varying, as is the case here), and because the B coefficients are often difficult to estimate accurately without a large amount of data (especially when there are sharp nulls in the spectrum). Because of these problems and because spectral nulls are not formed by interference signals, there seems to be no practical reason to incorporate the coefficients into the filter. A practical alternative filter is

$$H(e^{jw_1}, e^{jw_2}) = A(e^{jw_1}, e^{jw_2}) \qquad (4)$$

where A is obtained by either:

1. Modeling the data with the ARMA model in equation (2) and using only the estimated AR coefficients in (4).

2. Modeling the data with an AR model and using the estimated AR coefficients in (4).

The main advance of the first alternative is that sharp spectial peaks present in the data are more effectively nulled; the main advantage of the second alternative is that some of the MA filtering is performance because the entire spectrum is approximated by the AR model. Thus, the two filter models represent a tradeoff between effectiveness in eliminating narrowband interference and effectiveness of spectrally whitening the data.

B. Two-Dimensional ARMA Algorithms

In this section we propose one class of AR coefficient estimation algorithms derived by considering the observed data as a sample from a two-dimensional stationary ARMA process. The derivation makes use of the fact that the antenna array elements are collinear and equally spaced, which results in inefficicent use of the data in generating autocorrelation estimates.

Consider the 2-D semi-casual ARMA recursion

$$\sum_{i=-P_1}^{P_2} \sum_{j=0}^{P_3} a_{ij} X(m-i, n-j) = \sum_{i=-q_1}^{q_2} \sum_{j=0}^{q_3} b_{ij} V(m-i, n-j) \ , \ a_{00} = 1 \qquad (5)$$

If we follow the earlier suggestion of using only AR coefficients in the filtering operation, the corresponding prefilter outputs are given by

$$e(m,n) = \sum_{i=-P_1}^{P_2} \sum_{j=0}^{P_3} a_{ij} X(m-i, n-j) \ , \ 1 \le m \le M \qquad (6)$$

It can be seen from (6) that p_1 and p_2 must be chosen so that $1 \le m-i \le M$ for $-P_1 \le i \le P_2$. In general this may require a different choice of p_1 an p_2 for each of the M filter outputs. Thus, a different set of AR coefficients is in general necessary for each of the M outputs.

An effective AR coefficient estimation procedure can be derived by appealing to the well-known Yule Walker equations, which are found by multiplying both sides of equation (5) by $x^*(m-k, n-l)$ and then taking the expected value to give

$$r_x(k,l) + \sum_{i=-P_1}^{P_2} \sum_{j=0}^{P_3} a_{ij} r_x(k-i, l-j) = 0 \ , \ \text{For } l > q_3 \qquad (7)$$

By replacing exact autocorrelations in (7) with ones estimated from the given data, we can form a matrix of approximate Yule Walker equations whose solution yields AR coefficient estimates. To this end,

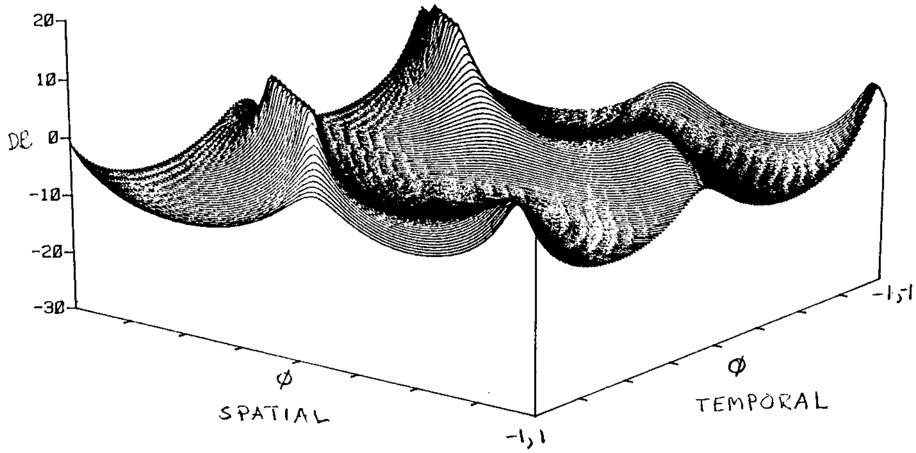


FIGURE A.

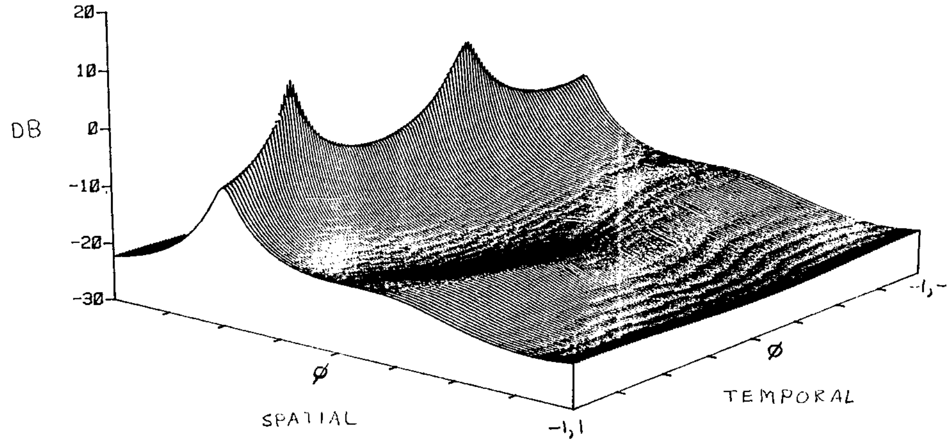


FIGURE B.

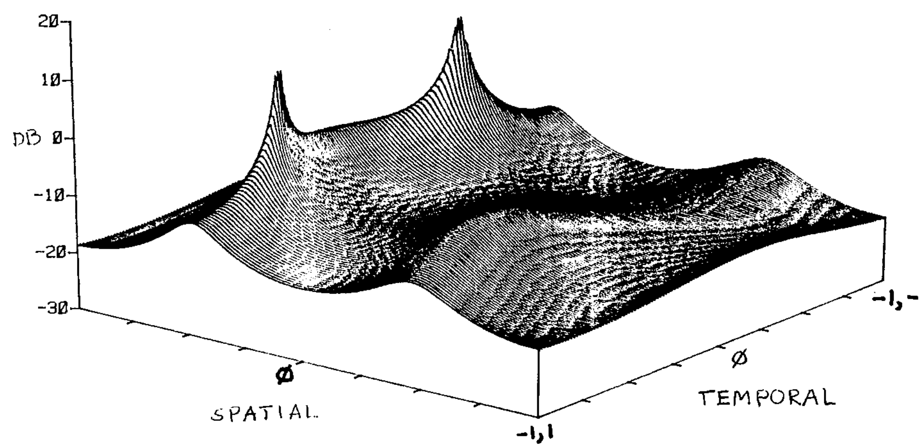


FIGURE C.

DISCUSSION

E.W.Lampert, Ge

What are the classes of discriminants you are using for your adaptive array system?

Author's Reply

No discriminants are used for the 2-dimensional spectral whiting filter discussed in the paper. The prefilter merely performs spectral smoothing. In terms of spatial filtering a number of discriminants are possible. They are, direction of arrival, a prior knowledge of desired signal and knowledge of receive frequency (frequency hopped).

R.W.Jenkins, Ca

Have you implemented the matrix inversion techniques yet, and if so, what are the matrix inversion times achieved?

Author's Reply

For the worst case matrix inversion would require $\frac{1}{2}n^3 + \frac{3}{2}n^2 + \frac{5}{2}n$ complex multiplies, where n is the number of antennas. For example, if $n = 3$ then $\frac{27}{2} + \frac{27}{2} + \frac{15}{2} = \approx 34$ complex multiplies. At 100 m flops the update time is 34 μs for Cholesky $\frac{27}{6} + 18 + 3 = 26$, i.e. 2.6 μs update approx.