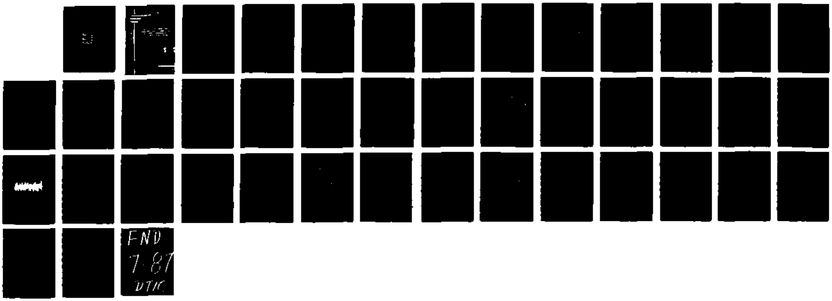
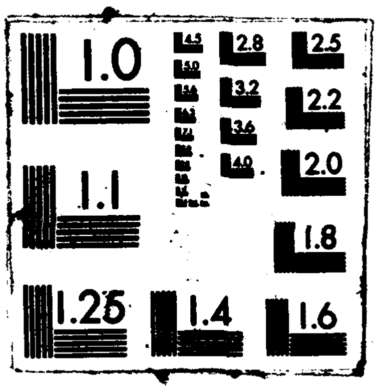
AD-A181 144 A METHOD FOR COMBATING FADING MULTIPATH CONDITIONS IN A DIRECT-SEQUENCE SPREAD-SPECTRUM SYSTEM(U) COMMUNICATIONS RESEARCH CENTRE OTTAWA (ONTARIO) G O VENIER JUN 86 CRC-1414 1/1

UNCLASSIFIED F/G 20/14 NL





1.0
1.1
1.25
1.4
1.6
1.8
2.0
2.2
2.5
2.8
3.2
3.6
4.0
4.5
5.0

3

DTIC FILE COPY

Communications Research Centre

A METHOD FOR COMBATING FADING MULTIPATH CONDITIONS IN A DIRECT-SEQUENCE SPREAD-SPECTRUM SYSTEM

by

G.O. Venier

The work described in this document was sponsored by the Department of National Defence under Project No. 32B76 (0117C11).

CRC REPORT NO. 1414
OTTAWA, JUNE 1986

Government of Canada – Department of Communications
Gouvernement du Canada – Ministère des Communications

Canada

COMMUNICATIONS RESEARCH CENTRE

DEPARTMENT OF COMMUNICATIONS
CANADA

A METHOD FOR COMBATING FADING MULTIPATH CONDITIONS IN A DIRECT-SEQUENCE SPREAD-SPECTRUM SYSTEM

by

G.O. Venier

(Directorate of Radio Propagation and Systems (DRL))

CRC REPORT NO. 1414

June 1986
OTTAWA

The work described in this document was sponsored by the Department of National Defence under Project No. 32B78 (0117C11)

CONTENTS

A METHOD FOR COMBATING FADING MULTIPATH CONDITIONS
IN A DIRECT-SEQUENCE SPREAD-SPECTRUM SYSTEM

G.O. Venier

ABSTRACT

The large bandwidth of a direct-sequence spread-spectrum communications system provides increased time-delay resolution which can resolve multipath components that would not be resolved by the system without spreading. However, normal direct-sequence systems do not fully exploit this capability. In these systems the received signal is despread by correlation with the reference signal (the spreading waveform) at a single delay. In a multipath situtation where there are multiple resolvable delays one of these delays must be chosen for the correlation. If the path with that delay is fading then the signal may be lost at times, even when a strong signal exists at a different delay.

In the system proposed here the despreading correlation is performed at a number of delays and the results are combined to provide the improvement expected from a diversity system under fading multipath conditions. The combining is simplified by the use of differential-phase-shift modulation of the data. An added advantage of the proposed system is that the outputs of the multiple-delay correlator are ideal inputs to a simple but effective code-delay tracking system based on keeping the "centre of gravity" of the values in the correlation window centred in the window.

A design using the multiple-delay-correlator despreader was tested in a general-purpose spread-spectrum software simulator developed at CRC. Bit-error-rate plots were generated under various propagation conditions, including Rayleigh fading and multiple paths, and with both Gaussian and impulsive interference. As expected, when there were multiple fading paths a significant improvement over conventional systems was observed. The improvement was obtained in both Gaussian and impulsive noise, but much greater improvements were possible in impulsive noise when a hard limiter was added in front of the despreader.

1 INTRODUCTION

Fading is a serious problem for digital communications systems in the high frequency (HF) band, where it is common, since it causes severe degradation in the bit-error-rate performance, especially at high signal-to-noise ratios. A major source of fading is multipath propagation which results in interference between two or more signal components received over different paths. Very small variations in the relative path lengths can cause large changes in the relative phases of the signal components and, therefore, of the amplitude of the component sum. Multipath propagation also causes intersymbol interference in digital communication systems when the spread in path delays is a significant fraction of a data-symbol duration.

Diversity techniques are often used to reduce the effects of fading. Since the state of the multipath interference varies with time, frequency, and antenna position, the use of time, frequency, or space diversity can provide two or more independent samples of the fading function which can be combined to give improved performance against fading. These techniques require the use of extra bandwidth or equipment, however. The usual way of avoiding the effects of intersymbol interference is to use a low data-symbol rate. When this is not acceptable equalization techniques are necessary, but these are complex and have not as yet proven themselves at HF.

The large bandwidth of direct-sequence spread-spectrum transmissions provide a means of resolving, in time, the multipath components that cause the fading and intersymbol interference. If the spread bandwidth is greater than the inverse of the time-delay separation of the received signal components, then one of the components may be selected and the intersymbol interference and the fading due to interference largely avoided. Unfortunately, particularly at HF, each of the individual paths may itself be fading as a result of such effects as polarization variations, ionospheric disturbances, or the presence of very-closely-spaced components. However, these resolvable fading paths themselves provide a type of diversity that we may call "path-delay diversity". The increased bandwidth of a direct-sequence system is frequently employed as an anti-jamming (AJ) technique. In such circumstances it can also be used to combat fading by means of path-delay diversity without further cost, other than that of some additonal signal processing.

A normal direct-sequence system synchronizes to only one signal delay, effectively ignoring any other resolvable paths. As mentioned above, this can provide improvement when the individual paths are not fading, but not when they are fading. If the system can be made to look at more than one delay, the signals at these different delays can be combined in a path-delay-diversity scheme to improve the overall performance.

A method of accomplishing this by means of a matched filter matched to a short spreading code that repeats for each information symbol has been developed and tested before [1]; but the use of a short repeating code does not provide the good AJ performance expected from a spread-spectrum system using a long non-repeating spreading code. It is the purpose of this report to show how the path-delay diversity may be exploited in a direct-sequence system using a long non-repeating spreading code which does not compromise the AJ performance of the spread-spectrum system.

After this work was completed it was discovered that a similar technique called Rake [2] had been developed about thirty years ago. Rake was developed as an optimum receiver for multipath conditions rather than as a modification to a spread-spectrum system, and is different is some respects. At that time implementation of the technique was difficult; today the availability of digital signal processing components makes implementation of this kind of technique much simpler.

A direct-sequence spread-spectrum system using the proposed technique was tested under various propagation and interference conditions by means of a spread-spectrum simulation facility. This facility, developed at CRC, has been implemented in software for the VAX-11/750. The results of the simulation, which are presented in this report, indicate the improvement possible under multipath fading conditions.

2. DESCRIPTION OF THE TECHNIQUE

2.1 Resolving the paths

The time-delay resolution of a waveform is reflected in its autocorrelation function [3], and the main peak of the autocorrelation function has a width of about the inverse of the waveform bandwidth. Thus, the bandwidth increase from the direct-sequence spreading code provides much greater resolution than is available from the unspread information-signal. As a result, multiple propagation paths with delay spread of much less than a data-symbol duration may be resolvable by the spread-spectrum signal.

The despreading in a conventional direct-sequence system is performed by correlation with a modulated replica of the spreading sequence, as shown in Figure 2.1. The integration, which may be performed by a filter, is over one data-symbol duration. The autocorrelation (assume for the moment that there is no noise or distortion) is computed at only one value of delay, and so the output of the correlator is a sample of the autocorrelation function at that particular delay. If two paths are present the correlation function will be repeated for the second signal, but offset from the first by the delay difference. If this delay difference is greater than the duration of a spreading code element the delay of the reference can be set equal to the delay of one of the paths, and the correlation sample will fall on the correlation peak for one path and will not be much affected by the other. Actually, since the correlation is over a finite length of the pseduo-random spreading sequence, the correlation outside the main peak will not be zero but will have random value with an average magnitude much less than that of the peak. Therefore there will be some interference between paths, but it will be relatively small.

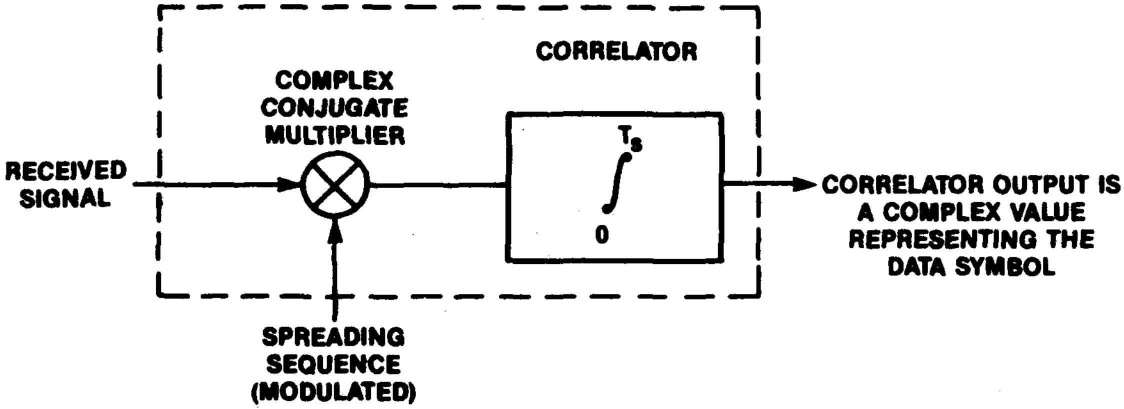


Figure 2.1 Direct-sequence demodulation using a correlator.

2.1.1 Matched Filter

One method of utlizing the correlation output from more than one path is to replace the correlator with a "matched filter" matched to the spreading waveform. A matched filter is a device whose output is the correlation function between the input and the waveform to which the filter is matched. The entire correlation function is generated instead of just a single sample as in the case of the correlator described above. For a large spreading ratio the matched filter must be quite complex since it must have stored in it the entire waveform to which it is matched. Another serious drawback is that the matched filter is matched to a particular waveform, and, since it is desirable that the spreading waveform not repeat from data symbol to data symbol, satisfying that requirement would mean that the filter would have to be reprogrammed for each new data symbol. For a particular data symbol the output of the matched filter represents the correlation as a function of delay. Therefore, it is necessary to combine samples of the output at different delays in a diversity scheme to provide a single value from which the information symbol can be determined.

2.1.2 Multiple-Delay Correlator

Another method of generating the correlation function at more than one point is simply to use multiple correlators at different delays as shown in Figure 2.2. This gives us a sampled version of the correlation function in some time-delay window. If the entire correlation function (over a data symbol) were required this method would be as complex as the matched filter technique, but normally only a relatively small portion of the correlation function is required to encompass the multipath spread. We shall refer to the device of Figure 2.2 as a "multiple-delay correlator".

The integration in the correlation process is over one data symbol duration. This implies block processing with integration over exactly one data symbol. But, since the multipath causes time spreading of the symbol, a block encompassing more than one data symbol must be processed. This

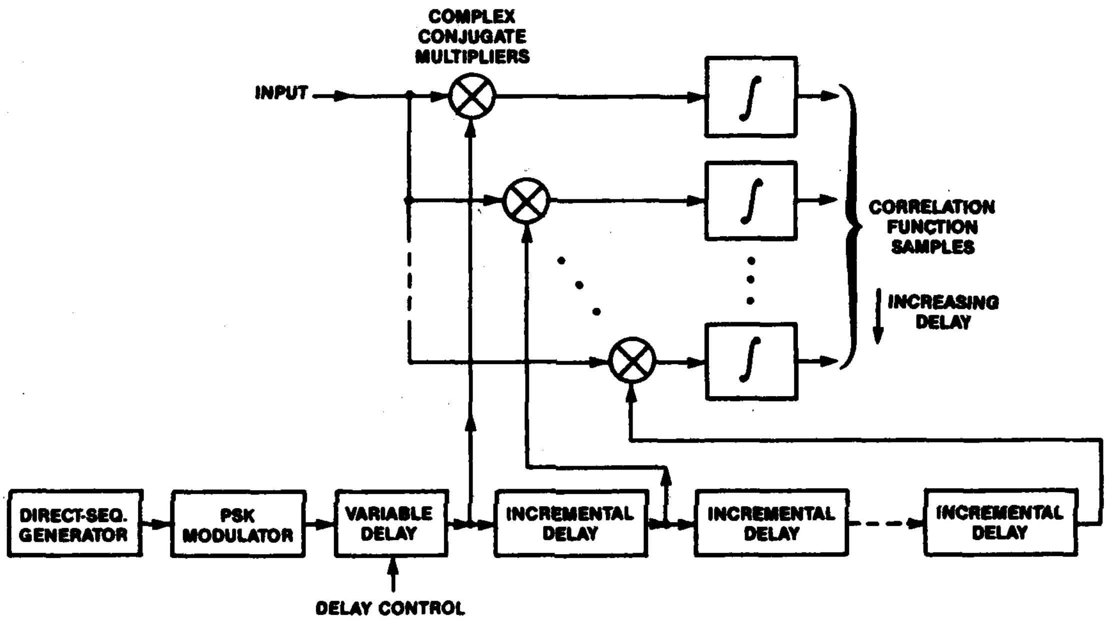


Figure 2.2 Multiple-delay correlator.

means successive blocks of input signal must be overlapping. The procedure is illustrated in Figure 2.3 which shows a rather trivial example where there are only ten samples per data symbol. The spreading code elements are not shown as they are not important to the explanation, but these elements, of course, must each contain less than ten samples. The top line represents the input signal over 13 samples. This is the block that has been chosen to provide a four-sample window for the correlation function. The four lines below the signal samples represent the reference samples (replica of the spreading sequence) for four correlators with delays incremented by one sample for each correlator. Only the ten samples corresponding to one data symbol (it is assumed that the time relationship between the spreading sequence and the data symbol edges is known as a result of timing control at the transmitter) are used and extended as shown with zeros to match the 13 signal samples. For the next block the last three signal samples of the block just processed are repeated at the beginning of the new block to advance the signal by exactly one data symbol. The overlap of three samples provides a window of three-plus-one or four samples. The restriction of the reference to only one data symbol duration reduces the adverse effect of adjacent symbols on the correlation.

2.1.3 Frequency-Domain Processing

When the delay window is a significant fraction of the data-symbol duration it may be more efficient to compute the correlation by transformation to the frequency domain, where a sample-by-sample multiplication of the signal and reference spectra would generate the transform of the entire correlation function. A Fourier transform of both the signal and reference and an inverse transform of the product are required. The transforms must be computed over a data-symbol duration plus the appropriate correlation window, and zeros must be added to the end of the reference to extend it from one data symbol to the transform length to avoid problems resulting from the cyclic nature of this process. The blocks of input data processed must advance by exactly one data symbol for each information bit processed; that is, the blocks must overlap by the extent of the window, as was the case for the multiple-delay correlator.

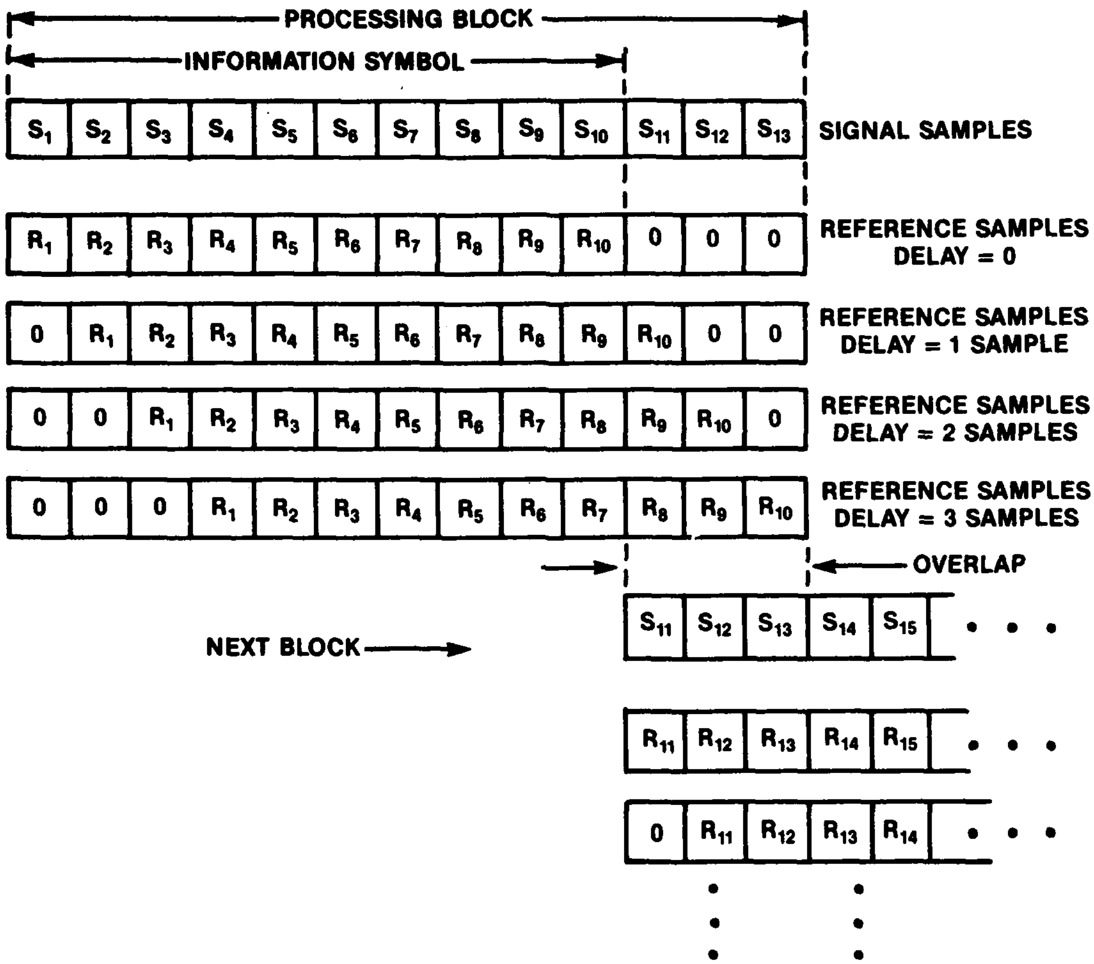


Figure 2.3 Illustration of multiple-delay correlator block processing timing.

This method may also be viewed as the implementation of a matched filter, but as described it is reprogrammed for each data symbol. If the reference were a fixed code repeated for each data symbol the reference transform would only have to be computed once and the result then used repeatedly, a process more normally associated with a matched filter. The frequency-domain method can require less computation than correlation in the time domain if the correlation output window is large relative to a data-symbol duration.

2.2 Combining the resolved paths

There is a problem with combining the outputs at different delays. We would expect the relative phase among the different rays to be random, and since the information is contained in the phase of the phase-shift-keyed (PSK) signal used in direct-sequence spread-spectrum systems, we would like to combine them to add coherently and give the correct phase from which to determine the information. To accomplish this in a fully coherent system would require the use of a separate phase-tracking system for each delay. After subtracting the corresponding reference phase from each correlation sample we could coherently add the results to determine the signal phase. But if we differentially encode the information bits (we will consider only binary information) before spreading the PSK modulation, and perform a differential process after correlation but before combining, as shown in Figure 2.4, each delay will give the same differential phase and the combining can be performed coherently. In this case the preceding information bit plays the role of the phase reference at each correlation delay. Since the use of differential PSK instead of fully coherent PSK is the rule at HF because of the unstable propagation paths, the small loss involved in differential PSK should not be considered a penalty resulting from the use of the multiple-delay correlator.

Consider in a little more detail the problem of combining the output at the different delays in an optimum manner. Figure 2.5 shows a hypothetical correlation function generated by the multiple correlators. Only the magnitude of the correlation is shown although the phase

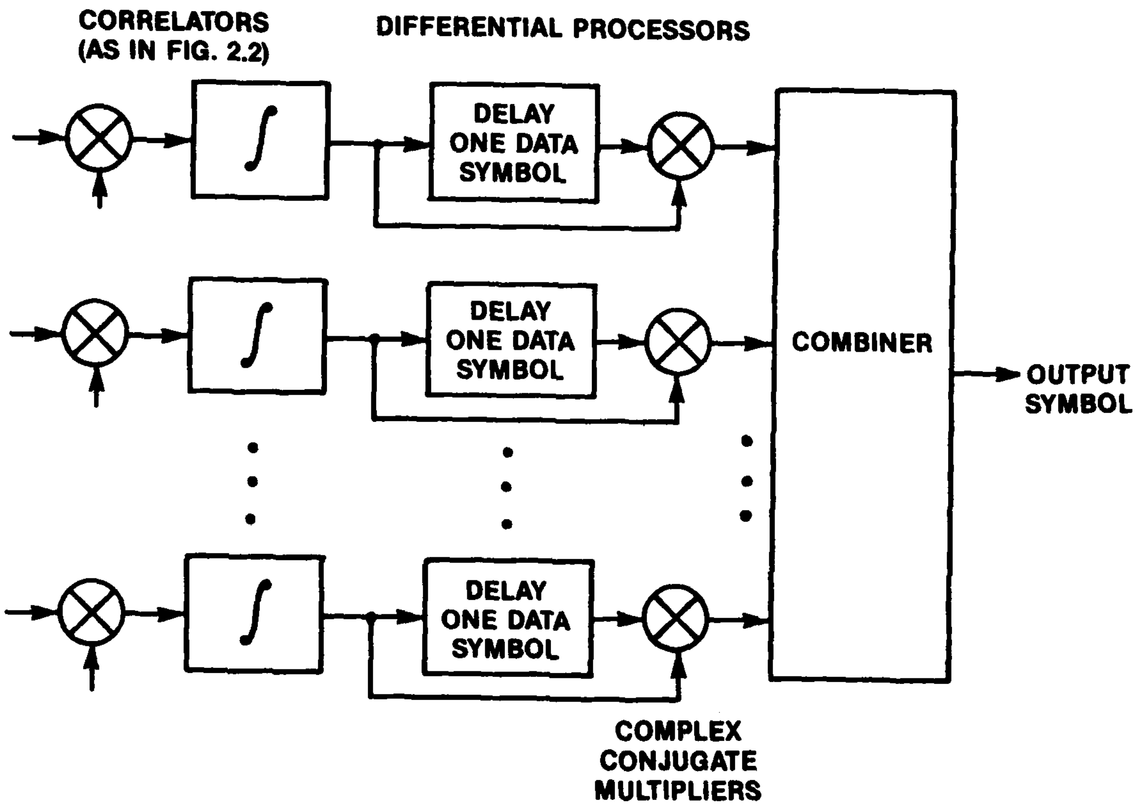


Figure 2.4 Differential demodulation and combining.

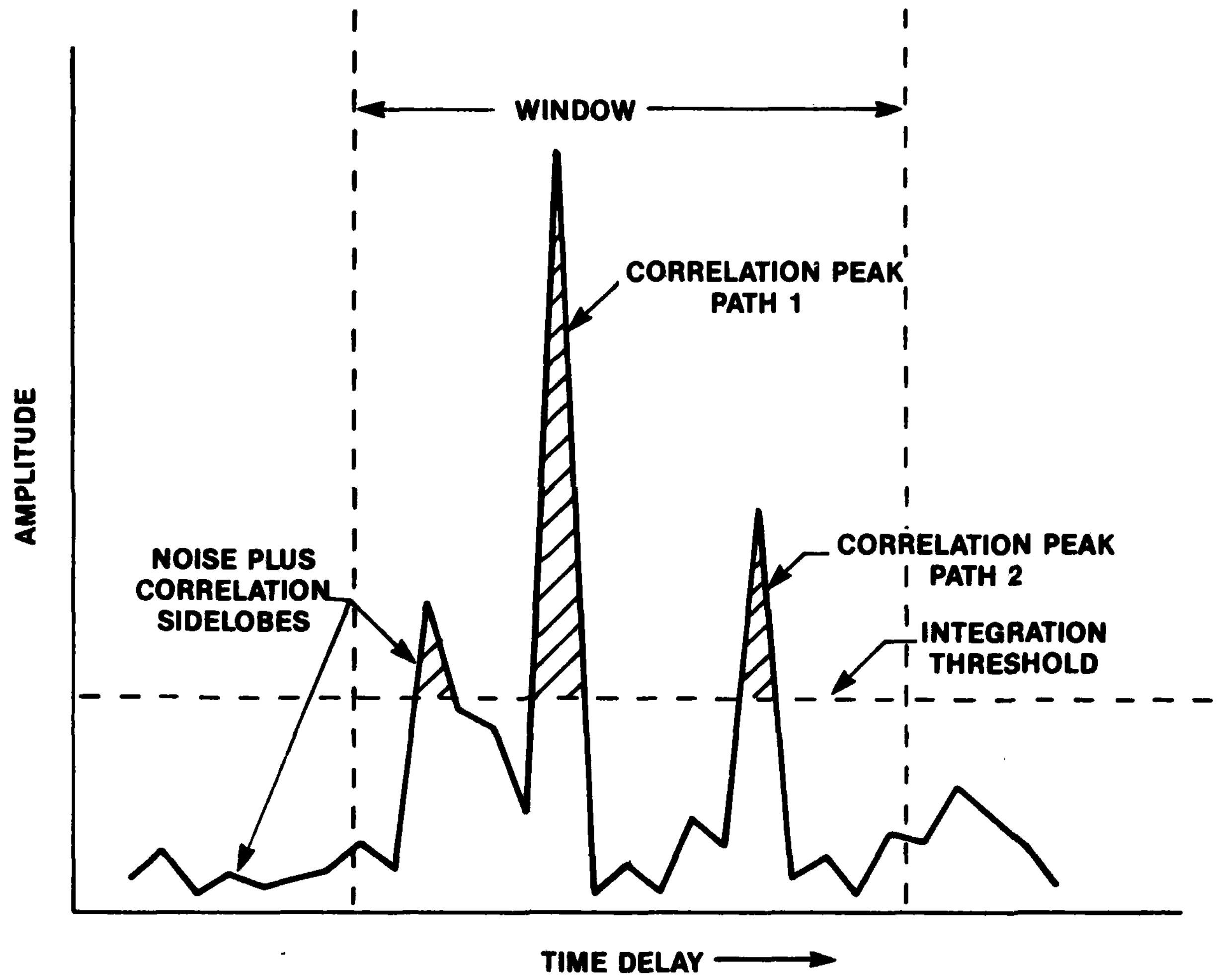


Figure 2.5 Example of correlation function and window.

information is assumed to be available as well. The horizontal axis represents the delay, and although the function consists of only discrete samples, a continuous function is shown for clarity. In this example there are two separate paths each generating a correlation peak. Of course these peaks will contain noise as well as signal. Also shown is the interference which consists of noise and correlation "sidelobes" that are caused by the random correlation occuring for delays that do not correspond to actual path delays. These sidelobes may be considered as noise since their phase is random. Ideally, as much of the energy in the peaks corresponding to the paths as possible would be integrated, and as little of the noise as possible. If the signal level at each point were known the best procedure would be to integrate over all delays in the window, with a weighting equal to the signal level at each delay.

When the signal voltage is much greater than the noise voltage the signal-plus-noise voltage is a good estimate of the signal voltage and therefore the square of the signal-plus-noise voltage can be used as the input to the integration. In fact, since the differential process involves a multiplication of the signal by itself, the correlation output samples already represent a squared signal voltage and no other squaring is necessary. When the signal voltage is in the range between zero and slightly above the noise voltage the signal-plus-noise voltage is not a good estimate of the signal level. Generally, most of the delays in the correlation window will contain only noise and it is reasonable, therefore, to set the estimated signal voltage to zero for delays where the signal-plus-noise voltage fails to exceed a threshold set a little above the mean noise level across the window. Thus, the proposed combining scheme is to estimate the mean noise level in the window, multiply this by a suitable factor to fix a threshold, and sum all correlation samples in the window that exceed this threshold. Such a scheme would not be far from optimum if the threshold were well-chosen.

The estimate of the noise level may be obtained by averaging the magnitude of the samples in the correlation output window. The threshold should be set at the measured rms level multiplied by some factor that will result in a reasonably high probability of it being exceeded by a signal with the minimum useful signal-to-noise ratio. The presence of the communication signal in the window will affect the noise estimate, but if the part occupied by the signal is a small fraction of the window, this will not be too serious. In any case, the threshold setting is not a very critical factor, and is most critical when the signal-to-noise ratio is low, which is the time when the presence of the signal will least influence the estimate. If the threshold is exceeded nowhere in the window the largest value in the window should be taken as the integrated signal, since it most likely represents the signal. However, if error-correction coding is used it may be preferable not to make a decision in this case but to mark the bit as an erasure. For computation of the integration threshold, a multiplication factor in the range of one to three has been found to give reasonably good results. Large multiplication factors give better performance in single-path propagation conditons since ideally only the peak should be integrated, while a somewhat smaller factor will give better performance under multipath conditions. A large multiplication factor (i.e., high threshold) generally results in selection of the largest peak for the symbol decision.

2.3 Time-Delay Tracking

When only a single-delay correlator is used time-delay tracking is required to keep the output at the autocorrelation peak. With a matched filter, tracking may not be required at all. In the case of the multiple-delay correlator, some tracking is advisable to keep the correlation window centered on the multiple paths, but the required precision is much relaxed from the single-delay case. In addition, the multiple correlation values provide an ideal input to the tracking algorithm.

The proposed tracking method is based on keeping the "centre of gravity" of the correlation samples in the centre of the window. The "centre of gravity" in samples relative to the centre of the window is defined as:

$$D_{cg} = \frac{\sum_{i=1}^{N_w} (i - (N_w + 1)/2)|X_i|}{\sum_{i=1}^{N_w} |X_i|}, \qquad |X_i| \geq V_t$$

where N_w is the number of correlation samples in the window, X_i is the i^{th} sample in the window, the V_t is the integration threshold (that is, the voltage below which the samples are not included in the integration). Since the "centre of gravity" is measured in relation to the centre of the window, it is a measure of the tracking error, and can be used as a correction to the direct-sequence reference-signal delay. Some smoothing should be applied by multiplying this error by a factor that is less than unity before it is used to correct the delay. The tracking time-constant will be the inverse of this factor.

3. SIMULATION EXPERIMENT

The multiple-delay correlator technique was tested under various conditions of propagation and interference by simulation using the CRC spead-spectrum simulation facility. In section 3.1 the parameters of the design chosen for testing are described and in 3.2 a description is provided of the simulation method and of the detailed characteristics of the simulated propagation and interference. Finally, the results of the tests are presented and discussed in section 3.3

3.1 Parameters of the Simulated Communication System

In order to evaluate the technique outlined, a design was chosen for testing on the CRC spread-spectrum simulation facility. This consisted of a direct-sequence system with a moderate spreading ratio of 100, and a spreading-code rate of 2000 elements per second. These two parameters determine the information rate of the binary DPSK system as 20 bits per second. This rather low information rate is a consequence of the intention of providing reasonably good AJ performance while restricting the spread bandwidth to about 6 kHz or a double HF channel. Actually, a spreading code rate approaching the channel bandwidth would be possible if appropriate filtering were applied, but a different approach using Hanning shaping of the code element envelope was chosen for the tests. This shaping broadens the main lobe of the signal spectrum and therefore effectively increases the processing gain by factor of about 1.7 over the nominal value of 100 given by the ratio of spreading code rate to data rate.

The design modeled in the simulation was based on a restriction of the transmitted bandwidth to 6 kHz. Removal of this restriction would allow the spreading code rate to be increased by a factor of about five or ten. Increasing the spreading code rate to 10,000 elements per second should provide better multipath performance than that of the tested system as a result of the higher resolution associated with the higher spread bandwidth.

A correlation delay window equal to the duration of four spreading elements, or two milliseconds, was used for most tests, but a window of only one sample was used for some tests to simulate a conventional direct-sequence system. The threshold was set at 2.5 times the rms value in the window, a value that turned out to be a little higher than optimum. The higher-than-optimum thereshold value assured that in most multiple-path cases the peak from only the strongest path was integrated, reducing the combining algorithm to one of selection of the strongest path. A few tests with smaller thresholds confirmed that for the multipath cases, the performance was slightly better with a threshold approximately equal to the rms value, but that the loss with the higher threshold was small (usually less than one dB) and did not seriously affect the results. On the other hand, the lower threshold caused slightly poorer performance (less than one dB) for single-path cases. As a result, the experiments were not repeated for the lower threshold except for the few cases mentioned.

In some of the tests a hard limiter was inserted before the direct-sequence correlators in the receiver. The hard limiter operated on complex signals. It had unity small-signal gain for input magnitudes up to 0.1 volts and a constant magnitude output of 0.1 volts for inputs above this value. Since the average signal level was about one volt, and the noise at this point was usually much higher, the limiter was in its non-linear range almost all the time. The limiter maintained the phase of the signal; i.e., the output phase was always equal to the input phase.

3.2 Simulation Details

The CRC simulator can simulate the operation of a complete spread-spectrum system including transmitter, propagation path, and receiver, and provides all the data and signal generation and analysis capabilities necessary to predict the performance of the simulated system under operational conditions. The system described above was tested on this simulator under various propagation and noise conditions. The set-up of the simulation is shown in Figure 3.1.

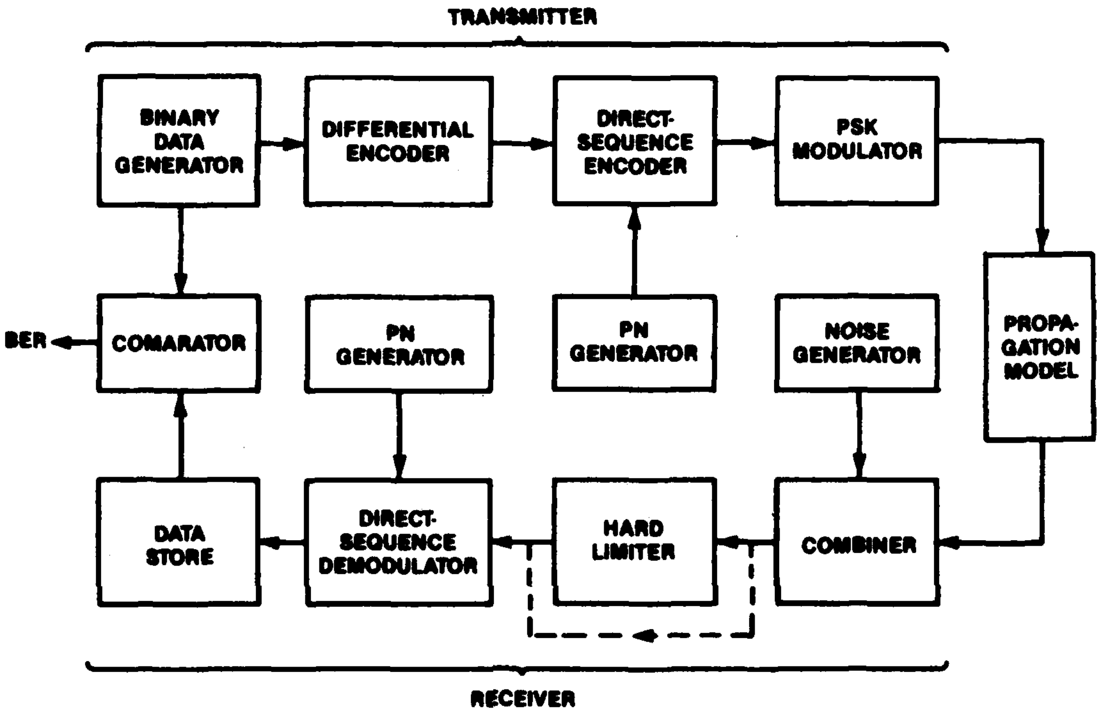


Figure 3.1 Simulation experiment setup.
Hard limiter used only in some tests.

3.2.1 Simuation of Waveforms

The waveforms were simulated by complex samples at zero carrier frequency. A sample rate of 6 samples per spreading element (12000 samples per second) was used. The adequacy of this sample rate can be seen from the spectrum of a spreading element shown in Figure 3.2. The power is more than 40 dB below the peak for frequencies beyond the half-sample-rate point (three frequency-times-symbol-duration units) where "folding" of the spectrum occurs; thus aliasing should not be significant at the chosen rate. In the receiver the demodulation was performed by the frequency-domain method described above. This was done because a general-purpose matched-filter capability was available in the simulator using this method, although for the window used the multiple-correlation method would certainly have been more efficient. The results would be identical, however, for the two methods, so this is not a restriction.

3.2.2 Propagation Path

The propagation path models used in the simulation tests included a single non-fading path, a single Rayleigh-fading path, a combination of two Rayleigh-fading paths, and a combination of 16 closely-spaced Rayleigh-fading paths intended to simulate spread-F conditions. The general algorithm used in the simulator for generating the Rayleigh-fading paths is described in Figure 3.3. The tapped delay line provides N_3 (specified by the user) delayed versions of the input signal which are multiplied by individual complex multipliers and summed to produce the multipath signal. The incremental and initial delays are specified by the user in terms of sample intervals. The complex multipliers, G_{ik}, are the normally distributed coefficients that determine the Rayleigh fading characteristics; each is generated by one of the generators shown in the blocks at the lower left. The root-mean-square (rms) amplitude and average Doppler frequency for each delay are determined by A_k and ϕ_k which are specified by the user. The fading rate is determined by the low-pass filter which is also specified by the user. This filter may be considered a narrow-band filter for complex waveforms, centred at zero frequency. The

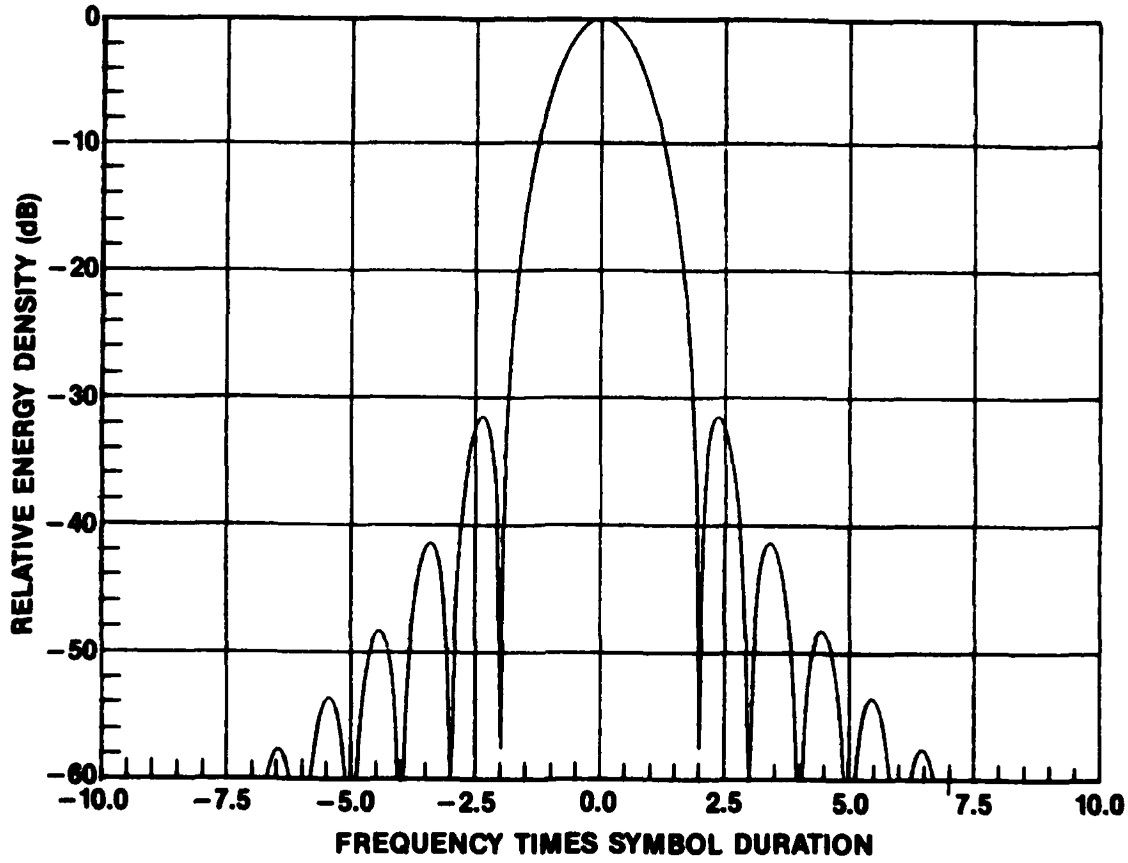


Figure 3.2 Spectrum of Hanning-shaped element.

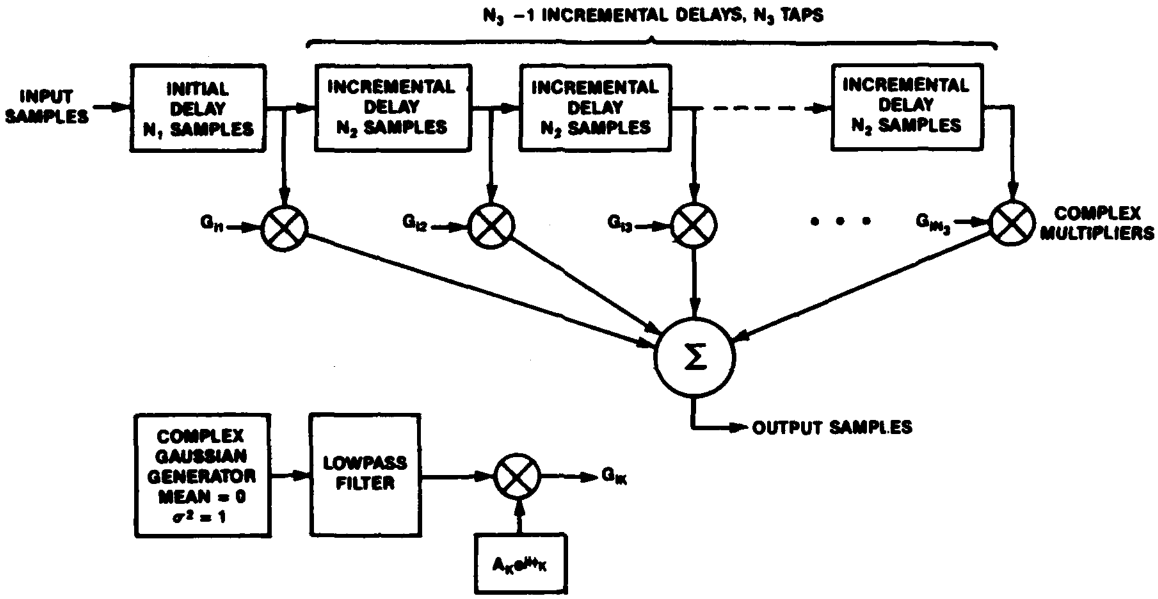


Figure 3.3 Multipath generation.

A_k = user-specified real multiplier for k^{th} tap.

ϕ_k = phase change between samples (Doppler) for k^{th} tap (user-specified).

G_{ik} = complex multiplier for i^{th} sample from k^{th} tap.

For each k, G_{ik} is generated as a time-series with time index i.

A seperate G_{ik} generator is used for each tap.

G_{ik} may be set to fixed values to simulate non-fading conditions.

For the tests performed on the system described here all rms amplitudes were made equal and the average Doppler frequencies were set to zero. The low-pass filter was a three-section filter resonant at zero frequency. Each section had a pole pair on the real axis of the Z-plane at 0.99951154. This gave a 3-dB half-bandwidth (from zero frequency to the 3-dB point) of 0.000035 times the sample rate, or 0.42 Hz, and an 18-dB half-bandwidth of 0.0000777 times the sample rate, or .933 Hz. Thus, the fading rate was a little less than one Hz. In the case of the two-delay path the incremental delay was 10 samples, or 1.7 spreading elements, and in the 16-delay case the incremental delay was just one sample. This means that the adjacent delayed signals were separated by less than the resolution capability of the spreading signal; the 16 samples spanned almost three spreading elements, however. In both of these cases the delay spread was much less than a data-symbol duration which was 600 samples.

3.2.3 Noise

Tests were run with two types of noise, Gaussian noise and noise modeled on the CCIR noise amplitude probability distributions [4]. In both cases the noise samples were independent, making the spectrum white across the 12,000 Hz bandwidth defined by the sample rate. The value of V_d, the ratio of rms to mean of the noise magnitude, was chosen from information in reference [4] for the bandwidth simulated. The value predicted in the reference is a function of time of day, time of year and transmission frequency. The value of 20.5 dB may be described as typical for the 12,000 Hz bandwidth.

The CCIR curves for the amplitude probability distributions are based on a limited number of measurements and represent a smoothed estimate. The true distributions may vary significantly from them at any given time and place, and this should be remembered when considering the results of the performance tests. Tests of the receiving system performed without the hard limiter shown in figure 3.1 should not be very sensitive to differences in the curves, but those in which the limiter was used will be far more sensitive to the differences. The reason for this is that the bandwidth reduction in the despreader tends to convert any distribution into a Gaussian one, but when the limiter is inserted before the despreader the noise power into the despreader is reduced by the limiter and the reduction is a strong function of the distrbution. The method of generating samples of the CCIR noise was taken from Akima [5], and tests have shown it to reproduce the CCIR distributions within about one dB. Some examples of noise generated by the simulator are presented below.

Figure 3.4 is a plot of the amplitude of a sequence of 512 complex noise samples generated by the noise generator with V_d set to 1.05 dB, the value that should result in Gaussian noise (Rayleigh amplitude). The rms voltage was set to one volt. No filtering was applied and the samples were, therefore, independent. The plot looks typical of wideband Rayleigh noise amplitude. A histogram of the noise amplitude computed from one million samples is plotted in Figure 3.5. The number of occurences in 0.1 volt amplitude bins was counted and divided by the bin size and by the total number of samples to make the vertical scale a probability-density-estimate scale. The histogram can be seen to have the expected Rayleigh shape. The Rayleigh case is included here as a reference against which to compare the impulsive CCIR noise.

A typical time sequence of 512 samples and a histogram, computed from 500,000 samples, for the CCIR noise with Vd = 20.5 dB as used in the simulation tests are plotted in Figures 3.6 and 3.7 respectively. The impulsive nature of the noise is evident from the former. Another point that should be noted is the reduction of the level of the majority of the samples between the large spikes (the rms level was one volt as it was for the Rayleigh case). This is a consequence of the high energy in the large spikes which contribute greatly to the rms value, in spite of their infrequent occurrence, since the energy is proportional to the square of the amplitude. Figure 3.7 also shows the low level of most of the samples and illustrates the long tail of the probability- density function. Note that the scales in figure 3.7 were changed in order to show both the low

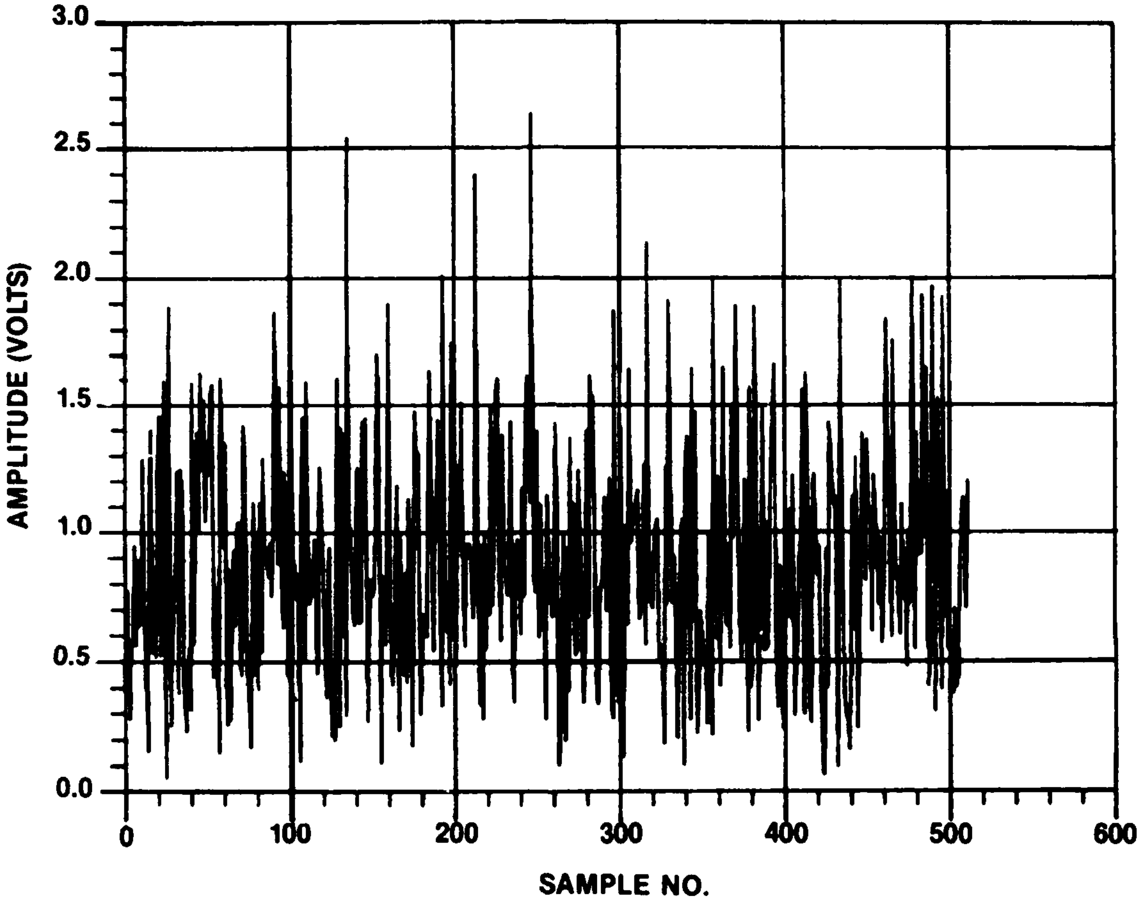


Figure 3.4 Sample of Gaussian noise (Rayleigh amplitude — V_d =1.05dB).

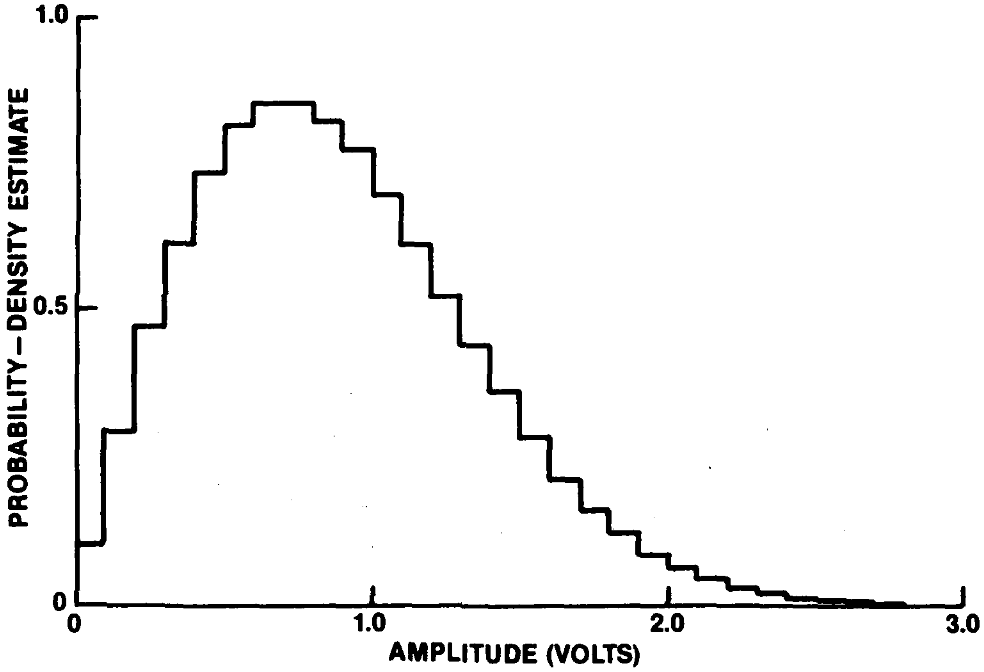


Figure 3.5 Histogram of Rayleigh noise amplitude — 10^6 samples.

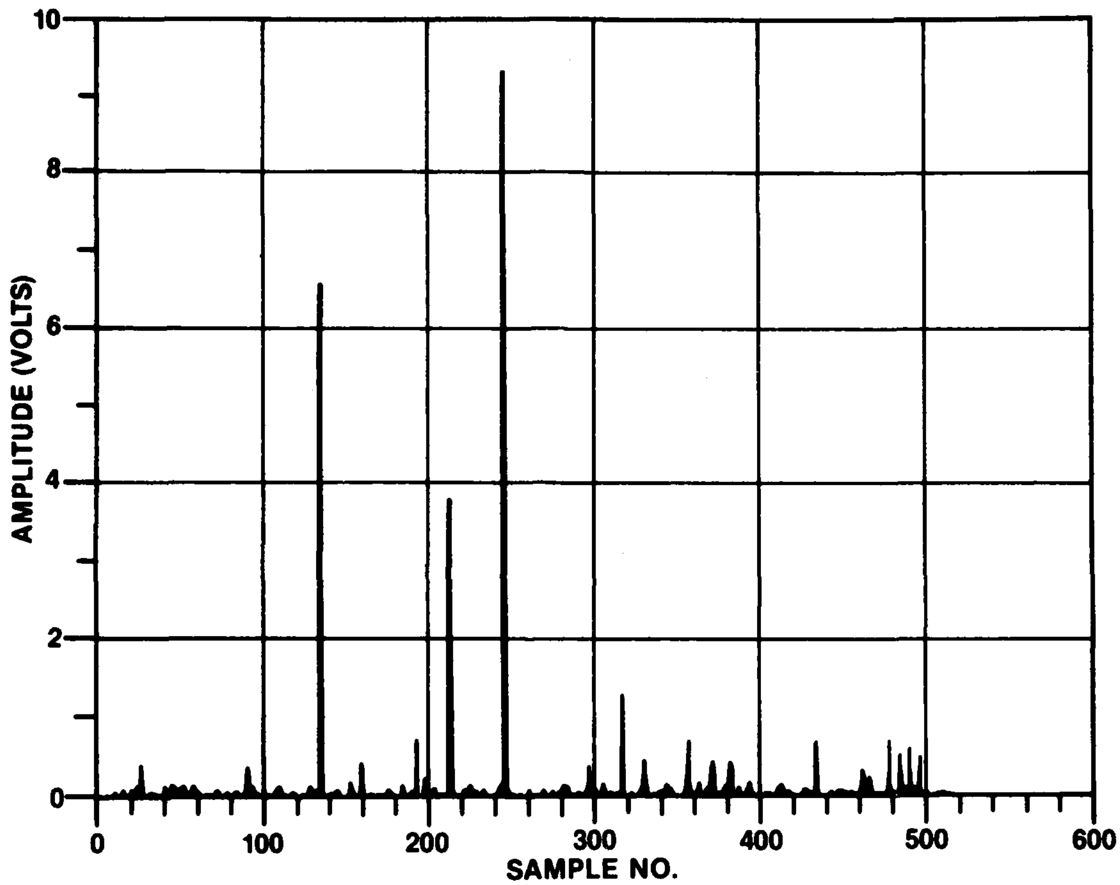


Figure 3.6 Sample of CCIR noise — V_d =20.5dB, 1 volt rms.

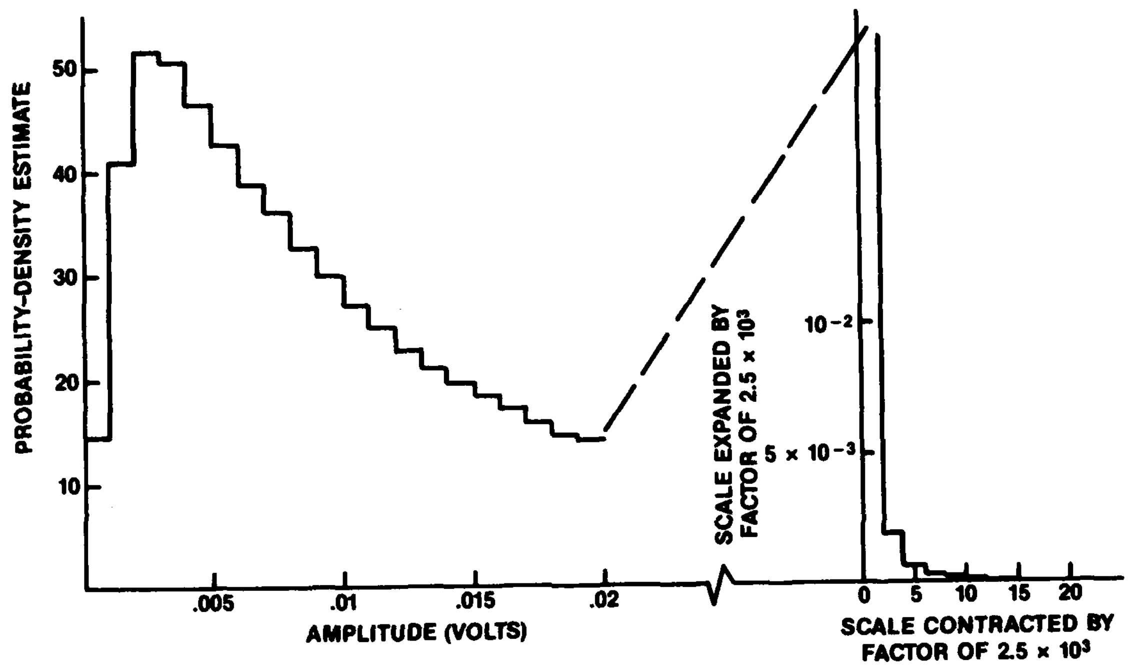


Figure 3.7 Histogram of CCIR noise amplitude — V_d = 20.5dB. Note change of vertical scale by a factor of 2.5×10^{-3} and horizontal scale by the inverse of this factor to conserve areas.

amplitude peak and the tail; the scales on the two axes were changed in opposite directions to conserve areas, which indicate the probability of amplitudes between different values. Although it can't be seen from Figure 3.7, from the 500,000 samples of impulsive noise examined 13 were above 80 volts, and the largest was 225 volts. By contrast, the maximum sample from the one million used in the Rayleigh test was 3.69 volts.

For comparison with the spread-spectrum system a non-spread binary DPSK system with the same data rate was simulated. The simulated non-spread DPSK signal had a bandwidth of only 0.01 times that of the spread signal, and, to save computing time, the signal after passing through the propagation medium was decimated by a factor of 100. Since the noise was added after this decimation the noise bandwidth was 100 times less than that for the spread system. This reduces the appropriate value of V_d to 4.2 dB according to curves in reference [4]. This is not surprising since reduction of bandwidth always makes non-Gaussian noise more like Gaussian noise.

3.3 Results

The results of the simulation tests with the various propagation models already described and for the Gaussian noise interference are shown in the bit-error-rate curves of Figure 3.8. The solid curves were added to show the theoretical expectations for DPSK for the cases of: no fading, Rayleigh fading, and Rayleigh fading with optimum diversity combining of two independent paths. Results from a simulation of a conventional DPSK system are included for comparison.

The curves for the multiple-delay correlator are marked "MDC". Error bars were not put on each point, but an indication of the 68% confidence interval (plus and minus one standard deviation) were added to the right of the graph for three different error rates. The confidence interval as a fraction of the error rate increases as the error rate decreases because it was not practical to increase the number of samples enough to compensate for the fewer errors generated at lower rates. The confidence intervals

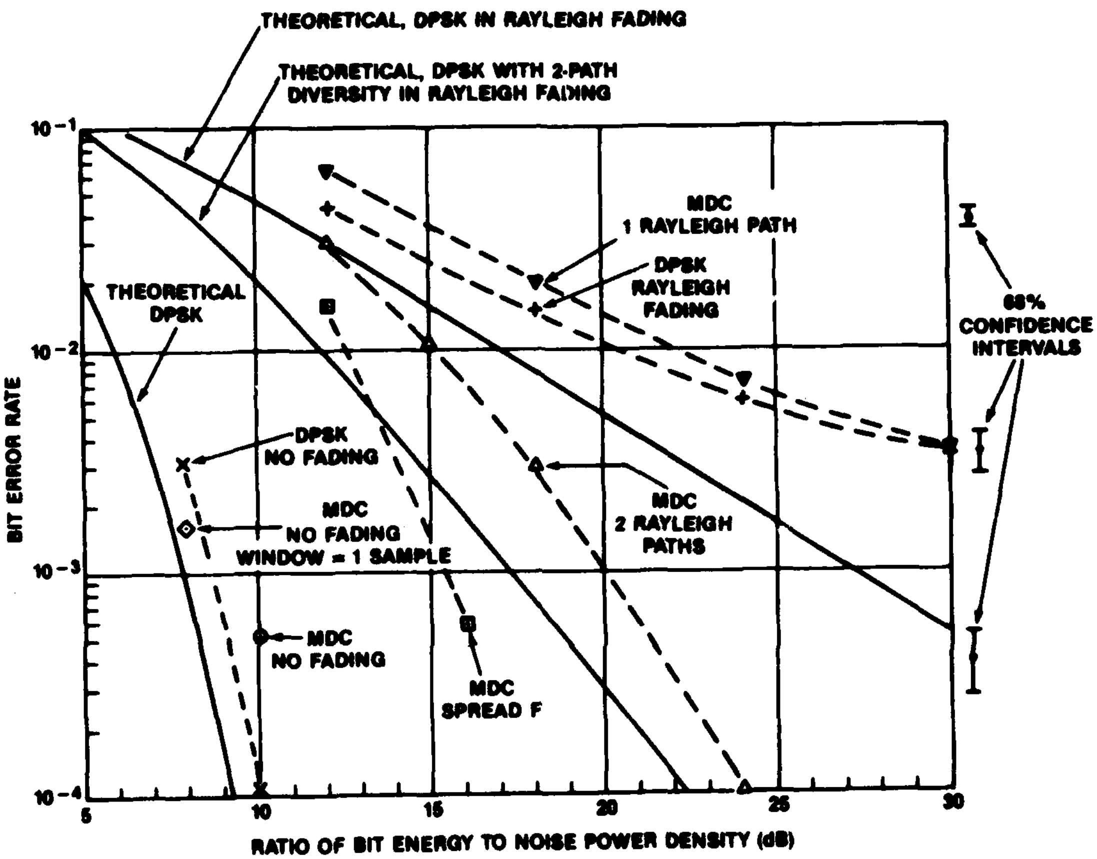


Figure 3.8 Simulation results for Gaussian noise.
MDC = multiple-delay correlator.

apply to the multiple-delay-correlator results; the DPSK results have somewhat smaller intervals.

The measured points for the simulated DPSK system without spreading when there was no fading are just to the right of the theoretical curve, indicating a loss of just under one dB. This loss is thought to be the result of the non-optimum filtering in the DPSK system, which permits noise-by-noise terms to be generated in the differential multiplication process that are not considered in the theoretical calculations. To test the multiple-delay correlator system under simple well-understood conditions the window was reduced to one sample and was centred exactly on the simulated delay. Under these conditions the matched-filter system reduces to a normal direct-sequence system with perfect synchronization, and the performance should be identical to an unspread DPSK system for white Gaussian noise. Figure 3.8 shows that the bit error rate falls within 0.5 dB of the theoretical no-fading curve for the one value of signal-to-noise ratio tested under these conditions. The result is slightly better than for the simulated DPSK system and this is attributed to the fact that the correlation (matched filtering) is performed before the differential process in this case, providing optimum filtering, while in the DPSK system the correlation is performed after the differential processing. In any case the difference is very small. When the window is widened to 24 samples, the multiple-delay correlator system performance in the non-fading case is degraded as a result of the integration over noise-only correlation values; there is a finite probability that noise alone will exceed the threshold, and also the correlation "sidelobes" act as noise, increasing the noise level slightly. The loss caused by the window appears to be less than two dB for the simulated results.

With a single fading path no path diversity is possible and the direct-sequence system provides no improvement relative to normal DPSK. In fact its performance is slightly inferior for the reason just mentioned for the no-fading case. Both the normal DPSK and the direct-sequence system show signs of an irreducible error rate at around 0.003. This is a result of the relatively rapid fading rate (just under one Hz) used in the tests which can cause phase reversals within one or two data-symbols on rare occasions.

When there are two fading paths the multiple-delay-correlator performance exceeds the theoretical single-path fading performance (by about seven dB at an error rate of 0.001) and falls within 3 dB of the theoretical two-path diversity curve. The total signal power for the two-path case was kept the same as for the single-path case, with the power in each path being reduced by a factor of two. No sign of an irreducible error rate is evident down to the lowest error rate tested. This limit of about 0.0001 in measured error rate was set by the enormous amount of computing time required.

The performance improves even further for the spread-F case, exceeding the theoretical single-path performance by more than 11 dB at an error rate of 0.001. This results from the fact that there are, in effect, about four independent paths (Although there are sixteen actual paths they are spaced closer together than the resolution of the direct sequence). The slope of the experimental curves show that the improvement for the multiple-delay correlator increases as the signal-to-noise ratio increases. Only one plot is shown for DPSK with fading since the results were essentially the same for one, two, or sixteen fading paths. This is to be expected, since the time separation of the paths was much less than a data bit interval.

Figure 3.9 shows the results when the Gaussian noise is replaced by CCIR noise for which there is a higher probability of very large values. This characteristic has the effect of reducing the slope of the bit-error-rate curves in the non-fading case and seems to reduce slightly the relative improvement of the multiple-delay correlator system. The multiple-delay correlator can take advantage of the impulsive nature of the CCIR noise, however, by the use of limiting before matched filtering. This has the effect of clipping the spikes, and although the desired signal is suppressed at the point of clipping, the subsequent correlation integrates the energy over a data-bit interval to provide good performance.

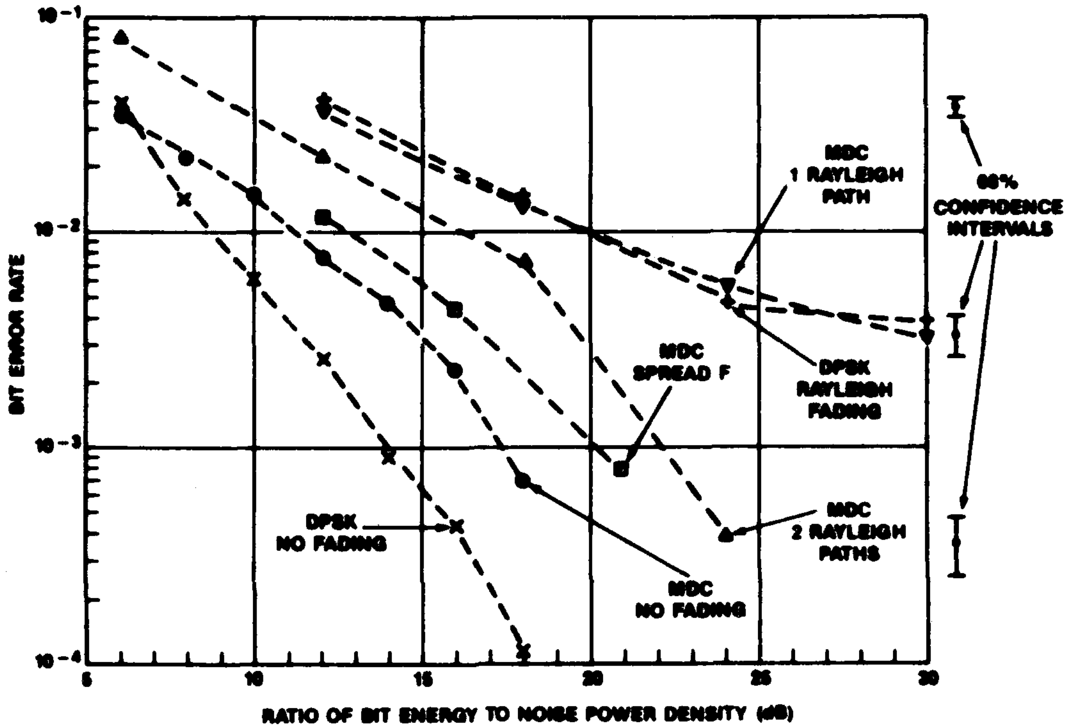


Figure 3.9 Simulation results for CCIR noise.
MDC = multiple-delay correlator.

Figure 3.10 shows the very great improvement obtained when a hard limiter precedes the correlators (see Figure 3.1 for position). The improvement is of the order of 40 dB, a rather surprising result. However, examination of the characteristics of the CCIR noise and of the operation of the demodulator provides an explanation for this improvement and gives confidence that this performance is to be expected when the noise has the characteristics specified by the CCIR model.

An explanation for the improvement with the use of hard limiting is the following. Much of the power in the impulsive noise resides in the relatively few very large impulsive samples since the power is proportional to the square of the amplitude. Therefore, when these samples are removed by a limiter while the bandwidth is still large the mean power in the noise is greatly reduced. Of course, the signal samples coinciding with these large noise samples are greatly suppressed and do not contribute significantly to the integrated signal value, but since the integration is over a large number of these samples (600 in the simulated system) the loss of even a substantial number of them will not cause a large degradation, and the remaining samples after integration must now compete with a much lower noise level. The integration also makes the amplitude probability distribution of the noise much closer to a Gaussian one than was that of the original wideband noise. If the integration were done before the limiting the energy in the large noise spikes would be averaged into the resulting near-Gaussian noise where it could not be removed by limiting. That is, once the bandwidth has been reduced, as in a non-spread DPSK system, the impulsive nature of the noise can no longer be fully expolited. Of course, in the non-spread system a wideband front end could be used before limiting and the limiting followed by filtering to reduce the bandwidth to a value suitable for the data signal.

The above argument might lead one to conclude that the pre-limiter bandwidth should be as great as possible, but it should be remembered that strong interference outside the signal bandwidth may be brought into the signal bandwidth by the nonlinear action of the limiter, and a compromise based on the characteristics of the interference in the band should be

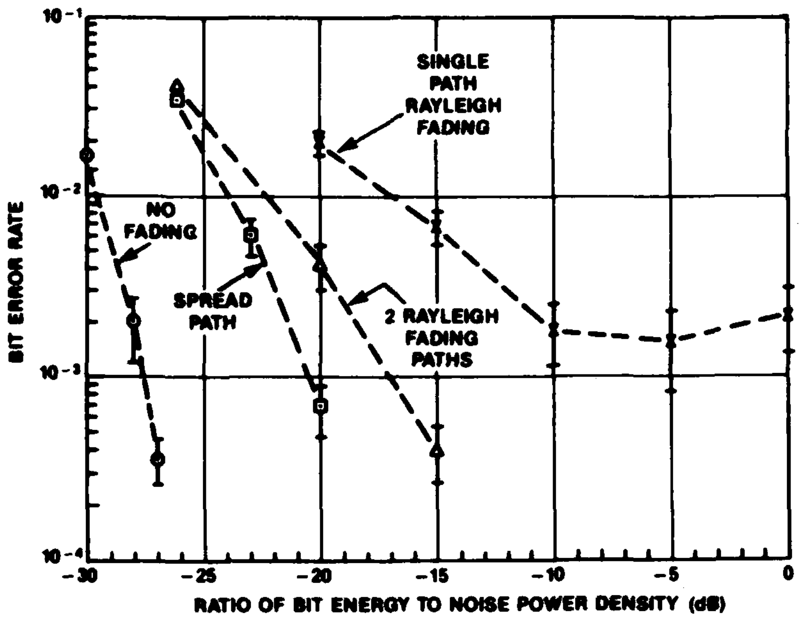


Figure 3.10 Simulation results for multiple-delay correlator in CCIR noise and limiting. 68% confidence intervals shown.

used.

While the use of hard limiting produces a large improvement when the noise is impulsive, it results in a small penalty when the noise is Gaussian. Simulation tests indicate that this penalty is about one dB.

4. CONCLUSIONS

The use of multiple-delay correlators to provide diversity in propagation conditions consisting of multiple Rayleigh-fading paths was found to be an effective technique for improving the bit-error-rate performance of direct-sequence spread-spectrum communication systems operating in those propagation conditions. The proposed method allows a long non-repeating direct-sequence spreading code to be used, and thus avoids the compromise in the ECCM capability that is associated with short repeating codes. Because it is expected that for most delays there will only be noise, the proposed combining scheme uses a threshold computed from the noise level to remove correlation samples with low signal-to-noise ratio from the integration process.

Simulation tests verified the performance improvement under fading multipath conditions when the noise was Gaussian and also when it was impulsive in accordance with the CCIR HF noise distributions. While the improvement was slightly less in the CCIR-type noise it was still significant and, in addition, it was found that the impulsive nature of the noise could be exploited by the use of limiting at the front-end of a direct-sequence system to provide very large improvements under all propagation conditions.

5. FURTHER WORK

The advantages of the path-diversity scheme proposed here depend on the occurrence of two or more fading propagation paths and on the amplitude probability distribution of the paths. Rayleigh fading, as used in the simulation tests, is the worst case for the non-diversity systems and, therefore, provides the greatest opportunity for improvement. It seems likely, however, that the fading of the individual paths at HF will be Rician rather than Rayleigh and consequently will have fades of less depth. It is important therefore to investigate the statistics of typical propagation paths to determine how well the model used here reflects actual conditions. It will be useful to carry out a measurement program using a system capable of resolving the multipath components down to at least 100 microseconds. Since the simulator is capable of generating paths with Rician statistics, the results of such an investigation could be used to set the simulation parameters for further tests.

Noise measurements with a bandwidth of at least 10 kHz, and preferably greater, would be useful to verify the impulsive characteristics of the noise in the HF band and therefore the advantage of the use of a limiter. Recordings of measured sampled complex baseband noise could be substituted in the simulator for the generated noise.

Finally, if the results of the above investigations indicate that the proposed techniques would offer significant benefits, the development of an experimental system should be considered. It is recommended that the signal processing for such a system be performed digitally using a microprocessor.

6 REFERENCES

[1] Chow, S., J.K. Cavers, and P.F. Lee, "A Spread Spectrum Modem For Reliable Data Transmission in the High Frequency Band", IEE Conference Publication Number 206 of the Second Conference on HF Communication Systems and Techniques, London, 1982.

[2] Price, R., "A Communication Technique for Multipath Channels", Proceedings of the IRE, vol. 46, pp555-570, March 1958.

[3] Woodward, P.M., "Probability and Information Theory, with Applications to Radar", Chapter 7, Pergamon Press, 1953.

[4] CCIR (International Radio Consultative Committee), Report 322, "World Distribution and Characteristics of Atmospheric Radio Noise", Figure 27, International Telecommunication Union, Geneva, Switzerland, 1964.

[5] Akima, Hiroshi, "A Method of Numberical Representation for the Amplitude-Probability Distribution of Atmospheric Radio Noise", Telecommunications Research and Engineering Report 27, Institute for Telecommunication Sciences, Office of Telecommunications, U.S. Department of Commerce, 1972.

7 ACKNOWLEDGEMENT

The work described here was sponsored by the Directorate of Maritime Combat Systems of the Department of National Defence.

UNCLASSIFIED

Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall document is classified)

1 ORIGINATING ACTIVITY Communications Research Centre	2a. DOCUMENT SECURITY CLASSIFICATION Unclassified 2b GROUP

3 DOCUMENT TITLE
A METHOD FOR COMBATING FADING MULTIPATH CONDITIONS IN A DIRECT-SEQUENCE SPREAD-SPECTRUM SYSTEM (U)

4. DESCRIPTIVE NOTES (Type of report and inclusive dates)
CRC REPORT

5. AUTHOR(S) (Last name, first name, middle initial)

Venier, Gerald O.

6. DOCUMENT DATE	7a. TOTAL NO. OF PAGES 38	7b. NO. OF REFS 5
8a. PROJECT OR GRANT NO.	9a. ORIGINATOR'S DOCUMENT NUMBER(S) CRC RPT 1419	
8b. CONTRACT NO.	9b. OTHER DOCUMENT NO.(S) (Any other numbers that may be assigned this document)	

10. DISTRIBUTION STATEMENT

Unlimited

11. SUPPLEMENTARY NOTES	12. SPONSORING ACTIVITY DMCS-6

13. ABSTRACT (U)The large bandwidth of a direct-sequence spread-spectrum communications system provides increased time-delay resolution which can resolve multipath components that would not be resolved by the system without spreading. However, normal direct-sequence systems do not fully exploit this capability. In these systems the received signal is despread by correlation with the reference signal (the spreading waveform) at a single delay. In a multipath situtation where there are multiple resolvable delays one of these delays must be chosen for the correlation. If the path with that delay is fading then the signal may be lost at times, even when a strong signal exists at a different delay.

In the system proposed here the despreading correlation is performed at a number of delays and the results are combined to provide the improvement expected from a diversity system under fading multipath conditions. The combining is simplified by the use of differential-phase-shift modulation of the data. An added advantage of the proposed system is that the outputs of the multiple-delay correlator are ideal inputs to a simple but effective code-delay tracking system based on keeping the "centre of gravity" of the values in the correlation window centred in the window.

A design using the multiple-delay-correlator despreader was tested in a general-purpose spread-spectrum software simulator developed at CRC. Bit-error-rate plots were generated under various propagation conditions, including Rayleigh fading and multiple paths, and with both Gaussian and impulsive interference. As expected, when there were multiple fading paths a significant improvement over conventional systems was observed. The improvement was obtained in both Gaussian and impulsive noise, but much greater improvements were possible in impulsive noise when a hard limiter was added in front of the despreader.

DSIS
77-065

UNCLASSIFIED
Security Classification

KEY WORDS

Spread Spectrum
Direct Sequence
Multipath
Fading
Limiting
Impulsive Noise
Diversity
Combining
Simulation
Code Tracking
Correlation
Differential Phase-Shift Keying

INSTRUCTIONS

1. ORIGINATING ACTIVITY Enter the name and address of the organization issuing the document.

2a. DOCUMENT SECURITY CLASSIFICATION: Enter the overall security classification of the document including special warning terms whenever applicable.

2b. GROUP: Enter security reclassification group number. The three groups are defined in Appendix 'M' of the DRB Security Regulations.

3. DOCUMENT TITLE: Enter the complete document title in all capital letters. Titles in all cases should be unclassified. If a sufficiently descriptive title cannot be selected without classification, show title classification with the usual one-capital-letter abbreviation in parentheses immediately following the title.

4. DESCRIPTIVE NOTES: Enter the category of document, e.g. technical report, technical note or technical letter. If appropriate, enter the type of document, e.g. interim, progress, summary, annual or final. Give the inclusive dates when a specific reporting period is covered.

5. AUTHOR(S): Enter the name(s) of author(s) as shown on or in the document. Enter last name, first name, middle initial. If military, show rank. The name of the principal author is an absolute minimum requirement.

6. DOCUMENT DATE: Enter the date (month, year) of Establishment approval for publication of the document.

7a. TOTAL NUMBER OF PAGES: The total page count should follow normal pagination procedures, i.e., enter the number of pages containing information.

7b. NUMBER OF REFERENCES: Enter the total number of references cited in the document.

8a. PROJECT OR GRANT NUMBER: If appropriate, enter the applicable research and development project or grant number under which the document was written.

8b. CONTRACT NUMBER: If appropriate, enter the applicable number under which the document was written.

9a. ORIGINATOR'S DOCUMENT NUMBER(S): Enter the official document number by which the document will be identified and controlled by the originating activity. This number must be unique to this document.

9b. OTHER DOCUMENT NUMBER(S): If the document has been assigned any other document numbers (either by the originator or by the sponsor), also enter this number(s).

10. DISTRIBUTION STATEMENT: Enter any limitations on further dissemination of the document, other than those imposed by security classification, using standard statements such as:

(1) "Qualified requesters may obtain copies of this document from their defence documentation center."

(2) "Announcement and dissemination of this document is not authorized without prior approval from originating activity."

11. SUPPLEMENTARY NOTES: Use for additional explanatory notes.

12. SPONSORING ACTIVITY: Enter the name of the departmental project office or laboratory sponsoring the research and development. Include address.

13. ABSTRACT: Enter an abstract giving a brief and factual summary of the document, even though it may also appear elsewhere in the body of the document itself. It is highly desirable that the abstract of classified documents be unclassified. Each paragraph of the abstract shall end with an indication of the security classification of the information in the paragraph (unless the document itself is unclassified) represented as (TS), (S), (C), (R), or (U).

The length of the abstract should be limited to 20 single-spaced standard typewritten lines, 7½ inches long.

14. KEY WORDS: Key words are technically meaningful terms or short phrases that characterize a document and could be helpful in cataloging the document. Key words should be selected so that no security classification is required. Identifiers, such as equipment model designation, trade name, military project code name, geographic location, may be used as key words but will be followed by an indication of technical context.

END

7-81

DTIC