dynetics, Inc.
P. O. DRAWER B
HUNTSVILLE, ALABAMA 35805

TR-78-BMDSCOM-0041-019

FINAL REPORT

OPTIMAL NONLINEAR FILTERING FOR JAMMING SUPPRESSION

OCTOBER 1978

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prepared for

BALLISTIC MISSILE DEFENSE SYSTEMS COMMAND
P. O. BOX 1500
HUNTSVILLE, ALABAMA 35807

DASG60-75-C-0041 Mod P0009

LEVEL II

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This Final Report documents the results and conclusions established by Dynetics, Inc. on Contract DASG60-75-C-0041, Mod P0009 for the period of technical performance from 1 August 1977 through 30 September 1978. The objective of this effort is to investigate the application of nonlinear signal processing techniques to the suppression of jammer signals that are potential threats to BMD radars. A broad class of potential jamming signals are identified on the basis of their statistical and spectral characteristics. Six candidate nonlinear → next page.

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ABSTRACT

This Final Report documents the results and conclusions established by Dynetics, Inc. on Contract DASG60-75-C-0041, Mod P0009 for the period of technical performance from 1 August 1977 through 30 September 1978. The objective of this effort is to investigate the application of nonlinear signal processing techniques to the suppression of jammer signals that are potential threats to BMD radars. A broad class of potential jamming signals are identified on the basis of their statistical and spectral characteristics. Six candidate nonlinear processing techniques, which were investigated for their applicability to the various jamming signals, are discussed. Three techniques are identified as applicable for BMD signal processors. These techniques were implemented in simulated signal processors, and their advantage over conventional matched filters was determined for several categories of jamming signals.

R. C. Cashion

Approved by:

M. C. Budge, Jr., Ph.D.
Project Manager

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1.0 INTRODUCTION

This report summarizes the activities performed by Dynetics, Inc. on Contract DASG60-75-C-0041, Mod P0009 during the period of technical performance from 1 August 1977 through 30 September 1978. The objective of this effort was to investigate the application of nonlinear signal processing techniques to the suppression of jammer signals that are potential threats to BMD radars. This effort has been divided into three major task areas: 1) jammer signal characterization, 2) investigation of modern detection and estimation techniques, and 3) evaluation of candidate signal processing algorithms. During the first period of technical performance, Tasks 1 and 2 received the focus of attention. The applicable results and conclusions were documented fully in Interim Report SAPR-78-BMDSCOM-0041-001, "Optimal Nonlinear Filtering for Jamming Suppression" (Reference 1). Those findings are summarized in this Final Report. Task 3, the evaluation of candidate signal processing algorithms, is the major topic of this report. Six nonlinear signal processing techniques were perceived as having potential applicability to BMD. Three of these techniques have subsequently been identified as applicable to the somewhat unique environment associated with BMD.

1.1 REPORT SUMMARY

Section 2.0 of this report is a summary of the investigations conducted under this contractual effort. The first area investigated was the classification of jammers signal types. This classification was developed from considerations of the characteristics of jamming signals that might directly affect the strategy or performance of jamming mitigation techniques. No specific foreign BMD jamming signals are known, and a survey of air defense radar jamming yielded little insight into BMD jamming signals.

The usual assumption concerning jamming signals for BMD is that they are wide band Gaussian signals. Under such an assumption, the optimum detection processing is well known to be the standard matched filter. Therefore, for the purposes of considering nonlinear processing techniques, it is paramount that other jamming signal types are allowed. In this report, several other types of jamming signals are considered with no judgment being made as to the likelihood of their occurrence. It must be pointed out, however, that the conditions under which wide band Gaussian jamming is optimum for the offense are quite nebulous. Consequently, consideration of other types of jamming signals should not be construed as academic. A total of thirteen generic jamming signal categories are discussed.

During the early phases of this study, research led to six nonlinear techniques that might provide some measure of improvement for one or more of the jamming signal types considered. These techniques were discussed in some detail in the Interim Report (Reference 1). Section 2.2 of this report summarizes these discussions. Of the six techniques initially examined, three were found to be worthy of detailed evaluation. These three were the Adaptive Generalized Matched Filter (AGMF), non-coherent clipping/blanking, and the Filter-Nonlinearity-Filter (FNF). These three techniques were evaluated by means of digital signal processing simulations. The simulations were exercised in a Monte Carlo fashion to estimate the improvement in signal plus interference-to-interference power ratio relative to that obtained by conventional matched filter processing. Where feasible, detection performance was directly estimated by counting detections. Section 3.0 documents the results for these exercises.

The AGMF, non-coherent clipping/blanking in the frequency domain, and the FNF processor are all suitable for narrow band jamming. Three forms of the FNF processor were considered: Taylor processing and two techniques of independent AGC's on sub-bands of the signal of interest. The two AGC techniques are denoted as peak-normalizing and RMS-normalizing AGC processing. Under similar conditions in the presence of narrow band interference, these techniques exhibited the improvement ratios given in the following table:

TABLE 1-1. COMPARISON OF NONLINEAR TECHNIQUES FOR THE REDUCTION OF NARROW-BAND JAMMING INTERFERENCE

NONLINEAR PROCESSING TECHNIQUES	IMPROVEMENT FACTOR (dB)
AGMF	2.9
Non-coherent clipping	2.9
Non-coherent blanking	2.7
Taylor FNF	2.3
Peak-normalizing AGC FNF	1.2
RMS-normalizing AGC FNF	1.7

This table shows that comparable performance was obtained by at least one form of the three basic techniques considered. Although not indicated by the table, the AGMF suffers a drawback that the other techniques do not possess. That is, non-stationary jamming signals can defeat the estimation procedure required by the AGMF.

The non-coherent clipping and blanking techniques are applicable to time domain processing for the reduction of the effects from pulsed jamming signals. Their performance was demonstrated to be quite spectacular for pulse jamming with even 50% duty cycles.

Section 4.0 of this report contains the conclusions established from all three phases of this study: the characterization of jamming signal types, the identification of candidate nonlinear processing techniques, and the performance evaluation of those techniques potentially attractive for BMD.

2.0 SUMMARY OF INVESTIGATIONS

2.1 CLASSIFICATION OF JAMMER SIGNALS

This task was concerned with the identification of those characteristics of jammer signals that are distinguishable from thermal noise and, as such, could affect the optimal detection strategy or processing. Since the optimum processor for signal detection in thermal (i.e., white, stationary, Gaussian) noise has been established to be the matched filter, (Reference 1) those characteristics of jammers which may differ from thermal noise may result in improved performance with nonlinear processing techniques. Thus, those jamming signal characteristics that differ from thermal noise are of interest, specifically non-white power spectral density, non-stationarity, and non-Gaussian properties.

With the above considerations in mind, a jammer classification scheme was developed. The initial classification breakdown was on the basis of stationarity, as follows:

(1) Stationary-I: Jammer signals whose statistics and power spectral density are essentially constant.

(2) Stationary-II: Jammer signals whose statistics and power spectral density are time varying, but remain constant for a period of time equal to several receiver processing time windows.

(3) Non-stationary: Jammer signals whose statistics and/or power spectral density vary significantly over a period of time equal to the length of a single receiver processing window.

The rationale for the above divisions is based upon the processing strategy suitable for each category. For Stationary-I signals, jammer characteristics may be established by long term observation. For Stationary-II type signals, an observation window immediately prior to the signal processing window is required to estimate the pertinent jammer signal parameters. Finally, no prior estimate of jammer signal characteristics is feasible for non-stationary signals. Thus, adaptive processing is apparently limited to Stationary-I and Stationary-II categories.

Beyond these three major divisions, many possible jamming signal types are conceivable. For the purposes of this study the Stationary-I jammer signals are subdivided further into Gaussian noise and non-Gaussian interference according to the characteristics of the respective probability density functions. Stationary-II and Non-stationary jammer signals are subdivided using the concept of true random noise and that of structured interference, i.e., distinct jamming signal forms. This chain of subdivisions leads to the thirteen categories of jammer noise as depicted in Figure 2-1. These categories are certainly not exhaustive; however, they encompass a wide enough range of possibilities to provide a comprehensive set of "threats" for any candidate signal processor. Amplifying discussions of these jammer categories are provided in Reference 1.

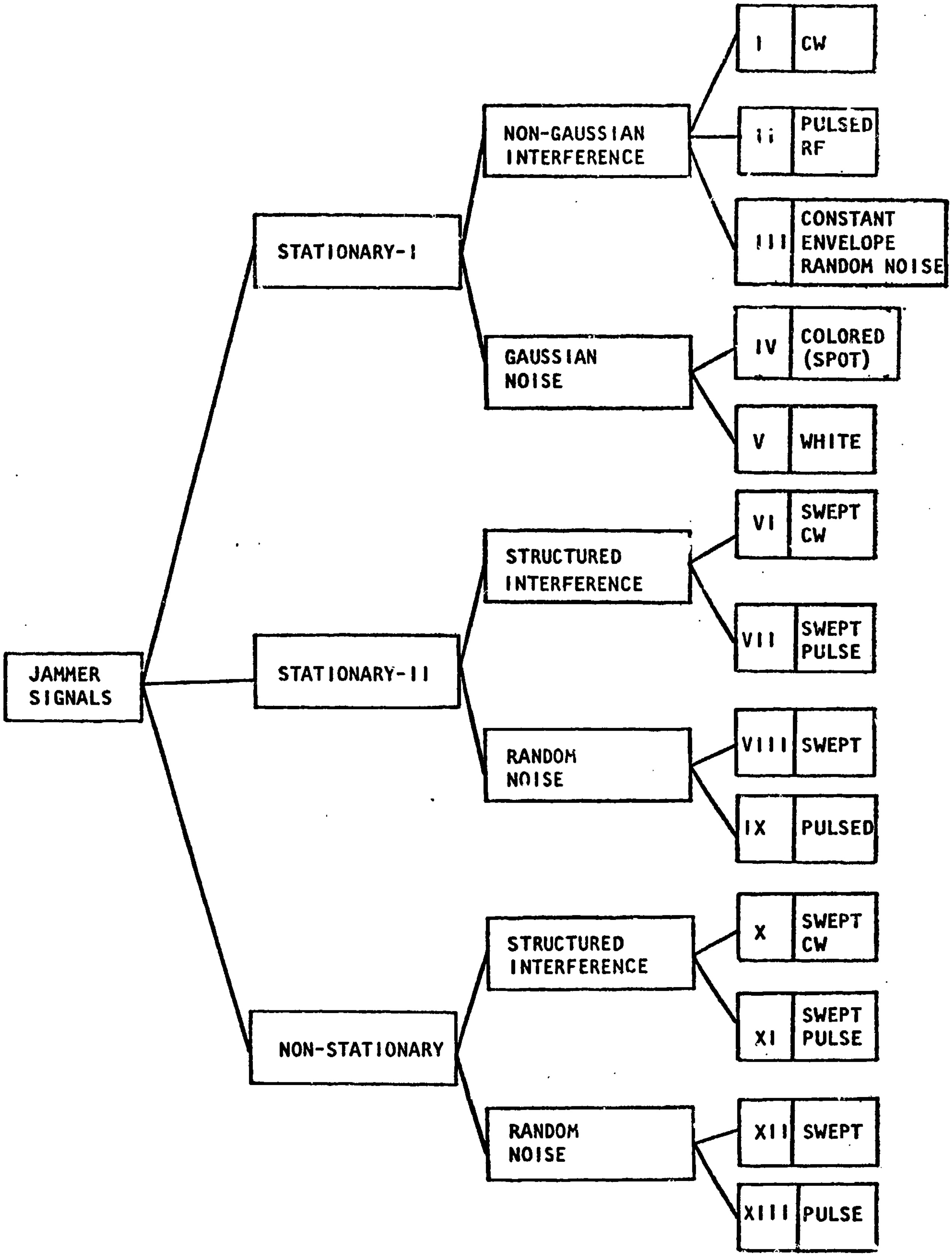


FIGURE 2-1. JAMMER SIGNAL CATEGORIZATION

2.2 CANDIDATE NONLINEAR PROCESSING TECHNIQUES

A total of six nonlinear processing techniques were examined in some detail during the contract performance period. This subsection discusses these six techniques. A more elaborate discussion of these candidate nonlinear signal processing techniques is presented in Reference 1. Results from the testing of three techniques deemed to be applicable to BMD are presented in Section 3.0.

2.2.1 Adaptive Linear Processing

Linear processing has long been applied to problems in detection and estimation. The classical matched filter processing of radar signals is known to be optimum for the detection of signals in white, stationary Gaussian noise. The matched filter is also the linear processing that maximizes the signal-to-noise power ratio for white, stationary noise, regardless of the probability distribution of the noise. Similarly, for colored, stationary noise, the generalized matched filter (GMF) is the linear processor that maximizes signal-to-noise power ratio. Futhermore if the probability distribution of that noise (interference) happens to be Gaussian, the GMF implements an optimum detection criterion. Thus, for the case of stationary interference, the GMF appears attractive. However, proper implementation of the GMF requires a knowledge of the power spectral density of the interference. Consequently, the performance of the GMF is dependent upon the ability to estimate the power spectral density of the interference. For Stationary-I type jamming signals this can be done with any desired degree of precision. For Stationary-II jamming signals (a more general case), the spectral estimation problem limits the achievable performance of the GMF. An implementation of an adaptive GMF was investigated and the results are presented in Section 3.1.

2.2.2 Non-Parametric Detection

There are a number of detection schemes which have been developed with the idea that the statistics of the corrupting noise are not known. These schemes are known as non-parametric or distribution-free detectors. Two such techniques were considered during the course of this study, the sign detector and the Wilcoxon detector. Both of these techniques have the property that, if the median of the corrupting noise or interference is 0, then these detectors provide a constant false alarm rate, regardless of any other characteristics of the distribution of noise.

The sign and Wilcoxon detectors are described in Reference 1. Additional discussions of these detectors are provided in Reference 2. It was shown that asymptotic relative efficiency (ARE) for a sign detector for a constant signal in Gaussian noise was 0.64, compared to the optimum (linear) detector. The ARE for the Wilcoxon detector under the same circumstances was determined to be 0.995. In the case of constant amplitude interference, i.e. Category III, the Wilcoxon detector has an ARE of 1.5 compared to the linear (matched filter) detector. Thus, superior performance with the Wilcoxon detector should be obtained for Category III jamming, in the case of constant signals.

Attempts to apply these non-parametric schemes to the BMD radar detection problem led to several severe problems. These problems result from

- Unknown signal phase
- Unknown Doppler frequency
- Pulse compression waveforms

Furthermore, it was determined that the constant false alarm features of these detectors could be defeated by narrow band jammers, which reduce the number of statistically independent samples of the interference. Because of these characteristics, non-parametric detectors were not considered to be suitable for jamming mitigation by BMD radars.

2.2.3 Non-Coherent Clipping/Blanking

One approach to the reduction of jammer interference is the elimination of certain components of an observation when those particular components are badly corrupted by the jammer. In such an operation signal fidelity is possibly comprised in order to enhance the remaining signal components relative to the remaining noise components, as compared to the original signal and noise components. A simple example of this is the clipping operation depicted in Figure 2-2 for a finite signal observation vector. Although this type of clipping is normally thought of in the time domain it is equally applicable to the frequency domain, or any arbitrary domain of signal definition. For the present study, time domain and frequency domain clipping was considered. The designator "non-coherent" is

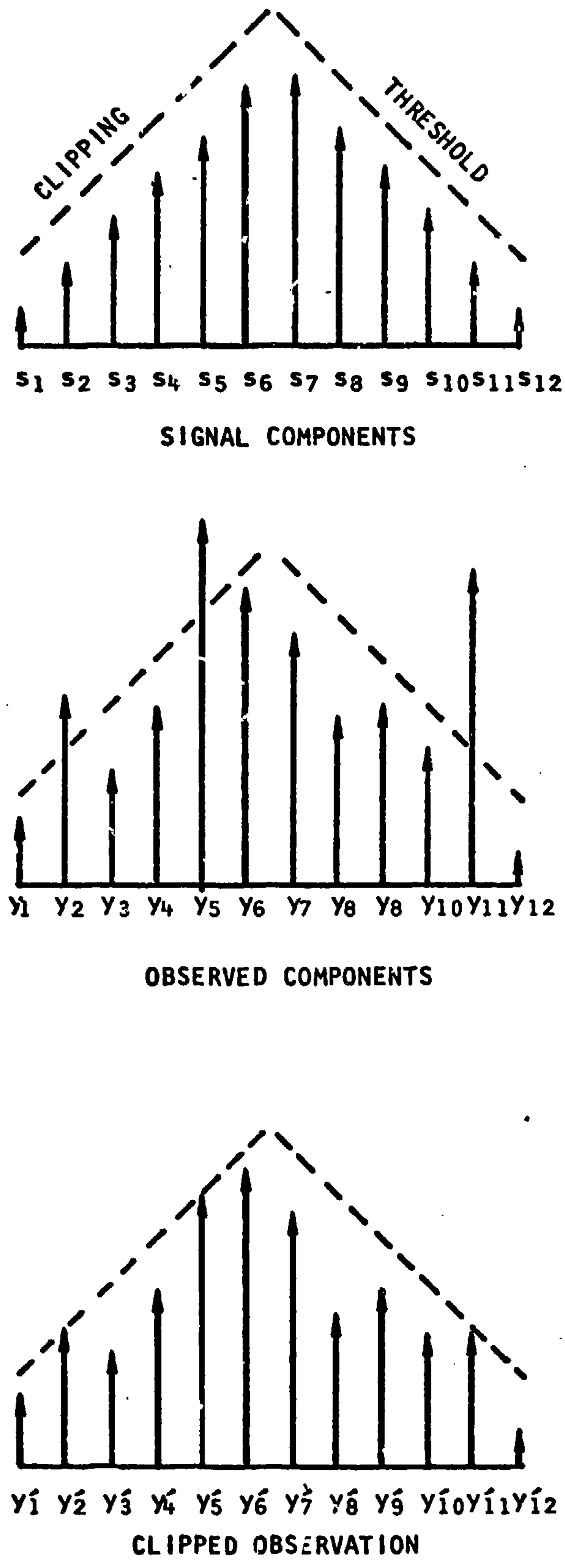


FIGURE 2-2. EXAMPLE OF NON-COHERENT CLIPPING

applied because the clipper is invoked whenever the magnitude of a component of signal plus interference exceeds a certain threshold. A similar procedure has also been evaluated which is called a non-coherent blanker. In this case, components of an observation vector are actually removed when their magnitude exceeds some threshold. This procedure is outlined in Figure 2-3

Mathematically, the non-coherent clipper and blanker procedures transform observation components (e.g. sampled signal plus interference) as follows:

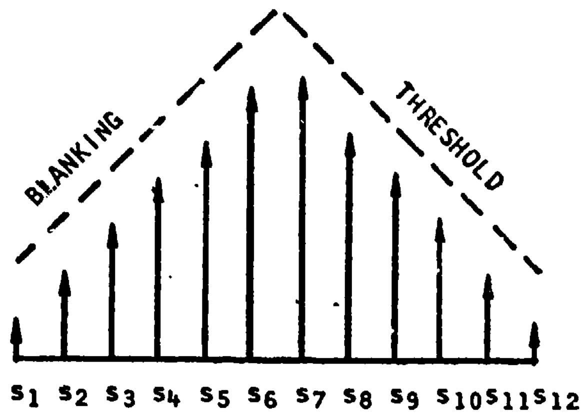
$$y_i' = \begin{cases} \dfrac{y_i c_i}{|y_i|} & \text{if } |y_i| > c_i \\ y_i & \text{otherwise} \end{cases}$$

for the clipper, and

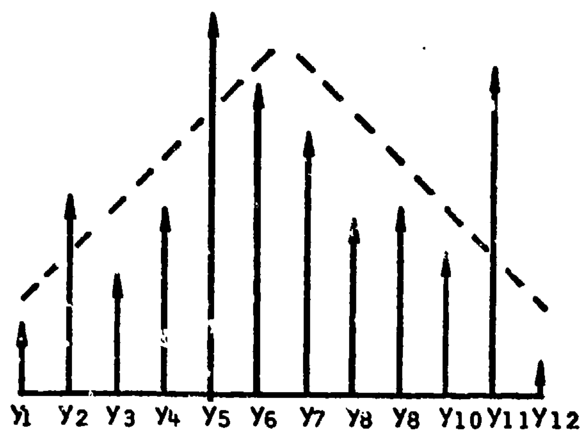
$$y_i'' = \begin{cases} 0 & \text{if } |y_i| > c_i \\ y_i & \text{otherwise} \end{cases}$$

for the blanker, where c_i is a threshold.

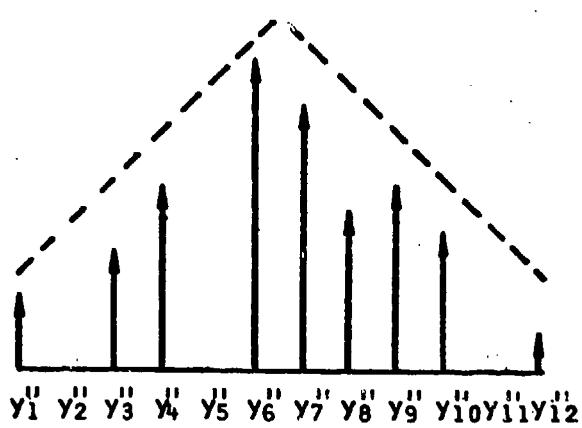
The initial approach to defining a suitable procedure for implementing these techniques was to determine the required threshold levels which maximized the probability that the ratio of signal



SIGNAL COMPONENTS



OBSERVED COMPONENTS



BLANKED OBSERVATION

FIGURE 2-3. EXAMPLE OF NON-COHERENT BLANKING energy-to-noise energy (or average power) in the transformed observation exceeded the original observation. This approach was carried to the point of generating the required probability distributions. However, unwieldy computational problems forced an abandonment of that approach in favor of a simulation study. The simulation study considered non-coherent clipping and blanking in both the time and frequency domain and treated the thresholds parametrically. The results are presented in Section 3.1.

2.2.4 Signal Space Blanker

In reference 1 a technique was developed which was intended to counter the effect on the false alarm rate that might be encountered by a fast swept (in frequency) jamming signal. A signal of this type (Categories X and XII) could defeat CFAR or AGC circuitry circuitry whose response time was insufficient to properly adjust the gain or threshold.

The basic premise for signal space blanking is that the vector space which spans the entire domain of the input to the signal processor contains a constant jamming power. However, that portion of the vector space spanned by the possible signals of interest does not necessarily contain a constant jamming power. As the jammer sweeps through the pass band of the signal processor, the signal space blanker, knowing the total jamming power, senses a decrease in the jamming power in that portion of the vector space orthogonal to the signal space. From this the jamming energy in the signal space can be determined and the detection threshold adjusted accordingly. A suggested implementation of the signal space blanker is shown in Figure 2-4, based upon the mathematical development in Reference 1.

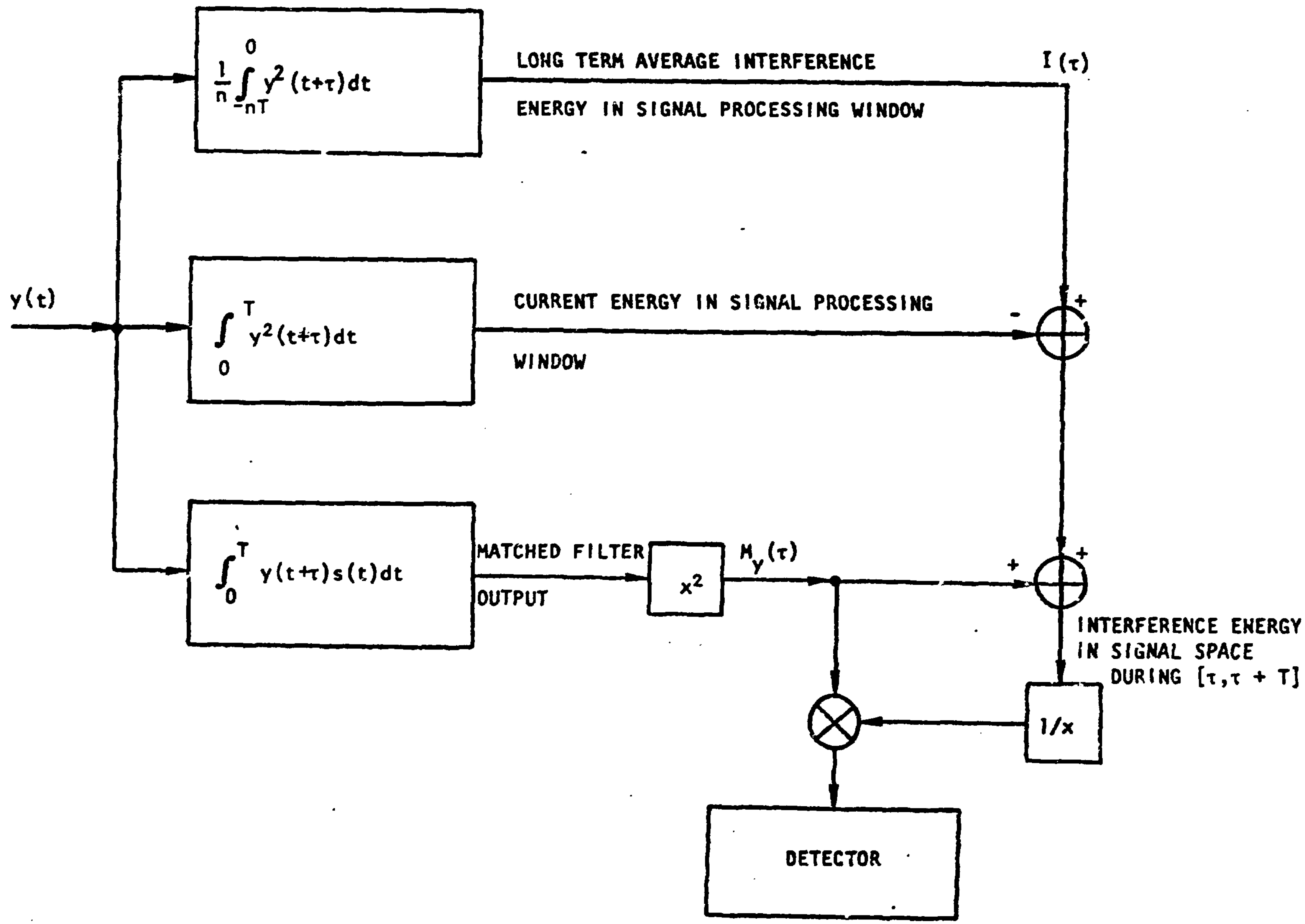


FIGURE 2-4. IMPLEMENTATION OF SIGNAL SPACE BLANKER

Because the signal space blanker appeared to have utility for a very limited class of jamming signals, the investigation of this particular nonlinear technique was not pursued beyond the initial development.

2.2.5 Markov Processors

A discrete parameter Markov chain is a stochastic process, $\{X(t_k),\ k=1,2,\ \ldots\}$, for which the conditional probability, $P_{X(t_i)|\{X(t_j),\ j<i\}}(\alpha|\beta_1,\beta_2,\ \ldots\ \beta_{i-1})$, is dependent only on the most recent previous value of the process, $X(t_{i-1})$, i.e.,

$$P_{X(t_i)|\{X(t_j),\ j<i\}}(\alpha|\beta_1,\beta_2,\ \ldots\ \beta_{i-1}) = P_{X(t_i)|X(t_{i-1})}(\alpha|\beta_{i-1}) \quad (2\text{-}1)$$

It is possible to structure a post-detector signal processing scheme such that there are a finite number of system "states" and the probability of a given state occurring depends only on the previous state.

Markov processors have been previously used in radars which operate by illuminating a target with many consecutive pulses. Normally, a BMD radar does not enjoy this advantage, however, Markov processing may be adapted to BMD radar operating modes. For example, several successive verification pulses could be processed by a Markov processor. In this case, the state transitions would depend upon whether or not a given verification pulse was correlated (range and angle) with the preceding pulse or possibly another of the preceding verification pulses. Markov processors are known to be inherently immune to certain types of jamming, particularly those types of jamming signals that may appear sporadically from pulse to pulse. Due to the fact that relatively long times on target are required for Markov processing and the fact that state transitions might depend upon system level considerations, the application of Markov processors to BMD has not been developed during these investigations.

2.2.6 Filter-Nonlinearity-Filter

A class of nonlinear processing techniques was considered which consists of a filter-nonlinearity-filter (FNF) pre-processor followed by a conventional matched filter. Figure 2-5 illustrates the general form of the FNF processor. In the most general form a FNF processor consists of a set of (usually contiguous) linear filters whose transfer functions are denoted by $G_1(s)$, $G_2(s)$,..., $G_n(s)$. Each of the filters is followed by some nonlinear function, denoted $F_1(x)$, $F_2(x)$, ... $F_n(x)$. Following each nonlinearity is another filter, $H_1(s)$, $H_2(s)$, ... $H_n(s)$ and the outputs of all of these filters are summed to form the output.

A form of the FNF processor termed the Taylor processor (Reference 3) was simulated in order to assess its potential for jammer suppression. The Taylor processor that was implemented consisted of a bank of 8 contiguous bandpass filters followed by a limiter in each channel and another bandpass filter identical to the initial channel separation filter. The summation of all 8 channels was then processed by a matched filter, compensating for the filter bank in the Taylor processor. The theory behind the design of the Taylor processor is that jamming signals which have narrower bandwidth than the signal would appear in less than the full eight channels occupied

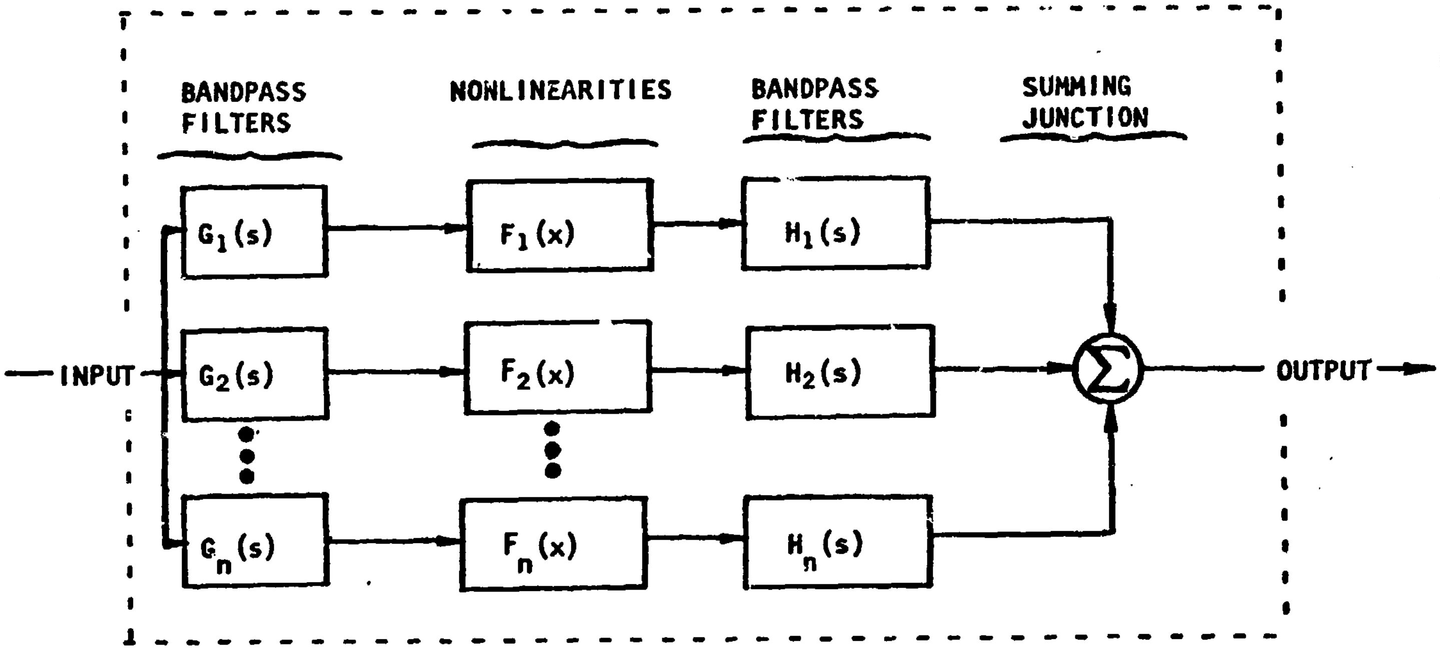


FIGURE 2-5. AN FNF PROCESSOR

by the signal. Thus, the jamming power in those affected channels can be limited without affecting the signal power in the remaining channels.

A second formulation of the Taylor processor was also considered. For this case the limiters for each channel were replaced by independent AGC's. The results discussed in Section 3.3 indicate that improvement in signal-to-noise can be achieved with both forms fo the Taylor processor in the presence of narrow band jamming.

3.0 SIMULATION RESULTS

During the second half of the contractual period emphasis was shifted from the characterization of jamming signals and the nomination of candidate nonlinear techniques, to the actual evaluation of those techniques. Attempts at analytical evaluation were stymied because of the varied jamming signal types considered and the resulting intractable mathematics. Therefore, a simulation was developed in several forms to test the several applicable nonlinear techniques in a Monte Carlo fashion.

The evaluation of each nonlinear technique was based upon the detection performance relative to conventional matched filter processing. Where computation times would allow, the measure of performance was the inferred probability of detection based upon detection counts from 100 trials. Levels of significance for false alarm probabilities could not be obtained by Monte Carlo techniques due to excessive requirements for numbers of trials. Therefore, threshold selections for false alarm probabilities were determined from theoretical distributions.

Where computation times did not allow direct estimation of detection probabilities, the measure of performance was taken to be an improvement in the signal plus interference-to-interference power ratio over a conventional matched filter. Symbolically:

$$\text{Improvement factor} = \frac{\left(\dfrac{<|Y(s+n)|^2>}{<|Y(n)|^2>}\right)}{\left(\dfrac{<|X(s+n)|^2>}{<|X(n)|^2>}\right)} \tag{3-1}$$

where $Y(\cdot)$ is the nonlinear operation, $X(\cdot)$ is the linear, matched filter operation, s is the signal, n is the jammer signal plus thermal noise, and $<\cdot>$ is the sample expectation (average) operator. In other words, the ratio of the average processor output when a signal is present to the average processor output with no signal is compared for the nonlinear process $Y(\cdot)$ and the matched filter process $X(\cdot)$. Note that

$$E\left[|X(s+n)|^2\right] = E\left[|X(s)|^2\right] + E\left[|X(n)|^2\right] \tag{3-2}$$

since X is linear and s and n are independent. However, the equivalent relationship for Y is not true. Thus, the determination of signal power alone out of such a nonlinear process is not so well defined. Consequently the ratio of signal plus interference-to-interference, and not signal-to-interference is used.

A common signal type was assumed for each technique evaluated. This was an up-chirped LFM waveform with a BT product of 256. Since digital simulation techniques were used to evaluate the candidate processors, the waveform was, in fact, a sampled LFM waveform. This sampling was performed at the waveform bandwidth and in I and Q, so that each observation consisted of a complex vector of length 256. The sampling rate and BT products were selected so that reasonable simulation run times could be achieved and yet a relatively high BT product waveform could be considered.

3.1 GENERALIZED MATCHED FILTER

An Adaptive Generalized Matched Filter (AGMF) was simulated in order to determine the performance characteristics of the GMF in a noise environment that was not known a priori. In Reference 1 the GMF was shown to be the linear processor that maximized signal-to-noise for known, stationary noise processes, and (preceeding a threshold detector) the implementation of the optimum detector for known, stationary Guassian noise processes. When the characteristics of the noise are not known, the implementation of the GMF is based upon an estimate of the noise characteristics. The AGMF includes an estimation procedure and its performance is dependent upon the ability to estimate the required parameters. These estimates in turn depend upon the length of the time those parameters are constant. Consequently the AGMF is not applicable to those type of jamming signals denoted as non-stationary in Section 2.1.

In Reference 1 it was shown that the frequency response of the generalized matched filter is given by:

$$H(\omega) \;=\; e^{j\omega\tau}\, S^*(\omega)/N(\omega)$$

where $N(\omega)$ is the power spectral density of the interference and $S(\omega)$ is the spectrum of the signal. Thus, in the presence of jamming, $N(\omega)$ is the sum of a thermal noise component (independent of frequency) plus a jamming signal term. If the jammer power spectral density is constant over the bandwidth of the signal (i.e., a broad band jammer), then the GMF reduces to the matched filter and no improvement is obtained.

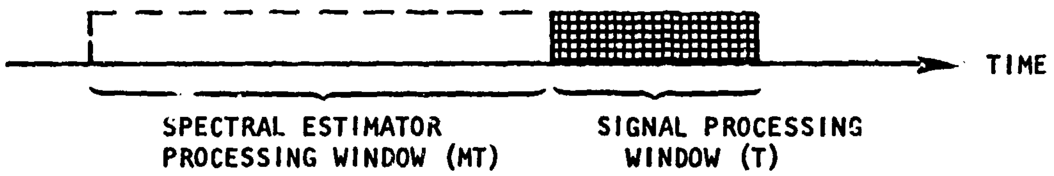
There are several techniques used to estimate power spectral density (Reference 4). The approach taken for this study was the averaging of periodograms as follows: The sampled observation under consideration can be represented as a segment of a sequence, say

$$\{y(k)\}_{k=p}^{p+255}$$

for some p. Then the Discrete Fourier Transform (DFT) of the observation may be denoted $F\left[\{y(k)\}_{k=p}^{k=p+255}\right]$. The periodogram is proportional to $\left|F\left[\{y(k)\}_{k=p}^{k=p+255}\right]\right|_{\ell}^{2}$. The estimated interference power spectral density, $\hat{N}$ (a vector), is determined by

$$\hat{N}(\ell) = \frac{1}{M}\sum_{j=1}^{M}\left\{\left|F\left[\{y(k)\}_{k=p-j\,255}^{p-(j-1)255}\right]\right|_{\ell}^{2}\right\} \qquad (3\text{-}3)$$

where ℓ is an index of the vector components. Thus, the M contiguous intervals of equal length to, and immediately preceding the interval to be processed, are used to compute the estimated power spectral density of the interference. This concept is illustrated in the following figure.



The tested AGMF performed the following operation:

$$D(p) = \sum_{\ell=1}^{256} \frac{F\left[\{y(k)\}_{k=p}^{p+255}\right]_{\ell} \; F\left[\{s(k)\}_{k=1}^{256}\right]^{*}_{\ell}}{\hat{N}(\ell)} \tag{3-4}$$

D is the output of the GMF and

$$y(k) = \begin{cases} s(k-p+1) + n(k) + r(k) & , \; p \le k \le p + 255 \\ n(k) + r(k) & , \; k < p \end{cases} \tag{3-5}$$

where n(k) is thermal noise, r(k) is the jamming signal, and s(k) is the sampled waveform.

Equation (3-4) is a frequency domain implementation of the GMF. The program flow is illustrated in Figure 3-1. Now $|D(p-255)|^2$ is the interference power in the processing window just prior to the signal and

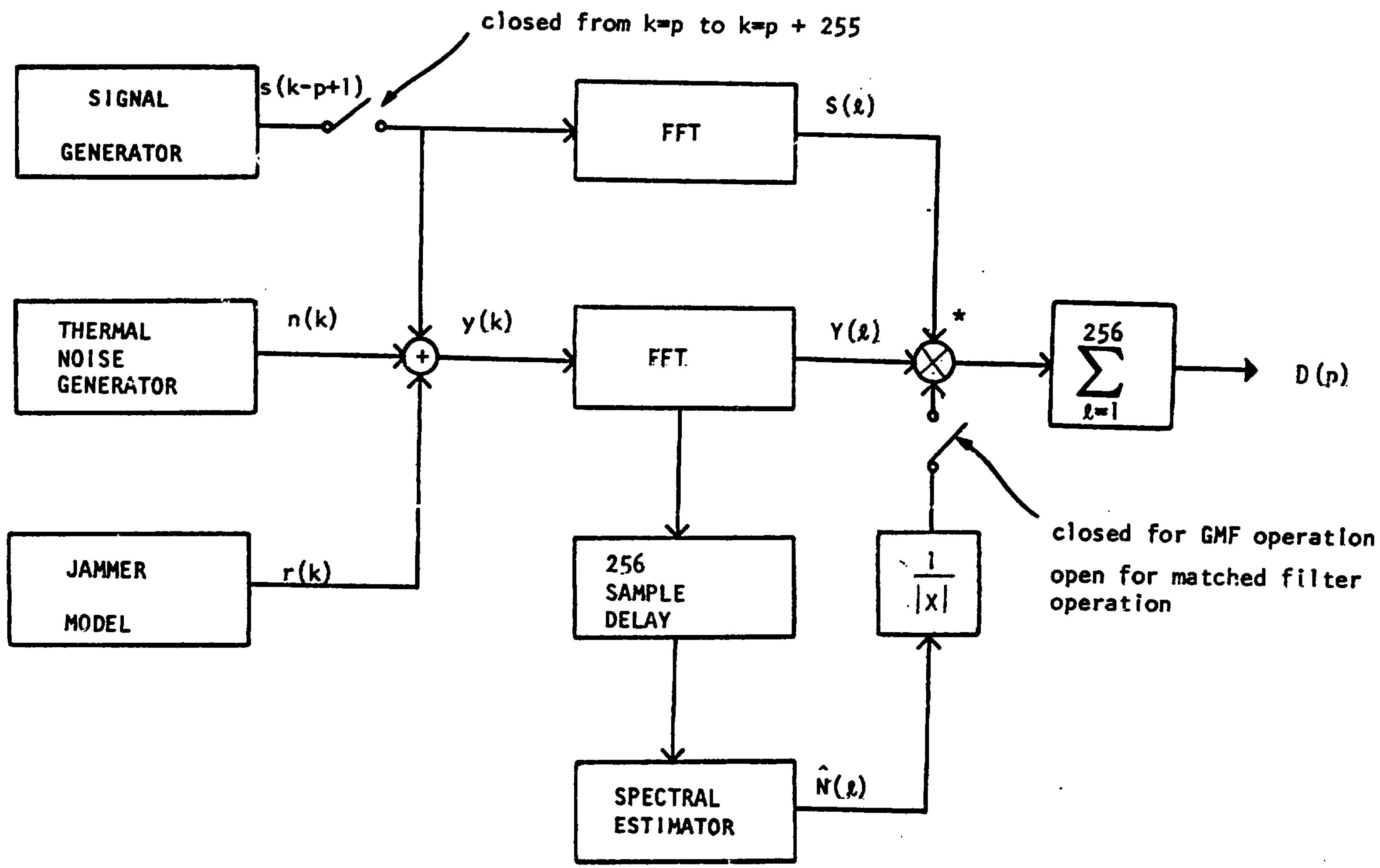


FIGURE 3-1. FLOW OF ADAPTIVE GENERALIZED MATCHED FILTER SIMULATION

$|D(p)|^2$ is the signal plus interference power. As indicated in Figure 3-1, the AGMF may be operated as a conventional matched filter (MF). In fact the simulation implements the AGMF and the MF simultaneously so that the improvement factor can be determined for a single set of inputs. Furthermore, based upon the interference power, $|D(p-256)|^2$ averaged over 100 cases, a detection threshold, C, is determined for a given false alarm probability, P_{fa}, by

$$C = \sqrt{-2\ I\ \ln\ (P_{fa})} \qquad (3\text{-}6)$$

where I is the average interference power. If $|D(p)|^2 > C$, a detection is declared. In order to estimate the probability of detection, the number of actual threshold crossings out of the 100 trials are counted.

Since the AGMF is only able to provide a benefit for jamming signals that are narrow band relative to the signal, this type of jamming signal was considered. The first case jamming signal was colored Guassian noise amplitude modulating a carrier, ω_0, at the center of the bandpass of the waveform. The noise was generated such that the autocorrelation function is of the form

$$R(u) = e^{-a|u|} \qquad (3\text{-}7)$$

and the power spectral density is of the form

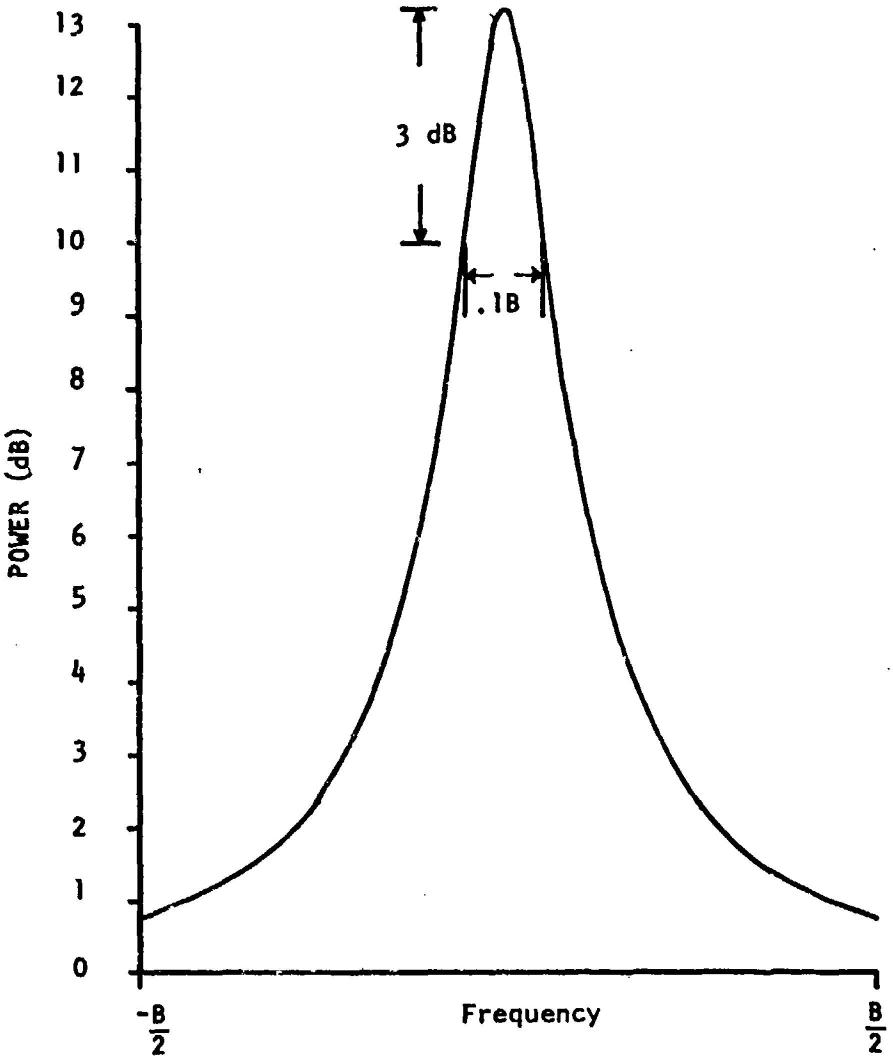
$$\overline{N}(\omega) = \frac{2a\left(\frac{J}{N}\right)}{(\omega - \omega_0)^2 + a^2} + 1 \qquad (3\text{-}8)$$

where 2a is the jammer bandwidth (3 dB) and $\left(\frac{J}{N}\right)$ is the jammer-to-noise power ratio. Figure 3-2 is a plot of the power spectral density for the case with $\frac{J}{N} = 10$ and $2a = 10\%$ of signal bandwidth, B.

It is of interest to know what the improvement factor is for the GMF under ideal conditions, that is, when the interference power spectral density is known. This is plotted in Figure 3-3 for the narrow band jamming signal described above. The improvement over the MF in signal plus interference-to-interference is plotted as a function of jammer bandwidth for a 25 dB jammer-to-noise ratio. For this example, the signal was assumed to have a rectangular power spectrum. As indicated by this figure, the improvement factor is not very pronounced except for relatively narrow-band jamming signals. The improvement factor drops below 1 dB as the jammer bandwidth exceeds 50% of the signal bandwidth.

Of course, the actual shape of the interference power spectral density determines the amount of improvement attainable. The shape of the power spectral density used here corresponds to simple RC-filtered thermal noise.

The performance of the AGMF for the interference model discussed above using 10% jamming signal bandwidth and 25 dB jammer-to-noise ratio



F. JRE 3-2. INTERFERENCE POWER SPECTRUM FOR 10 dB JAMMER-TO-NOISE RATIO, 10% JAMMER to SIGNAL BANDWIDTH.

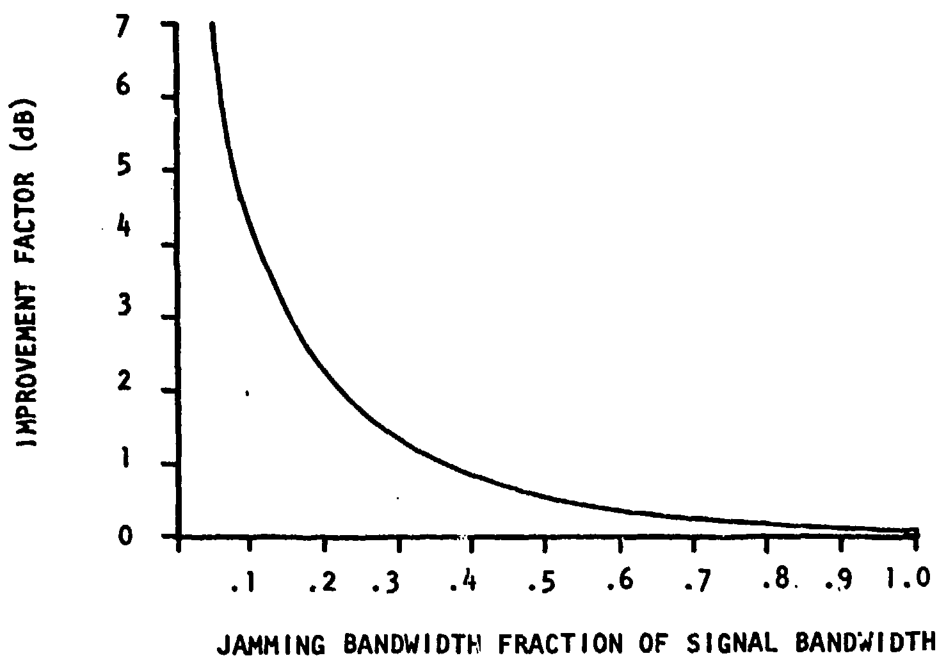


FIGURE 3-3. IMPROVEMENT FACTOR FOR IDEAL GMF

was examined for various spectral estimation window lengths. The plot of improvement factor (Monte Carlo estimation) versus length of spectral estimation window is shown in Figure 3-4. This figure shows that a spectral estimation window equal to the signal processing window resulted in almost a 3 dB loss in signal plus interference-to-interference power over the conventional matched filter. Even with relatively long estimation intervals, the performance of the AGMF does not equal the improvement expected for the ideal filter: 4.5 dB from Figure 3-3.

The length of the spectral estimation window determines the response time of the adaptive loop in the AGMF. Therefore, if the power spectral density of the jammer varies more rapidly than this length of time, the AGMF will perform poorly. On the other hand, a fast response time implies a short spectral estimation window which implies poor performance, as shown in Figure 3-4. With these considerations, a spectral estimation window five times longer than the signal processing window was assumed. This choice provides convergence of the interference power spectral density estimate for Stationary I and Stationary II type jamming signals. The AGMF will not provide a countermeasure for non-stationary type jamming signals.

The AGMF is of benefit only when the interfering source is narrow band relative to the signal of interest. In the event that wide band interference is encountered, it is appropriate to determining the level of degradation in performance of the AGMF compared to the appropriate MF. This performance degradation was found to be -1.0 and -1.2 dB for 6 dB and

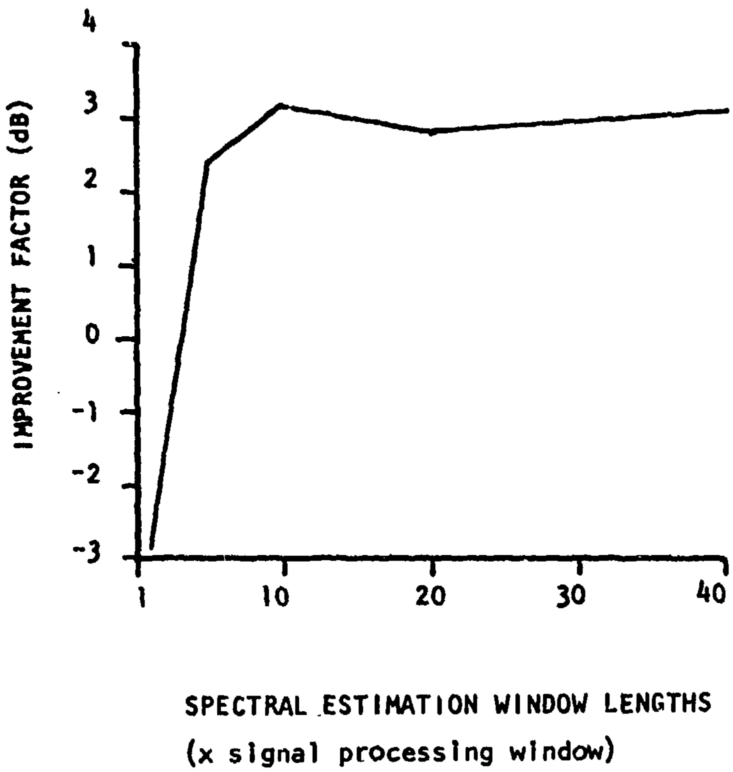


FIGURE 3-4. IMPROVEMENT FACTOR VERSUS ESTIMATION TIME FOR 10% JAMMING BANDWIDTH, 25 dB JAMMER-TO-NOISE RATIO

12 dB S/N ratio, respectively. By invoking the AGMF only when narrow band interference is sensed, it may be possible to eliminate this degradation.

The detection performance of the AGMF was examined via Monte Carlo techniques. One case is presented in Figure 3-5. This figure shows the number of detections out of 100 trials (estimated P_d) for the AGMF and the MF as a function of false alarm probability. The input signal-to-noise for this case is 10 dB, the input jammer-to-noise is 25 dB, and the jammer bandwidth is 10% of the signal bandwidth. The signal processor gain is 24 dB so that this example corresponds to a high signal-to-noise environment but a moderate (about 9 dB post processor) signal-to-jammer situation. The performance of the AGMF is quite superior in this example. At 10^{-5} P_{fa}, the estimated P_d for the MF is about 0.125 while the AGMF estimated P_d is near 0.87. The improvement factor for this case was 3.7 dB.

A second example is given by Figure 3-6. In this case a 0 dB input signal-to-noise ratio and a 18 dB input jammer-to-noise ratio were considered. Again, 10% jammer bandwidth was assumed. The detection performance for this case was considerably worse because the post processing signal-to-jamming ratio is only about 6 dB. Marked improvement was achieved by the AGMF. The improvement factor, for this case, is 2.9 dB.

The AGMF was also tested against constant amplitude jamming with phase noise. This type of signal, denoted Category III in Section 2-1,

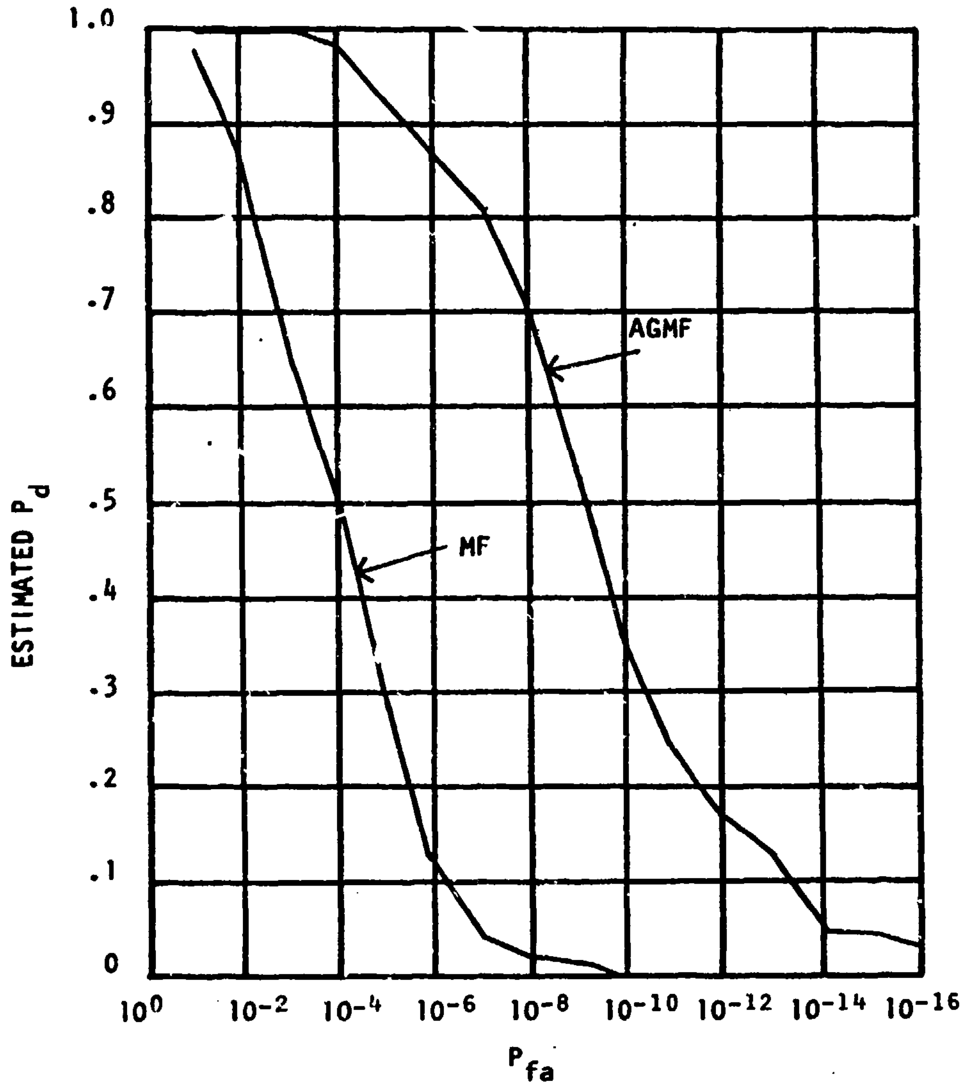


FIGURE 3-5. DETECTION PERFORMANCE OF AGMF FOR 10 DB S/N, 25 dB J/N, AND 10% JAMMER BANDWIDTH

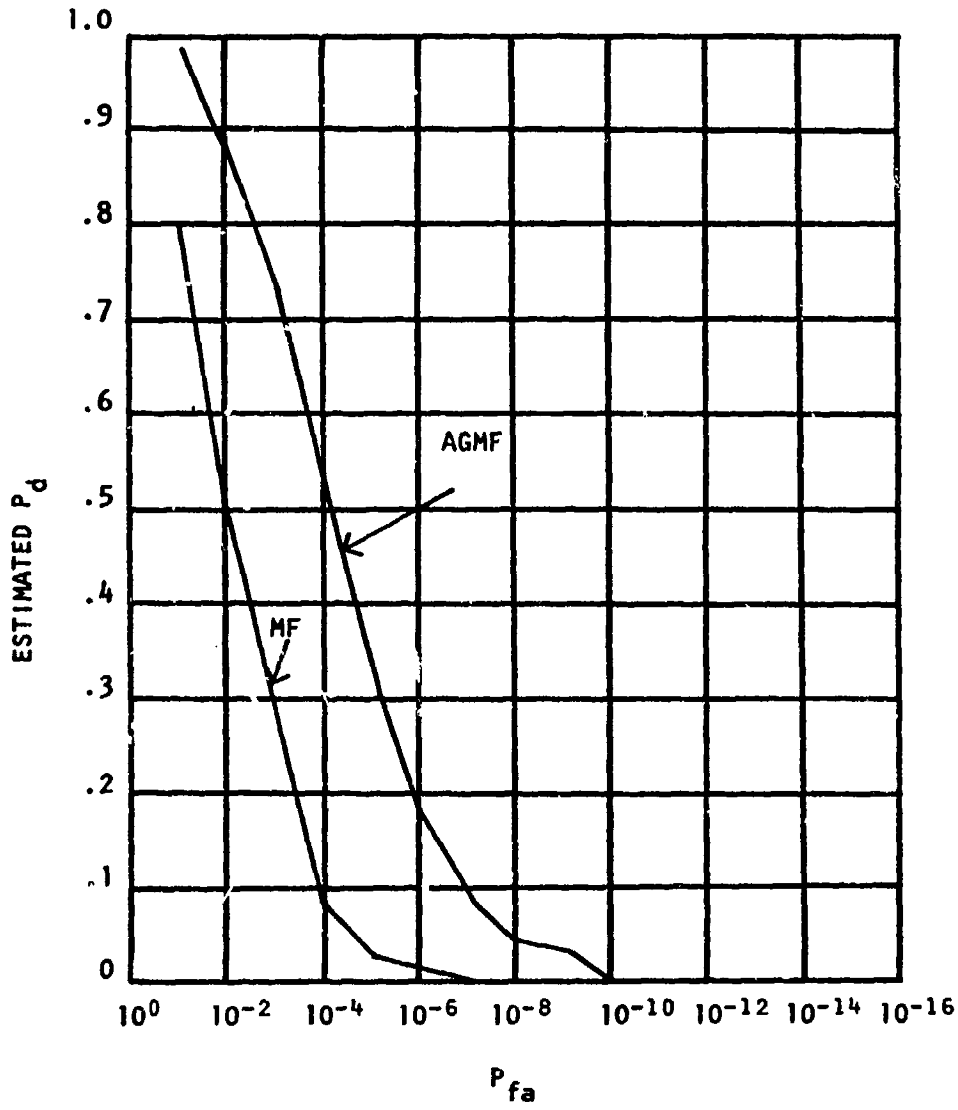


FIGURE 3-6. DETECTION PERFORMANCE OF AGMF for 0 dB S/N, 18 dB J/N, and 10% JAMMER BANDWIDTH

is considered to be consistent with efficient jamming power generation. Constant amplitude jamming signals were simulated by generating phase noise with a power spectral density of the form

$$\overline{N}_{\phi}(\omega) = \frac{2\,a}{\omega^2 + a^2} \tag{3-9}$$

where $\phi(k)$ represents the phase noise sequence. Then the constant amplitude noise sequence was formed from

$$r(k) = e^{j\phi(k)} \quad . \tag{3-10}$$

The AGMF demonstrated improvement over the MF for constant amplitude noise as well. The following table illustrates these improvement levels.

TABLE 3-1. AGMF PERFORMANCE FOR CATEGORY III (CONSTANT AMPLITUDE) JAMMING

JAMMER SIGNAL PERCENTAGE BANDWIDTH (APPROXIMATE)	IMPROVEMENT FACTOR (dB)
3%	6.0
7%	3.6
8.5%	2.9

NOTE: 0 dB input S/N; 18 dB input J/N

The results of this section have shown that the AGMF can provide superior performance to an MF in the presence of narrow band jamming signals with estimatable power spectral densities. It appears that lack of perfect knowledge of the power spectral density of the interference results in a 1 to 2 dB loss in S/N over the ideal GMF. Furthermore, the jamming signal bandwidth was required to be 10% or less of the signal bandwidth in order to achieve improvement factors greater than 3 dB for the cases considered.

3.2 NON-COHERENT CLIPPING/BLANKING

Non-coherent clipping and non-coherent blanking are two nonlinear digital signal processing techniques that are designed to exploit the fact that the jamming signals may be localized in either time or frequency, when compared to the signal of interest. Such a situation can occur in either pulsed jamming or narrow band jamming. Unlike the AGMF, discussed in Section 3.1, these two techniques do not require that any prior estimation of the jamming signal characteristics (with the possible exception of power). This opens the possibility of providing improved performance over the linear (matched) filter for non-stationary signals.

As mentioned in Section 2.2.3, the initial approach to evaluating the potential benefit of non-coherent clipping was to analytically derive the appropriate threshold vector which would maximize the probability of improving the signal-to-noise energy ratio prior to matched filtering. However, this approach resulted in unmanageable mathematical complexity, and hence a simulation approach was developed. The simulation approach included treating the threshold selection parametrically.

Figure 3-7 is a block diagram of the simulation. The two nonlinear elements are the time domain clipper/blanker and the frequency domain clipper/blanker. The time domain element is provided to reduce the energy of jamming signals that are localized in time. The frequency domain element is provided to accomplish the same result for jamming signals that are localized in frequency. It was discovered that having these elements in series in this manner caused adverse effects. The time domain nonlinearity tended to broaden the spectrum of narrow band jamming signals so that the frequency domain clipper/blanker could not perform as desired. Consequently, those two nonlinear elements were examined independently, which would correspond to a parallel implementation and dual detection logic, as shown in Figure 3-8.

The clipping and blanking threshold vectors were assumed to have equal components. This is consistent with the fact that the signal vector (as described in Section 3.0) in both time and frequency domains have components of equal amplitude. The threshold level was varied as a percentage of the measured rms noise level.

The frequency domain clipper/blanker was exercised against narrow band jamming signals including those used for evaluating the AGMF. The first case considered was for narrow band Guassian jamming with power

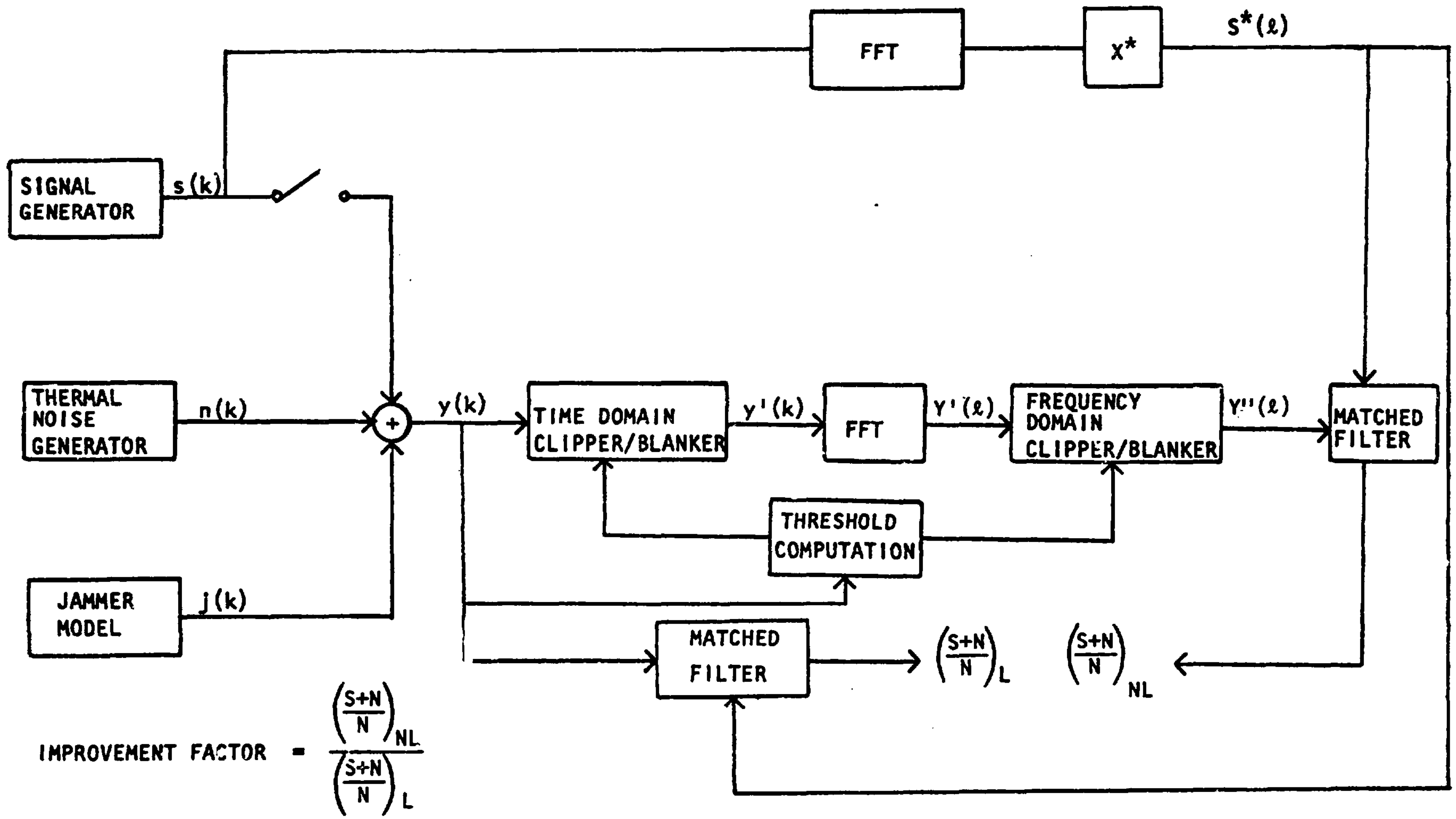


FIGURE 3-7. NON-COHERENT CLIPPING/BLANKING SIMULATION

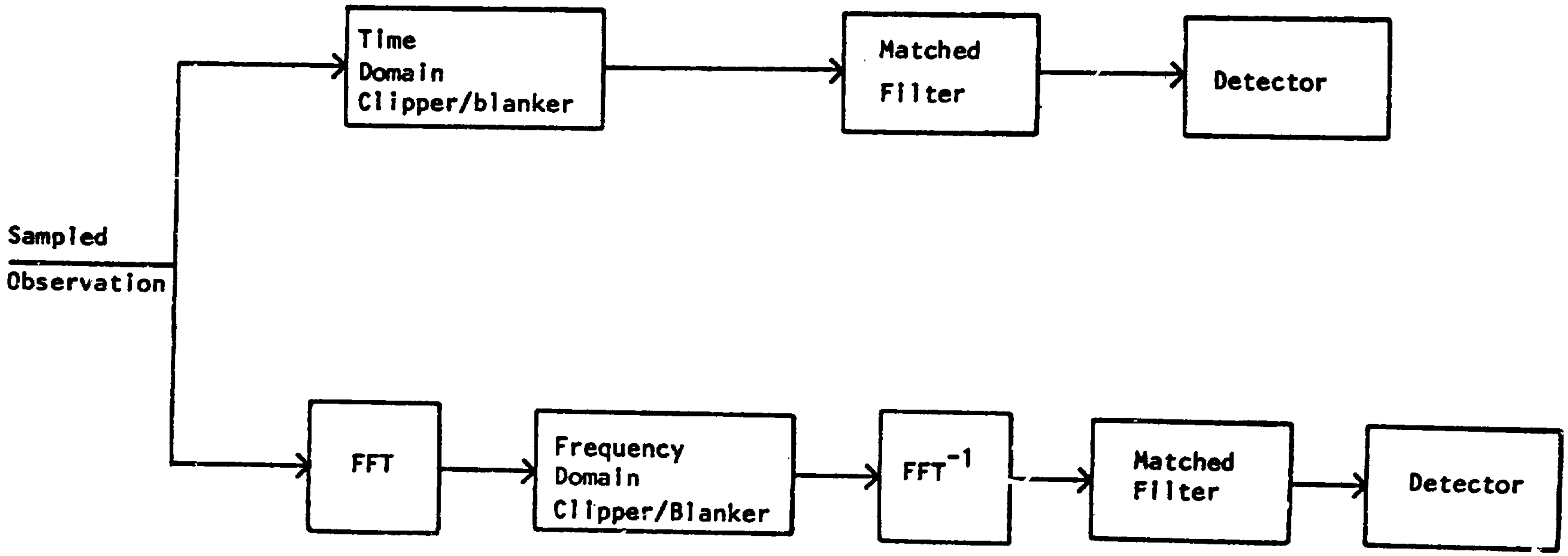


FIGURE 3-8. PARALLEL IMPLEMENTATION OF NON-COHERENT CLIPPING/BLANKING

spectral densities of the form given by Equation 3-8. The case of 10% (half-power) jammer bandwidth relative to signal bandwidth and 18 dB J/N ratio and 0 dB S/N ratio (input to processor) was examined. The improvement factors for non-coherent clipping as a function of threshold level is shown in Figure 3-9. The improvement factor for non-coherent blanking is also shown in this figure. These results indicate the maximum improvement for non-coherent clipping occurs when the threshold is set below the rms interference level. However, the blanking technique works best at a higher threshold setting. This result is not unexpected. The clipped observation vector preserves the phase content of the signal, even when all observation components are affected, such as the case when the clipping threshold is zero (i.e., hard limiting). On the other hand, the observation components whose magnitude exceeds the blanking threshold are set to zero and consequently all signal information is lost.

Figure 3-8 indicates that the optimum threshold selection for the clipping technique is about 0.4 times the RMS interference level. This result seemed to hold for a wide variety of jamming conditions and was accepted as fixed threshold selection strategy for the frequency domain clipper. Likewise, the 1.2 times RMS interference level was found to be near optimum for a wide variety of conditions and it too was fixed as the frequency domain blanking threshold.

Using these threshold selections, the improvement factor for clipping is 2.9 dB and for blanking is 2.7 dB. These values compare favorably with the 2.9 dB improvement attained under these conditions by the AGMF.

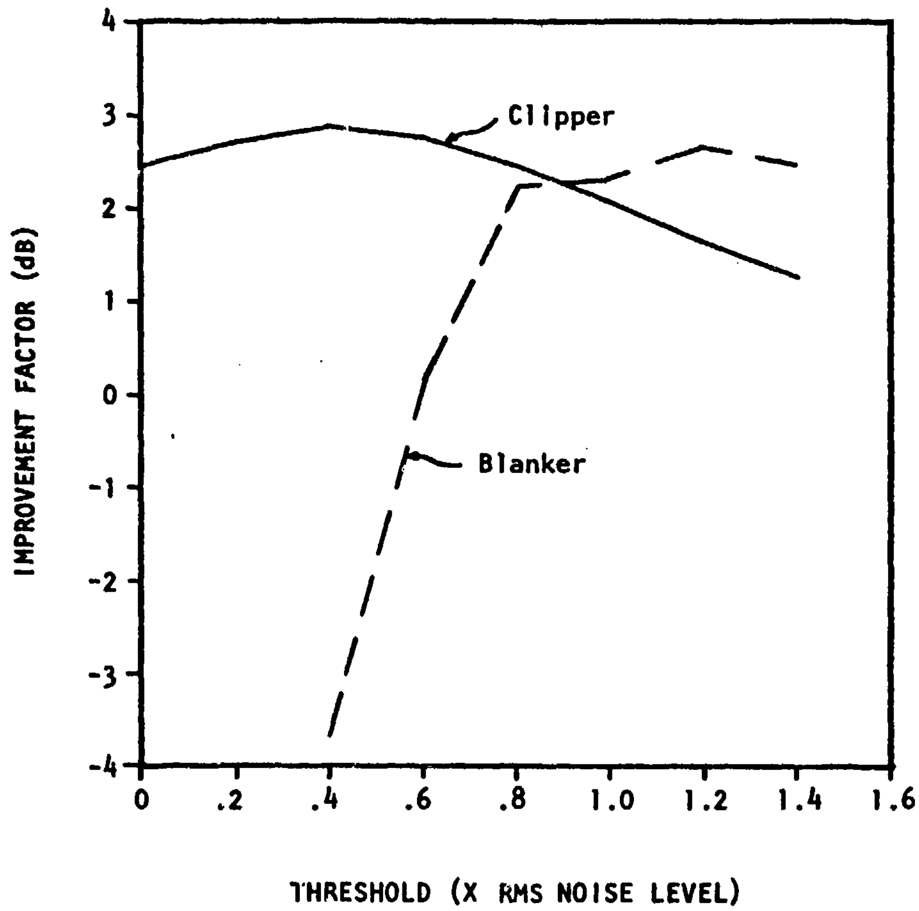


FIGURE 3-9. IMPROVEMENT FACTORS FOR THE NON-COHERENT CLIPPER/BLANKER FOR 0 dB S/N, 18 dB J/N, AND 10% JAMMER BANDWIDTH

Figure 3-10 shows the detection performance curves for the clipper and the blanker. The matched filter results are also shown for comparison. Although, single pulse detection performance is probably not acceptable under these conditions even with the improvement of these nonlinear techniques, the number of pulses required to achieve reasonable performance would be considerably reduced by the use of either the non-coherent clipper or blanker.

Figure 3-11 shows the detection performance for the case of narrow band Gaussian noise where the jammer-to-noise ratio is 15 dB and the jamming signal bandwidth is 10% of the signal bandwidth. The results are similar to the 18 dB jammer-to-noise case. Marked improvement in detection performance is achievable with the non-coherent clipper and the blanker when compared to the conventional matched filter. In this example, the performance of the matched filter is unacceptable while the performance of either the non-coherent clipper or blanker may be acceptable. The improvement factors for this case are 2.9 dB and 2.6 dB for the clipper and blanker respectively.

It has been demonstrated that these nonlinear techniques can provide an improvement in detection performance for narrow band Gaussian envelope jamming signals. However, this improvement depends upon the particular bandwidth of the jammer, as shown in Figure 3-12. This figure provides the improvement factor for non-coherent clipping and blanking as the jammer bandwidth varies from 10 to 25% of the signal bandwidth. This figure

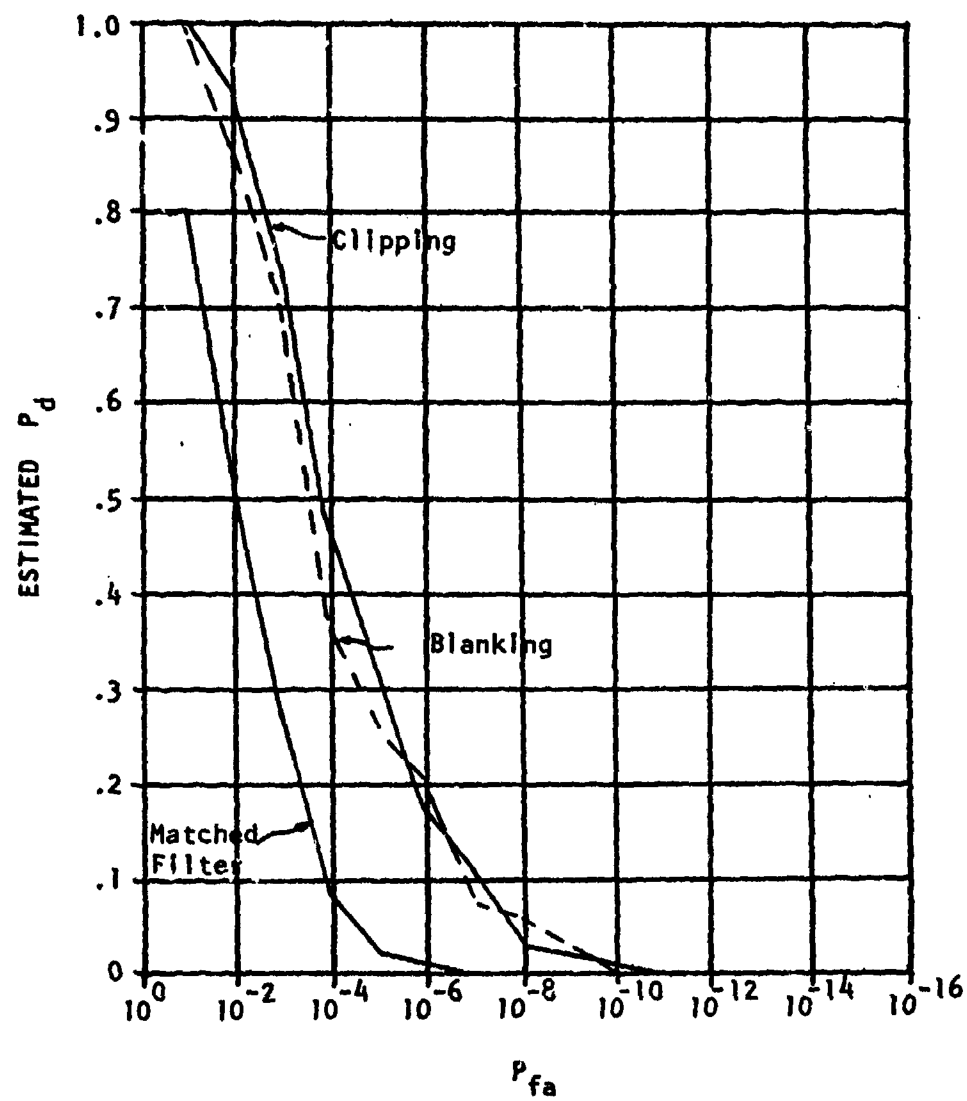


FIGURE 3-10. DETECTION PERFORMANCE FOR NON-COHERENT CLIPPING/BLANKING FOR 0 dB S/N, 18 dB J/N, AND 10% JAMMER BANDWIDTH

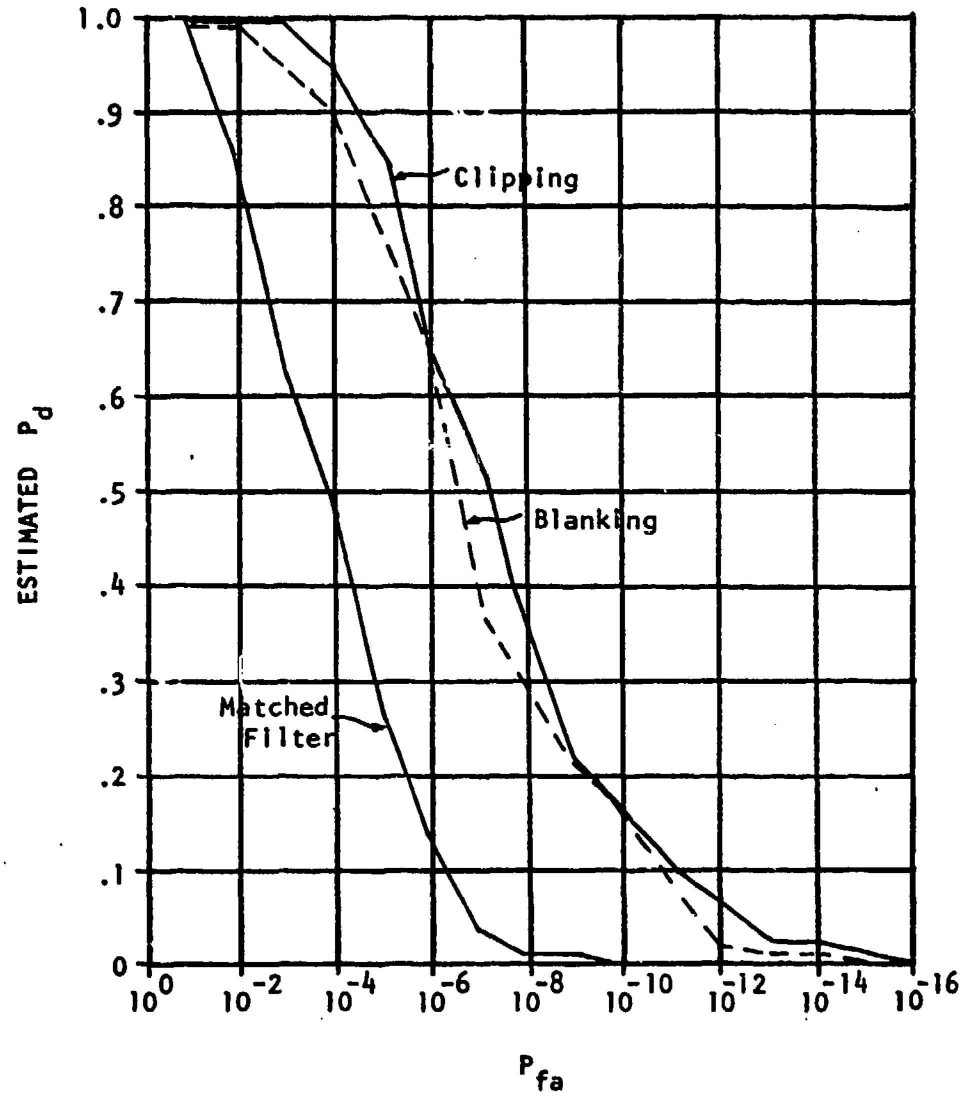


FIGURE 3-11. DETECTION PERFORMANCE FOR NON-COHERENT CLIPPING/BLANKING FOR 0 dB S/N, 15 dB J/N, AND 10% JAMMER BANDWIDTH

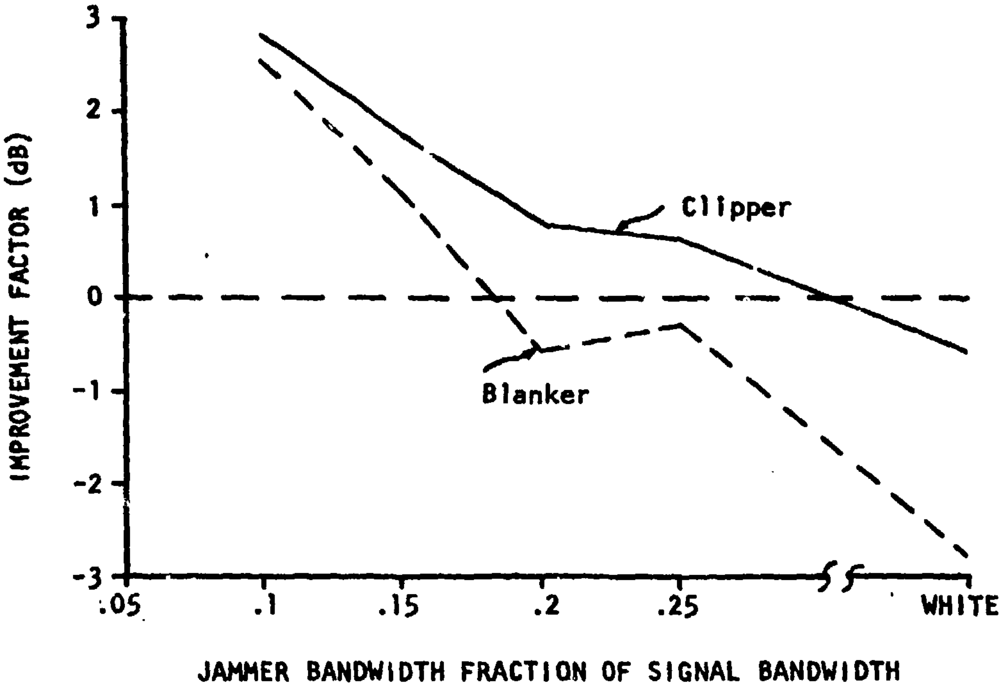


FIGURE 3-12. NON-COHERENT CLIPPER/BLANKER PERFORMANCE AS A FUNCTION OF JAMMER BANDWIDTH, 0 dB S/N and 15 dB J/N

shows the rapid fall off in improvement as the jammer bandwidth broadens. At 20% bandwidth the blanker results in about ½ dB loss in signal plus interference-to-interference ratio. At that point the clipper demonstrates only about 0.8 dB improvement.

Also shown in Figure 3-12 are the results for wide band (white) jamming. This result indicated that the blanker performance deteriorates unacceptably (-2.75 dB improvement factor) when wide band interference (including thermal noise alone) is dominant. On the other hand, the clipper only suffers about ½ dB loss in improvement factor. This is favorable when compared to the 1 to 2 dB loss in S/N demonstrated by the AGMF under similar conditions.

The large negative improvement factor demonstrated by the non-coherent blanker in wide band noise is not unexpected. Since the blanking threshold is proportional to the RMS noise level, as the interference bandwidth broadens, more components of the observation vector exceed the threshold. This effect is illustrated in Figure 3-13. In the presence of wide band jamming under small signal conditions, the probability of an observation component exceeding the blanking threshold (1.2 x RMS noise level) is approximately 0.49. Thus, about half of the observation components would be blanked and the compression gain of the signal processor would be reduced by about 3 dB. This difficulty with the non-coherent blanker can be alleviated by invoking an upper limit on the number of blanked components. An alternate procedure would be to blank the K largest (magnitude) components of the observation vector for some fixed number, K.

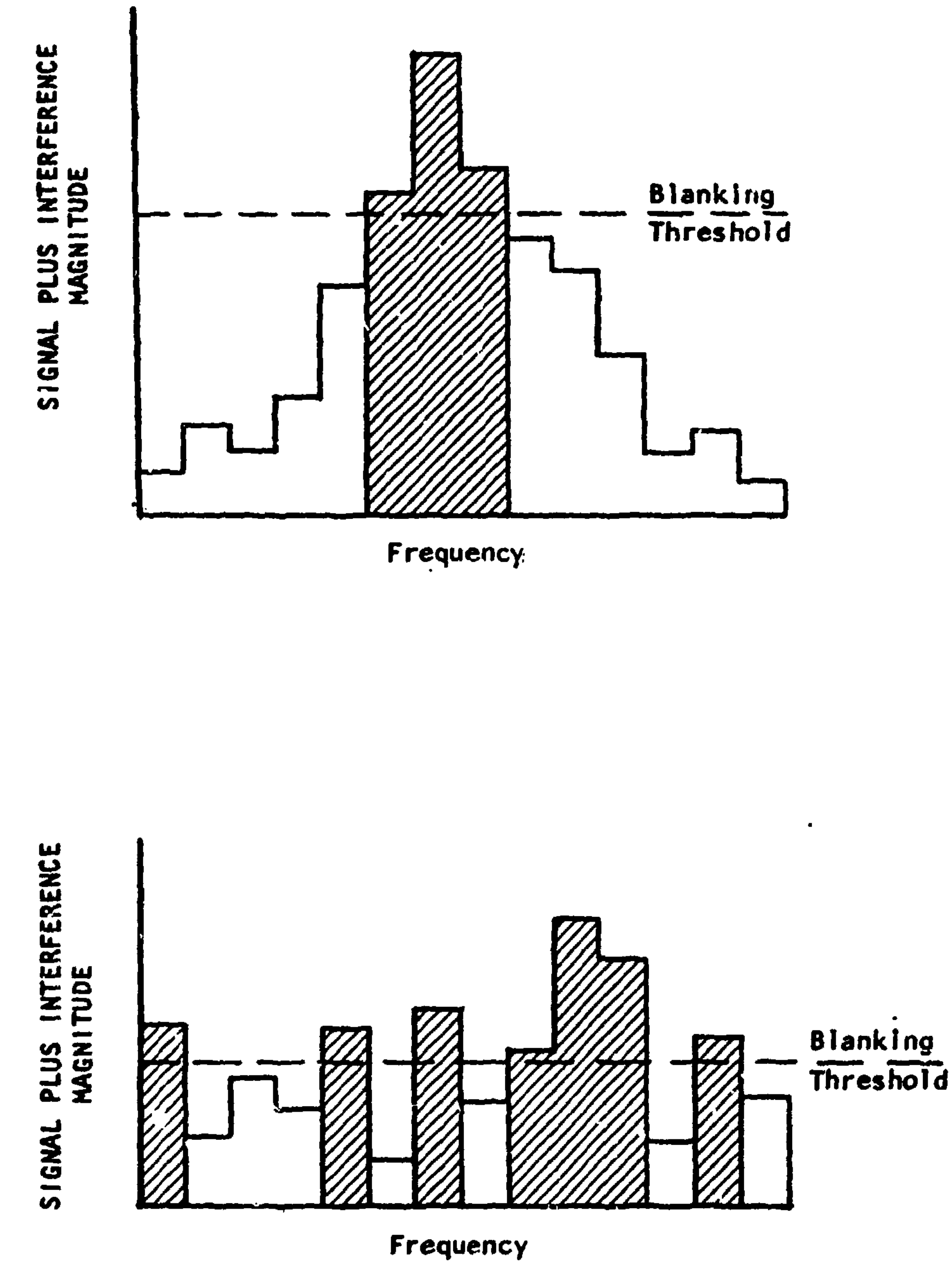


FIGURE 3-13. OPERATION OF NON-COHERENT FREQUENCY BLANKER IN NARROW BAND AND WIDE BAND INTERFERENCE

The performance of the non-coherent frequency clipper and blanker was also examined for the case of narrow band, constant amplitude jamming signal types. The case of 0 dB S/N with 18 dB J/N was considered for three jammer bandwidths. The following table summarizes these results.

TABLE 3-2. NON-COHERENT FREQUENCY CLIPPING/ BLANKING PERFORMANCE FOR CONSTANT AMPLITUDE JAMMING

JAMMER SIGNAL PERCENTAGE BANDWIDTH (APPROXIMATE)	IMPROVEMENT FACTOR (dB)	
	CLIPPING	BLANKING
3%	6.2	6.4
7%	3.4	3.0
8.5%	1.4	0.6

Comparing the above table with Table 3-1, which gives the analogous performance for the AGMF, it is apparent that the clipper and blanker perform slightly better than the AGMF for the 3% bandwidth case. At 7% bandwidth the AGMF is slightly better, and at 8.5% bandwidth, the AGMF, with a 2.9 dB improvement factor, performs significantly better than the non-coherent clipper or blanker techniques, at 1.4 and 0.6 dB respectively.

The alternate non-coherent clipping and blanking techniques operate in the time domain. These techniques were evaluated in the presence of pulsed jamming, for which they are particularly applicable. The basic attraction of these techniques is the fact that when a jamming signal occupies only a portion of the time window containing a signal, that portion can be modified so as to reduce the jamming energy subsequently processed by the matched filter.

Unlike the narrow band jamming cases considered for the frequency domain clipper/blanker, the pulsed jamming signals considered here are, in fact, zero for portions of the signal processing window. Consequently the clipping/blanking thresholds which provide the greatest improvement are not the same as were determined in the frequency domain case. This is illustrated by Figure 3-14, which shows the improvement factor for time domain clipping with 0 dB input signal-to-noise, 18 dB input jammer-to-noise (average jammer power) for a wide-band jammer with a Gaussian distributed random envelope. The threshold selection ranges from 0 (hard limiting) to 1.2 times the RMS interference level. Four cases of jammer duty cycles are considered, 12.5%, 25%, 50%, and 100% (continuous). With about 24 dB signal processor gain, the output S/I with matched filter processing for these examples is about 6 dB. This is not sufficient for reasonable detection performance. However, with the improvement factors indicated in the plot, satisfactory detection performance is possible, even with 50% duty cycle jamming, assuming high signal-to-thermal noise ratios.

For the cases considered in Figure 3-14, the optimum threshold level appears to be in the neighborhood of 0.2 times the rms interference level.

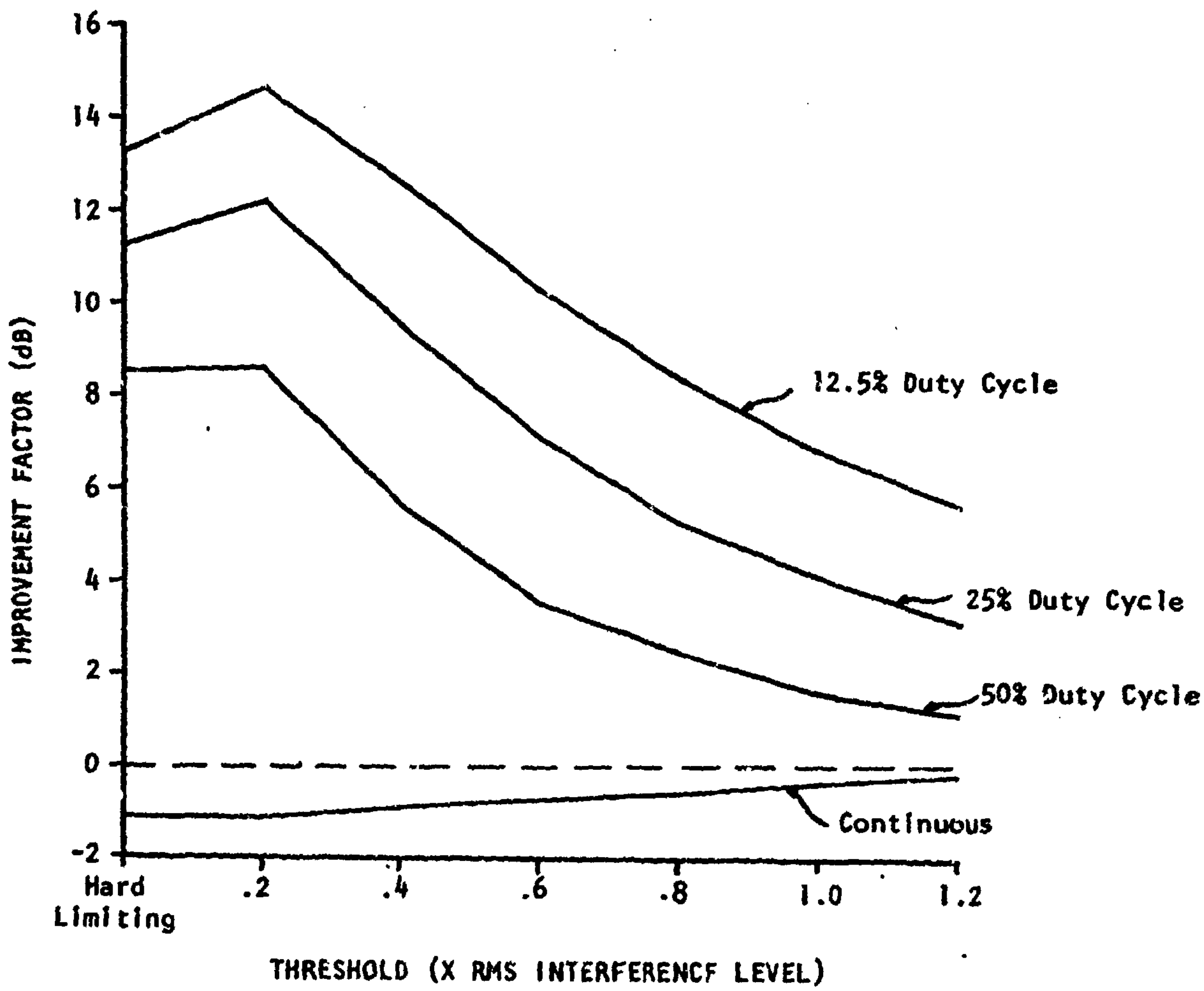


FIGURE 3-14. NON-COHERENT CLIPPING IMPROVEMENT FACTOR FOR WIDE BAND PULSED GAUSSIAN JAMMING, 0 dB S/N, 18 dB J/N

At this clipping threshold, the loss in S/I in the presence of continuous jamming is about -1 dB, as shown in the figure.

Figure 3-15 is a plot of the improvement factor for non-coherent blanking for the same cases as Figure 3-14. This figure shows that, at the optimum threshold level of 0.4 times the RMS interference level, the blanker performance is superior to the clipper performance. Up to 12.5 dB improvement was obtained for the 50% duty cycle case. This may be compared to the theoretical improvement factor of 14 dB which would result if the signal processing window were ideally blanked during the jamming pulse.

Also shown in Figure 3-15 is the loss associated with non-coherent blanking. At the optimum threshold, this loss is an unacceptable -6.5 dB. However, as mentioned in regard to frequency domain non-coherent blanking, part of that loss may be avoided by limiting the number of observation components that are blanked.

Figure 3-16 is a plot of the detection performance for the time domain non-coherent clipper/blanker in the presence of pulsed, wide band jamming with 50% duty dycle. This is for the case of 0 dB input S/N and 18 dB input J/N. The matched filter post processing signal-to-interference in this example is about 6 dB. This example represents a case where high S/N values are attainable yet jamming power is sufficient to substantially interfere with detection. In fact, out of 100 trials, Figure 3-16 shows that not a single detection resulted with matched filter processing and

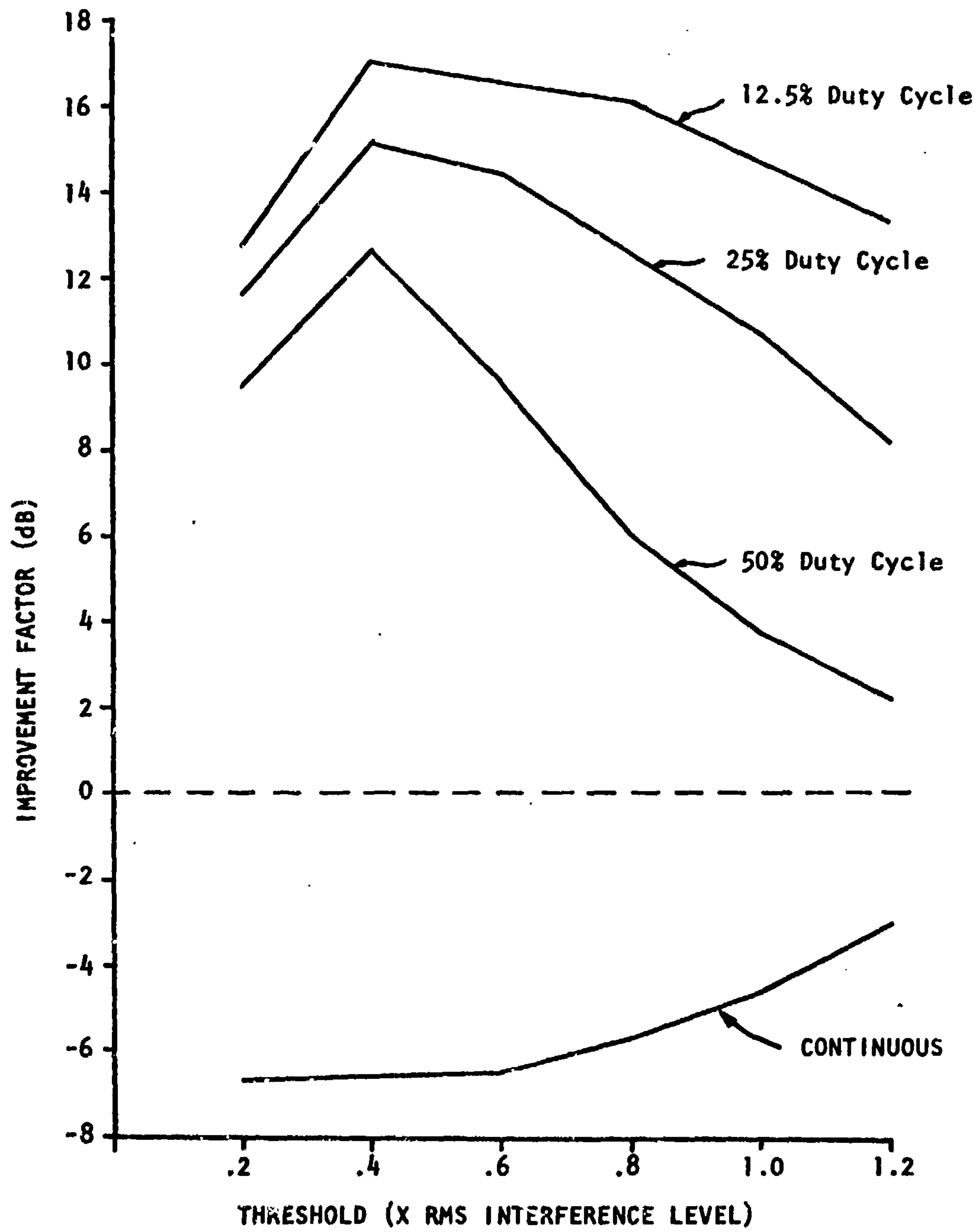


FIGURE 3-15. NON-COHERENT BLANKING IMPROVEMENT FACTOR FOR WIDE BAND PULSED GAUSSIAN JAMMING, 0dB S/N, 18 dB J/N

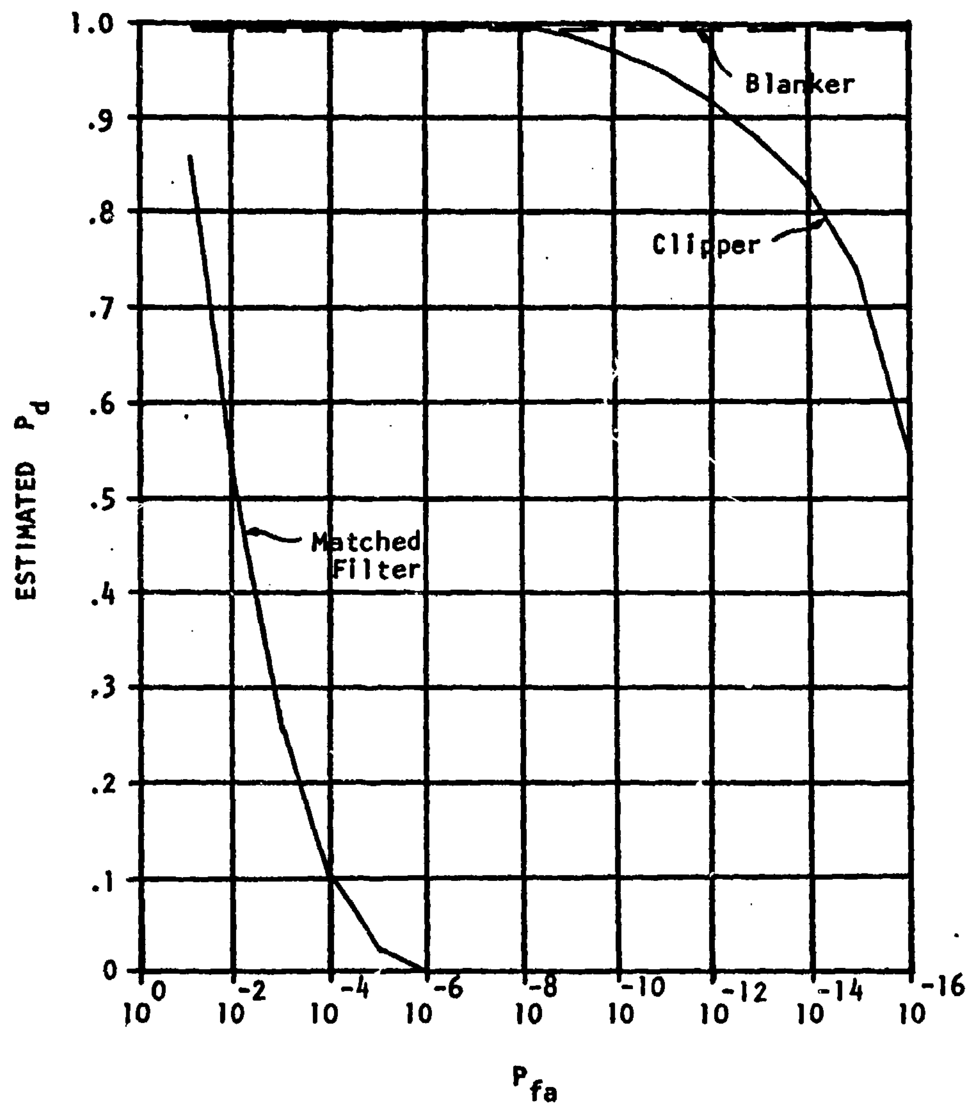


FIGURE 3-16. DETECTION PERFORMANCE FOR NON-COHERENT CLIPPING/BLANKING IN 50% DUTY CYCLE WIDE BAND JAMMING, 0 dB S/N, 18 dB J/N

with the detection threshold adjusted for 10^{-6} probability of false alarm. However, with non-coherent clipping or blanking, at 10^{-6} probability of false alarm, 100% success rate was achieved for detection. Because of the high S/N situation, extremely good detection performance is shown when the effects of the pulsed jamming are mitigated by the use of non-coherent clipping or blanking.

The results demonstrated in this section have shown that the non-coherent clipping and/or blanking techniques can provide superior performance to the conventional matched filter processor. These techniques are robust in the sense that they do not require stationarity for the jamming signals. However, they do require that the jamming signals be either localized in time or frequency when compared to the signal of interest. Thus, some knowledge of the offense jamming strategy would be desirable. A potential side benefit is the immunity gained from friendly unintentional interference sources.

3.3 FILTER-NONLINEARITY-FILTER PROCESSING

As discussed in Section 2.2.6, the FNF processor is designed to counter narrow band jamming by independently processing several channels or sub-bands of the band of the signal of interest. For the Taylor FNF processor, the theory is that the sub-band(s) containing the jamming signal can be limited. Hence the interference power would be reduced when the channels are recombined. The Taylor processor is in a sense an analogue to the non-coherent frequency clipping for non-digital processing. The testing of the Taylor processor was, however, accomplished via digital simulation.

The particular Taylor processor implemented for testing contained 8 channels. These channels were formed by inputting the observation through a bank of 8 contiguous Chebychev bandpass filters. This process is depicted in Figure 3-17. The filters were designed for peak sidelobes of -30 dB relative to the mainlobe, and were spaced so as to cover the entire band of the signal. Following the channel separation filters was a soft limiter (clipper) which performed the following transformation

$$f(X_k) = \begin{cases} X_k & \text{if} \quad |X_k| < T \\ \dfrac{TX_k}{|X_k|} & \text{if} \quad |X_k| \geq T \end{cases}$$

where T is a threshold and X_k is an observation component (complex for I and Q processing).

In addition to the above nonlinearity, two additional types were considered: an automatic gain control (AGC) based upon the peak signal in the particular channel during the processing window (signal duration), and an AGC based on the RMS voltage level for the particular channel.

Following the nonlinear transformation, each channel is filtered again using an identical filter to the appropriate channel separation filter. This second filter bank is desirable because the limiting nonlinearity has the tendancy to broaden the spectrum in each channel. Following the second filter bank the channels are recombined and processed by a matched filter. The appropriate matched filter is not matched to the transmitted signal,

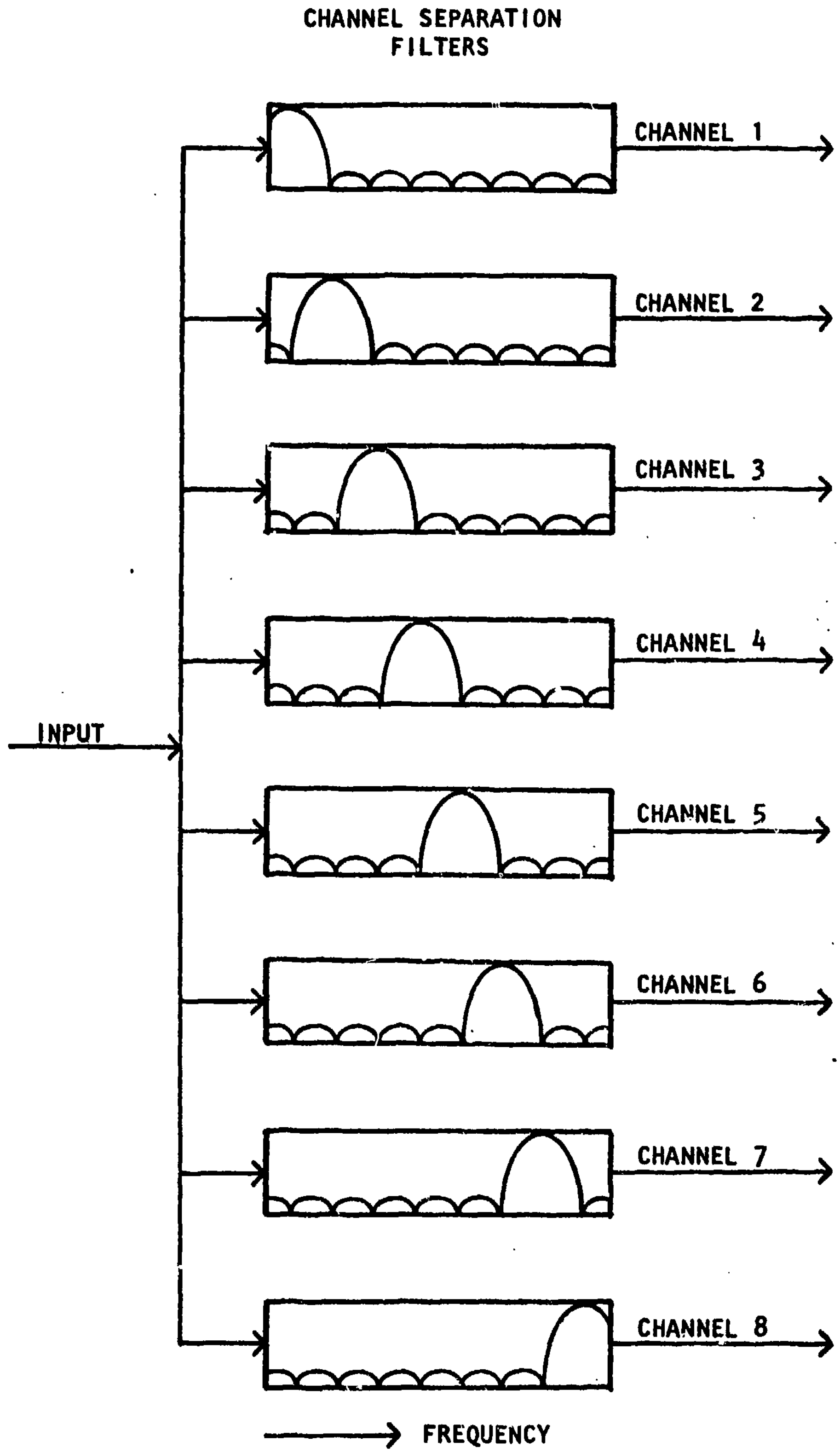


FIGURE 3-17. CHANNEL SEPARATION FOR 8 CHANNEL TAYLOR PROCESSING

but matched to that signal after passing through the filter banks of the Taylor processor. Figure 3-18 diagrams the simulation used to study the Taylor processor and the two AGC modifications. For the Taylor processor, the nonlinearities, $f_1(x)$ through $f_8(x)$ are all identical. But for the peak and RMS AGC implementations these functions are channel dependent. As shown in the figure, the appropriate matched filter has frequency response, $G(\omega)$ given by

$$G(\omega) = \left\{ \sum_{i=1}^{8} [H_i(\omega)]^2 \, S(\omega) \right\}^* \tag{3-11}$$

where $S(\omega)$ is the signal spectrum and $H_i(\omega)$ is the response of the i^{th} channel separation filter. For simulation purposes the signal was assumed to be the 256 BT product LFM waveform as described at the beginning of Section 3.0 and as used to evaluate the AGMF and non-coherent clipping/blanking techniques.

The Taylor processor was tested using narrow band Gaussian jamming signals as described by Equation 3-8. Because of the long run times required by the simulated signal processor, the number of Monte Carlo trials that could be reasonably performed were insufficient to allow for direct estimates of detection performance. Hence, the signal plus interference-to-interference ratio improvement over conventional matched filter processing is the standard by which the Taylor processor and the two AGC techniques were evaluated.

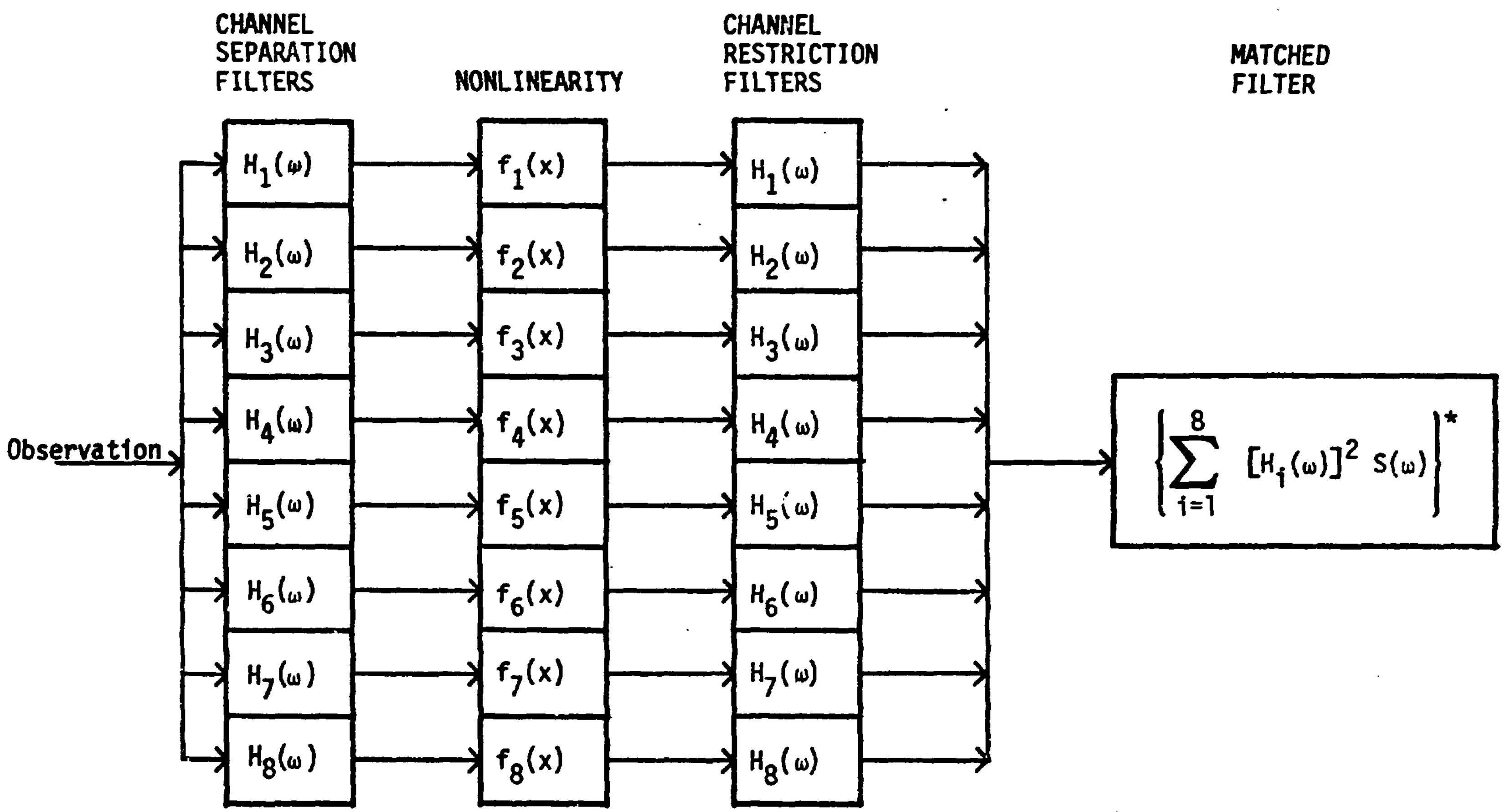


FIGURE 3-18. DIAGRAM OF TAYLOR PROCESSING SIMULATION

The clipping threshold for the Taylor processor was varied parametrically to determine the sensitivity in performance as well as the optimum levels for processing. Figure 3-19 indicates the threshold sensitivity for a jamming signal with 5% bandwidth (relative to signal bandwidth) for various levels of jamming power. The thermal noise power in this and subsequent examples has been set low enough so that the improvement factor is effectively improvement in signal-to-jamming ratio. The jamming power levels are indicated on the figure in terms of input signal-to-jamming (S/J) power.

For all but the -5 dB S/J case, threshold values below -10 dB effectively result in hard limiting so that the improvement factor is relatively insensitive to threshold. Similarly, for all but the -25 dB S/J case, the threshold level reaches a point where it is not exceeded and the performance levels off to a slight or negligible loss due to the filter banks. The region of maximum improvement encompasses a threshold range of about 4 dB for the -5, -10, and -15 dB S/J cases. At -20 and -25 dB S/J, the maximum achievable improvement level basically occurs at hard limiting. The greatest improvement factor occurs at -25 dB S/J where a 3.3 dB improvement in S/I is attained over the conventional matched filter. This results in an increase in S/I from about -1 to 2.3 dB after processing, still insufficient for reasonable detection performance.

The improvement factor attained at the optimum threshold was examined for several cases of jamming signal bandwidth. Figure 3-20 plots these

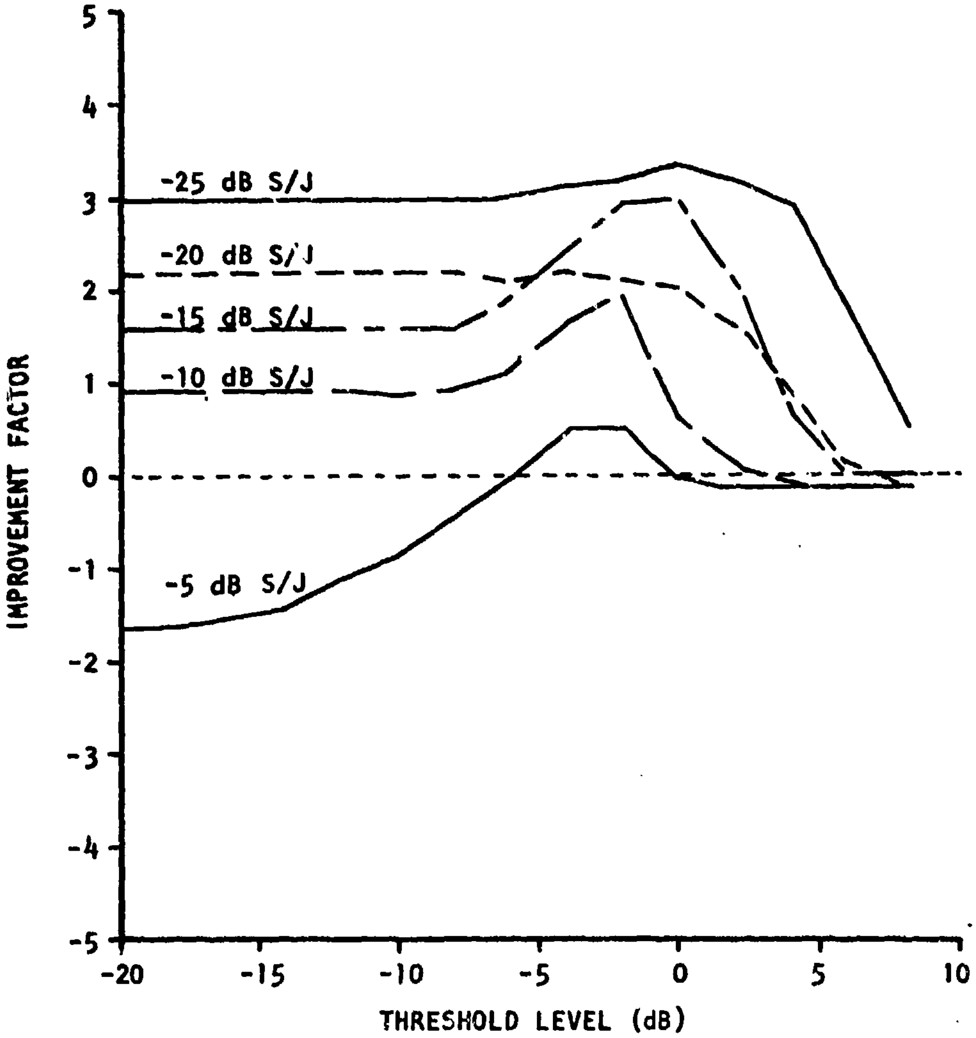


FIGURE 3-19. THRESHOLD SENSITIVITY FOR TAYLOR PROCESSING, 5% JAMMER BANDWIDTH

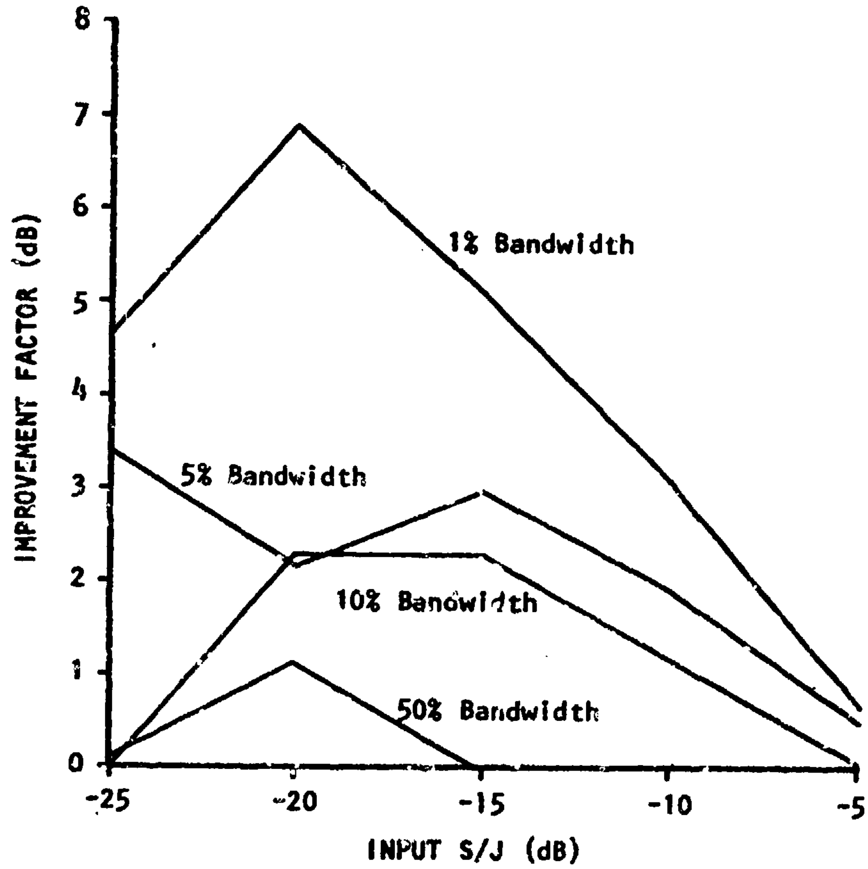


FIGURE 3-20. TAYLOR PROCESSING PERFORMANCE AS A FUNCTION OF JAMMER BANDWIDTH

results for input S/J ratios from -25 to -5 dB. The bandwidths considered ranged from 1% to 50% of the desired signal bandwidth.

Figure 3-20 indicates that the Taylor processor only provides significant improvement for jamming signals whose bandwidth is a small fraction of the signal bandwidth. The improvement factor of about 2.3 dB for 10% jammer bandwidths in the range of -15 to -20 dB input S/J is comparable to the improvement shown by the AGMF and the non-coherent clipping/blanking techniques under similar conditions. The Taylor porcessor however, unlike the AGMF requires no "learning" time and hence is applicable to non-stationary signals. On the other hand, the threshold selection problem cannot be dismissed because, under certain conditions, the performance is sensitive to proper threshold selection.

As an alternative to the selection of appropriate thresholds, it was reasoned that incorporating some form of AGC as the nonlinear element in each of the channels, might accomplish the desired result. Two types of AGC's were examined: a peak normalizing AGC and an RMS normalizing AGC.

The peak normalizing AGC basically determines the peak (magnitude) observation component in each channel and divides all components by that value. This FNF processor was examined for the same cases as presented in Figure 3-20. Figure 3-21 contains these results. This figure shows that the peak-normalizing AGC processor performs better than the optimum threshold Taylor processor for all of the 1% and some of the 5% bandwidth

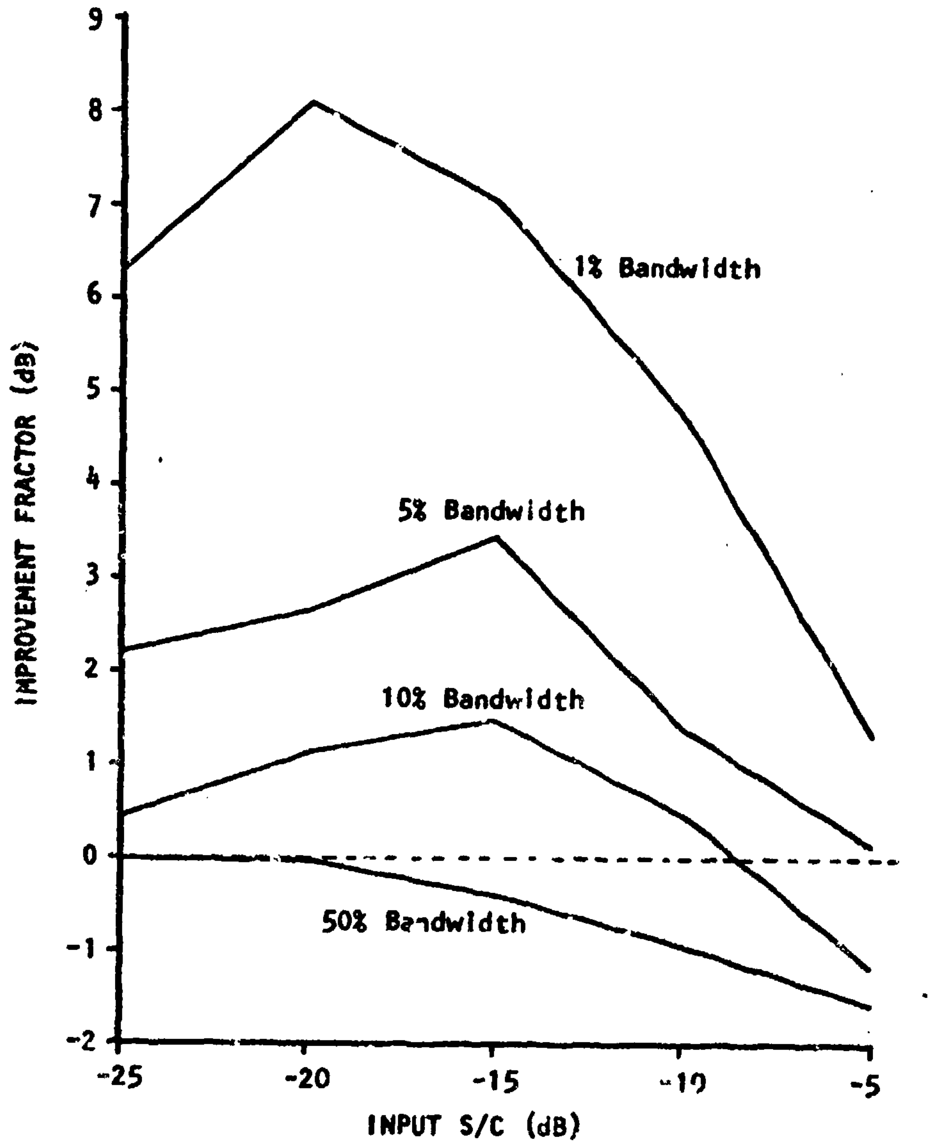


FIGURE 3-21. PEAK-NORMALIZING AGC PROCESSOR PERFORMANCE AS A FUNCTION OF JAMMER BANDWIDTH

cases. For the 10% to 50% jammer bandwidths the Taylor processor seems capable (with optimum threshold) of out-performing this AGC technique.

A second AGC was implemented which normalized the observations in each channel by the RMS power in that channel during the processing window. The results from this technique are shown in Figure 3-22. The performance of the RMS-normalizing AGC is, for the most part, comparable to the peak-normalizing AGC. The same conclusions concerning the relative merits of the Taylor processor also apply here. At 10% jammer bandwidth and in the range of -15 to -20 dB input S/J, where detection performance is marginal, both AGC techniques show disappointingly small improvements.

The results presented in this subsection reinforce the notion than a jamming signal that is confined in frequency to less than the bandwidth of the signal of interest, can be reduced in its effect without the precise knowledge of or estimate of its particular spectral density. However, the jammer bandwidth must be a relatively small portion of the signal bandwidth in order for any significant improvement to be realized. Although specific implementation details of the Taylor processor or the modifications considered may influence their performance capabilities, it is felt that the results from this simulation indicate the general level of improvement that may be attained using these nonlinear processing techniques.

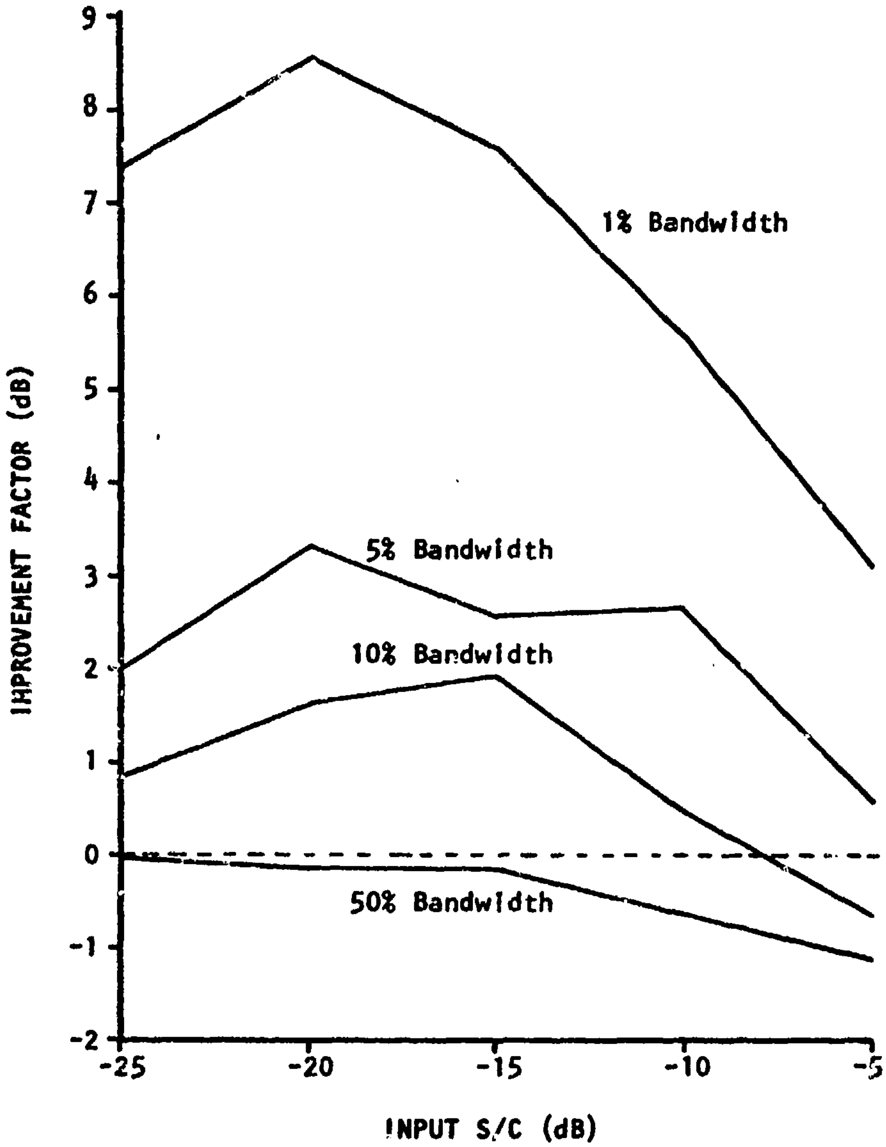


FIGURE 3-22. RMS-NORMALIZING AGC PROCESSOR PERFORMANCE AS A FUNCTION OF JAMMER BANDWIDTH

4.0 CONCLUSIONS

The investigations documented in this report are fundamentally investigations into ECCM techniques for radars performing BMD functions. A number of ECCM techniques such as sidelobe cancellation and frequency agility are known to be applicable for BMD radars. These present investigations are limited to ECCM techniques which can be implemented in the receiver or signal processor portion of the radar. In particular, nonlinear techniques for the mitigation of the effects of jammers have been studied. As a prerequisite to examining and generating suitable techniques, the potential vulnerabilities of jamming signals had to be identified.

It is a standard assumption that BMD jammers will be wide band thermal-like (i.e., Gaussian) noise sources. With this assumption, the optimum linear signal processing for maximizing S/N is the conventional matched filter. Moreover, the implementation of the optimum detector in either the Neyman-Pearson or Ideal Detector sense can be realized by the matched filter. Therefore, under the assumption of wide band Gaussian noise jammers and the assumption of stationarity of the interference, there appears to be no benefit to nonlinear processing. In the event that stationarity is not present, the well-known constant false alarm rate (CFAR) processing is applicable. CFAR processing is indeed nonlinear.

If the objective of jamming is to increase the interference level in the signal porcessor, it is certainly true that wide band Gaussian noise jammers are not optimum. In fact, for a fixed amount of jamming power, the offense could maximize the interference level in the signal processor by reproducing the radars signal, were it possible. This extreme is, however, not anticipated for BMD jammers because of the inherent covertness resulting from high BT product BMD waveforms.

The optimum strategy on the part of a BMD jammer designer could depend upon his knowledge of the defense's waveforms and signal processing, the level of sophistication possible for the jammers, and the cost extracted from the defense. This cost may be detection denial, false alarm saturation, track corruption, or even destruction of discrimination capability. It was beyond the scope of this effort to assess these issues and to formulate an offensive jamming signal repertoire. However, in light of these considerations, it was necessary to establish certain generic jamming signal categories that might encompass a range of signal characteristics which could 1) be particularly detrimental to conventional linear processing, and 2) possess vulnerabilities which might be exploited by nonlinear signal processing.

The classification of jamming signals resulted in three broad classes based upon stationarity. The first of these classes is denoted as Stationary I which contains signals which are essentially stationary over long times. The second class consists of those signals which are essentially stationary only over several signal processing windows. These are denoted as Stationary II. Finally, the third class of signals are non-stationary to the degree that signal characteristics vary significantly over one or two signal processing windows.

The rationale for these three jamming signal classes is based upon the ability of the defense to estimate pertinent characteristics of the jamming signals in order to appropriately modify the signal processing. For Stationary I signals these characteristics are assumed to be determinable with an arbitrary precision. For Stationary II signals, estimation errors of these characteristics may significantly affect processor performance. For non-stationary signals, all pertinent characteristics cannot be determined.

Below the three main classes of jamming signals, thirteen categories are established based upon power spectral density, duty cycle, and amplitude distribution. Six nonlinear techniques were examined for applicability to these jamming signal categories. Of these six, three were determined to have reasonable BMD-compatible implementations. These three were evaluated by signal processing simulations. One or more variations on each of these techniques displayed the ability to provide some measure of improvement in signal plus interference-to-interference ratio over conventional matched filter processing. Improvements in detection probabilities were determined when possible with reasonable simulation execution times.

The following paragraphs delineate the six nonlinear signal processing techniques examined and state the conclusions concerning each one:

- <u>Adaptive Generalized Matched Filter (AGMF)</u>

  The AGMF is suitable for Stationary I and Stationary II jamming signals whose power spectral density is not constant over the frequency band of the signal of interest. The proper operation of the AGMF requires a spectral estimation procedure which may be defeated by non-stationary jamming signals. For the purposes of evaluation, the spectral estimation procedure was performed during the five signal processing windows immediately preceding the window being processed. It was shown that for narrow band Gaussian and constant envelope jamming signals the AGMF does perform better than the standard matched filter. For Gaussian jamming signals with 10% bandwidth compared to the signal of interest, about 2.9 dB improvement was attained. For wide band Gaussian noise, the AGMF demonstrated about a 1 to 2 dB loss in S/N.

- <u>Non-Parametric Detectors</u>

  Non-parametric detection techniques were investigated because they are designed so as to be somewhat independent of the particular distribution of the interference. Two types of detectors were considered - the sign detector and the Wilcoxon detector. However, it was discovered that narrow band jamming had the effect of reducing the number of independent noise samples and, as a result, destroyed the constant false alarm characteristics of these detectors. Additional implementation problems were uncovered relating to unknown signal phase, unknown Doppler frequency, and pulse compression waveforms. Because of these problems, the non-parametric detectors were not considered for detailed performance evaluation.

- <u>Non-Coherent Clipping/Blanking</u>

These are two similar techniques whose benefits lie in the ability to reduce interference levels for jamming signals that are either narrow band (frequency domain non-coherent clipping/blanking) or pulsed (time domain non-coherent clipping/blanking). These techniques are implemented prior to the conventional matched filter. They are attractive in the fact that no jamming signal estimation parameters, other than perhaps power, are required. This implies that effects from non-stationary jamming signals may be reduced. Against narrow band jamming, 10% bandwidth, the frequency domain non-coherent clipper and blanker were able to demonstrate improvements of 2.9 dB and 2.7 dB respectively. Against pulse jamming, both the time domain non-coherent clipper and the blanker performed exceptionally well. Cases in which signal-to-noise (thermal) was high but detection performance was marginal because of the jamming were transformed into high probability of detection situations. The fact that time domain clipping/blanking performed better on pulsed signals than frequency domain clipping/blanking on narrow band signals is partially due to the fact that narrow band jamming signals were modeled as having continuous, spreading spectrums covering the entire signal band, while the pulsed jamming signals were actually zero during a portion of the signal.

- <u>Signal Space Blanker</u>

  This technique was judged to be useful in a situation where a swept, constant power jamming signal is present in the operating range of the radar at all times, but present in the receiver pass band only momentarily. It is based upon sensing the amount of jamming energy projected on the signal space and instantaneously adjusting the gain or detection threshold to prevent a false alarm. Although this technique was judged to be useful, its applicability is limited to such a restricted class of jamming signals that it was not analyzed or simulated in great detail.

- <u>Markov Processor</u>

  Markov processing refers to a class of post-detection signal processors that register a number of system "states." The probability of transitions between states is, theoretically, dependent only on the current state. One state might correspond to a declared detection, while another might correspond to a request for another verification pulse. It was concluded that Markov processors inherently require a long time or many pulses on target and hence are not particularly suited for BMD radars.

- Filter-Nonlinearity-Filter (FNF)

  All FNF processors have the characteristic of operating on sub-bands of the signal of interest by forming channels with banks of filters. The objective is to isolate narrow band jamming in a fraction of the total number of channels. Three forms of non-linear operations were tested via simulation. A clipping operation and two types of AGC's were performed on each of eight channels. The clipping operation FNF is called a Taylor processor. It was effective in reducing 10% bandwidth narrow band jamming by up to 2.3 dB. Comparable results for the AGC implementations are about 1.5 dB.

Several broad generalizations are apparent from these investigations. The first is that it is possible via nonlinear processing to reduce the effects of jamming for BMD. Furthermore, there are techniques which are applicable for a broad range of jamming signals. All techniques found to be effective are predicted on the ability to isolate the jamming from the signal of interest in either time or frequency. No technique was uncovered which would capitalize on specific probability distributions of random jamming signals. The performance of the nonlinear techniques in narrow band jamming proved to be disappointing except for very narrow band (<10%) jamming signals. For pulsed jamming the time domain noncoherent clipper and blanker proved to be very beneficial when the environment was favorable (high S/N). The merits of these techniques must ultimately be considered for particular waveforms and particular environments in order to determine their effectiveness in reducing the detrimental effects of ECM.

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