A Multiprocessing Approach
To
Spread Spectrum Code Acquisition
Under Jamming

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ABSTRACT

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Spread Spectrum (SS) communication systems, although primarily used for military applications, are currently being examined for use in a commerical environment (such as mobile networks). Howevever, before spread spectrum systems adapt to commercial applications, key aspects such as multi-user interference, mobility, cost, acquisition and tracking techniques, etc., must be considered. In this thesis two applications of a new parallel detection scheme for spread spectrum signal acquisition are proposed. Both applications employ matched filtering as a first stage of signal acquisition, and several sequence detection circuits (SDCs) to allow the receiver to begin tracking.

The first application of the parallel detection scheme is for a direct sequence (DS) spread spectrum system. By modelling the system as a flow graph, the overall generating function is obtained, and expressions for the mean and standard deviation of the acquisition time are determined. The system is examined to determine what effect changing the codelength and the number of sequence detection circuits has on the acquisition time statistics.

The second application of the parallel detection scheme is for a frequency hopping (FH) spread spectrum system. With slight modifications, the expressions, obtained for the mean and standard

deviation of the acquisition time from the DS system, are applied to the FH system. The effect of multi-user interference, jamming, deployment, and sequence detection circuitry on the acquisition time is analyzed, and the results presented.

The computed results allow for the study of the interaction of many of the design parameters, such as codelength, number of SDC systems, acquisition threshold limits, number of potential interferers, and jamming levels.

A summary and concluding remarks are made about the effectiveness of the new multi-SDC acquisition technique, in reducing mean acquisition time, for both direct sequence and frequency hopped SS signals.

ACKNOWLEDGEMENT

I wish to express my gratitude and appreciation to my supervisor, Dr. A.K. El-Hakeem, for his continuous encouragement, guidance, and friendship during the course of my graduate studies.

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LIST OF SYMBOLS AND ABBREVIATIONS

SS	Spread Spectrum
DS	Direct Sequence
FH	Frequency Hopping
PN	Pseudonoise
SDC	Sequence Detection Circuits
LFSR	Linear Feedback Shift Register
R	Autocorrelation of Two Pseudonoise Sequences
p	The Number of Bits in a Maximal Length Sequence
R'	Normalized Autocorrelation of Two Pseudonoise Sequences
S(f)	Power Spectrum of the Waveform with Normalized Autocorrelation
RF	Radio Frequency
R_{PN}	Bit Rate of the Pseudonoise Code Sequence
R_{INFO}	Bit Rate of the Information (Data)
G_P	Processing Gain
s	Number of Registers in a Linear Feedback Shift Register
J	Number of Channels the Interferer Can Hit in a Frequency Hopping System
τ	Integration Time for a Serial Search Acquisition Scheme
TADs	Tapped Analog Delays
SDC	Sequence Detection Circuits
ℓ	Codelength for the Direct Sequence Acquisition Code
n	Number of Tapped Analog Delays
M	Number of Taps per Tapped Analog-Delay
P_{align}	Probability of Alignment
λ_m	Probability of Miss

μ_d	Probability of Detection
λ_r	Probability of Recognizing Misalignment
μ_f	Probability of False Alarm
z	Delay Associated with False Acquisition
P_{d_i}	Probability that the Correctly Peaked i^{th} Tapped Analog Delay is Detected
P_{f_i}	Probability that the i^{th} Tapped Analog Delay will Falsely Peak
α_i	Probability that the Correctly Peaked i^{th} Tapped Analog Delay is Detected
ε_i	Probability that the Correctly Peaked i^{th} Tapped Analog Delay is not Detected
β_i	Probability that the i^{th} Tapped Analog Delay will Falsely Peak
γ_i	Probability that the i^{th} Tapped Analog Delay will not Falsely Peak
ℓ'	Total Number of Available Frequencies for Frequency Hopping System
M'	Number of Frequencies per Matched Filter Bank
n'	Total Number of Matched Filter Banks
P_b	Probability of Bit Error
E_{jk}	Event That j Interferers Use the Transmission Channel and k Interferers use One of the Two Adjacent Channels
$P(E_{jk})$	Probability of the Event E_{jk} Occuring
$P_{bE}(j,k)$	Probability of Bit Error Given E_{jk}
N	Total Number of Potentially Interfering Hoppers in Networks
d	Probability that an Interferer is Emmiting Power During a Bit Interval.
P_L	Lower Limit of Bit Error Probability

P_U	Upper Limit of Bit Error Probability
j	Interferers that use the Transmission Channel
k	Interferers that use one of the Adjacent Channels
L	Limit Chosen to Reduce the Integration in Evaluating $P_{bE}(j,k)$
X_i	Ratio of Power From the i^{th} interferer Using the Transmission Channel.
z_i	Ratio of Power From the i^{th} Interferer Using One of the Two Adjacent Channels.
K	Ratio of Power From an Adjacent Channel interferer to the Corresponding Power that would be produced if the interferer were using the Transmission Channel
y_i	Variable Change Used for Simplicity in Evaluating $P_{bE}(j,k)$
P(x')	Equivalent Expression to $P(\sum_{i=1}^{j} X_i + K\sum_{i=1}^{k} y_i)$ Used For Simplicity
P(E)	Probability of Bit Error
ξ	Variable Depending on the Demodulation Characteristics
R_s	Power in the Desired Signal
N_T	Total Noise Power
N_{th}	Thermal and Background Noise Power
N_j	Jamming Power
$R_s x'$	Summation of Power Resulting From Other Users in the Network
$S_0(x')$	Probability or Bit Error Given x' and the Event That the Transmission Channel is Not Jammed
$S_1(x')$	Probability or Bit Error Given x' and the Event That the Transmission Channel is Jammed
$f(x_i)$	Probability Density Function for an Interference-to-Signal Ratio given that the Interference Enters the Transmission Channel.

$f(y_1)$	Probability Density Function for an Interference-to-signal Ratio Given that the Interference Enters One of the Adjacent Channels.
R_0	Maximum Outer Radial Distance the Interferer can be from the Receiver to Cause Significant Interference.
R_I	Minimum Inner Radial Distance Between the Receiver and Interferer
D	Distance Between the Transmitter and Receiver
r	Distance Between the Interferer and Receiver
$f(R_0)$	Probability Density Evaluated at a Radial Distance of R_0
$f(R_I)$	Probability Density Evaluated at a Radial Distance of R_I
U	Interference-to-Signal Ratio

TO MY FAMILY AND FRIENDS

CHAPTER 1

INTRODUCTION

1.1 General

Although spread spectrum, (SS) communication is used primarly for military applications, there is an increasing interest in the commercial aspects of SS communication, especially in mobile communication networks. By spreading the spectrum, several desirable results can be obtained:

(i) Code division multiplexing allows several users in the network,

(ii) Interference rejection capabilities,

(iii) Anti-jamming properties,

(iv) High resolution ranging.

Clearly, the benefits are overwhelming.

Many aspects of spread spectrum research are currently under investigation, with a large portion of the work focusing on signal acquisition. In this thesis, a new spread spectrum signal acquisition technique is proposed, and examined under various transmission schemes.

The reader is encouraged to read the remainder of Chapter 1, for background into some areas of spread spectrum communication. However, a reader with sufficient knowledge in the area of spread spectrum communication can proceed to Chapter 2 without any loss of understanding.

1.2 Pseudonoise (PN) Sequence Generation

In order for a spread spectrum system to obtain the benefits outlined in Section 1.1, the spectrum of the signal has to be spread at the transmitter, and then despread at the receiver. The spreading effect is achieved by a random sequence and various modulation techniques. However, if a purely random sequence is used to spread the signal, recovery (synchronization) will be very unlikely. Thus, a pseudorandom or pseudonoise (PN) sequence is used to spread the spectrum of the signal. The pseudorandom sequence satisfies the following properties.

(i) the sequences are simple to generate,

(ii) the sequences have randomness properties,

(iii) the sequences have long periods,

(iv) the sequences are difficult to reconstruct from a short segment.

Often a linear feedback shift register (LFSR) is employed to generate the pseudorandom sequence. The LFSR exhibits most of the above properties. With a LFSR it is possible to generate codes of various length, however, for spectrum spreading we are primarily concerned with maximal length sequences having a period of 2^s-1 (where s is the number of stages in the LFSR). If the feedback connection is chosen properly, to generate a maximal length sequence, the following properties will result

(a) in one complete period, the number of ONES will differ from the number of ZEROS by 1.

(b) in any period, among the runs of consecutive ONES and ZEROS, one half of the runs will be of length one (for both ONES and ZEROS), one quarter of the runs are of length two (for both ONES and ZEROS), one eigth of the runs are of length 3 (for both ONES and ZEROS), and so on.

(c) in one complete period, a bit by bit comparison of a sequence and any shifted version of itself, will result in the number of agreements differing from the number of disagreements by at most 1.

Property (c) is extremely important in spread spectrum communication. We note for two binary sequences, that the correlation can be defined by taking bit-by-bit MOD-2 sum of the sequences, and comparing the number of agreements and disagreements between the two. It can be shown, [1] that the auto-correlation is

$$R = \text{\# of ZEROS} - \text{\# of ONES} \tag{1.1}$$

where ZEROS refer to agreements between the two codes and ONES refer to disagreements between the two codes. Further, expressing the auto-correlation in a normalized form yields

$$R' = \begin{cases} -1/p & \text{for all shifts} \neq 0 \\ 1 & \text{for shift} = 0 \end{cases} \tag{1.2}$$

where $p = 2^s-1$ and p is the number of bits in a codelength and R' refers to the normalized auto-correlation. The auto-correlation is shown in Fig. 1.1. By taking the Fourier transform of the auto-correlation, it can be shown [2], that the power spectrum, S(f), of the auto-correlation in (1.2) is,

$$S(f) = \frac{p+1}{p^2} \sum_{\substack{n=-\infty \\ n\neq 0}}^{n=\infty} \operatorname{sinc}^2\left(\frac{n}{p}\right) \delta\left(f-\frac{n}{pT}\right) + \frac{1}{p^2}\delta(f) \qquad (1.3)$$

where p is the number of bits in the codelength and T is the bit duration. The spectrum is shown in Fig. 1.2.

Cross-correlation is also an important consideration in choosing a PN sequence. However, there is no simple method for predicting the cross-correlation between two codes.

1.3 Direct Sequence (DS) Spread Spectrum Systems

Direct sequence (DS) is a common technique used in spread spectrum (SS) communication, primarily because of the relative ease in implementing these systems, over other SS technqiues. Direct sequence modulation is precisely the modulation of a carrier by a code sequence (usually a PN sequence), as shown in Fig. 1.3. The baseband information is digitized and then modulated with the carrier frequency ω_0, to form $S_1(t)$. The final transmitted spread spectrum signal, $S'(t)$, is generated by modulating the spreading binary code, or PN code with the signal $S_1(t)$. Biphase balanced modulation (using a PSK modulator), is most commonly used for signal modulation. In biphase modulation , the carrier is transmitted with one phase when the code sequence is a "1" and by a phase shift of 180° when the code is a "0". Biphase balanced modulation is prefered for several reasons [3];

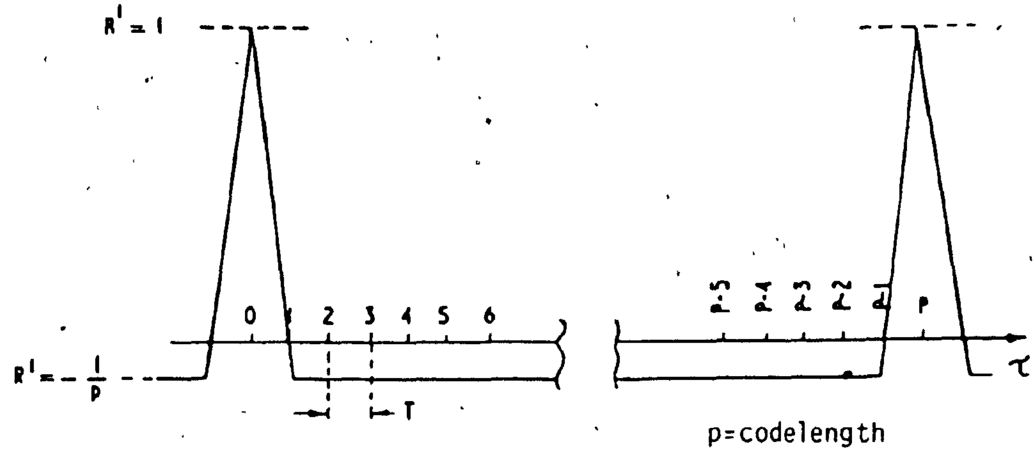


FIG. 1.1: NORMALIZED AUTOCORRELATION FUNCTION WITH BIT DURATION T

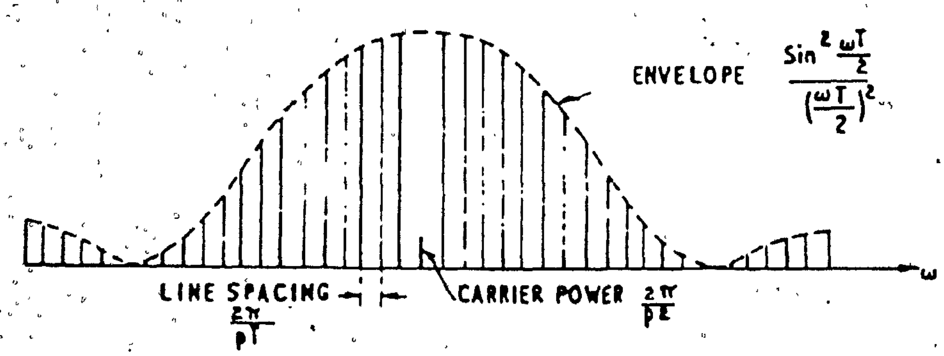


FIG 1.2: POWER SPECTRUM OF A BINARY PN WAVEFORM

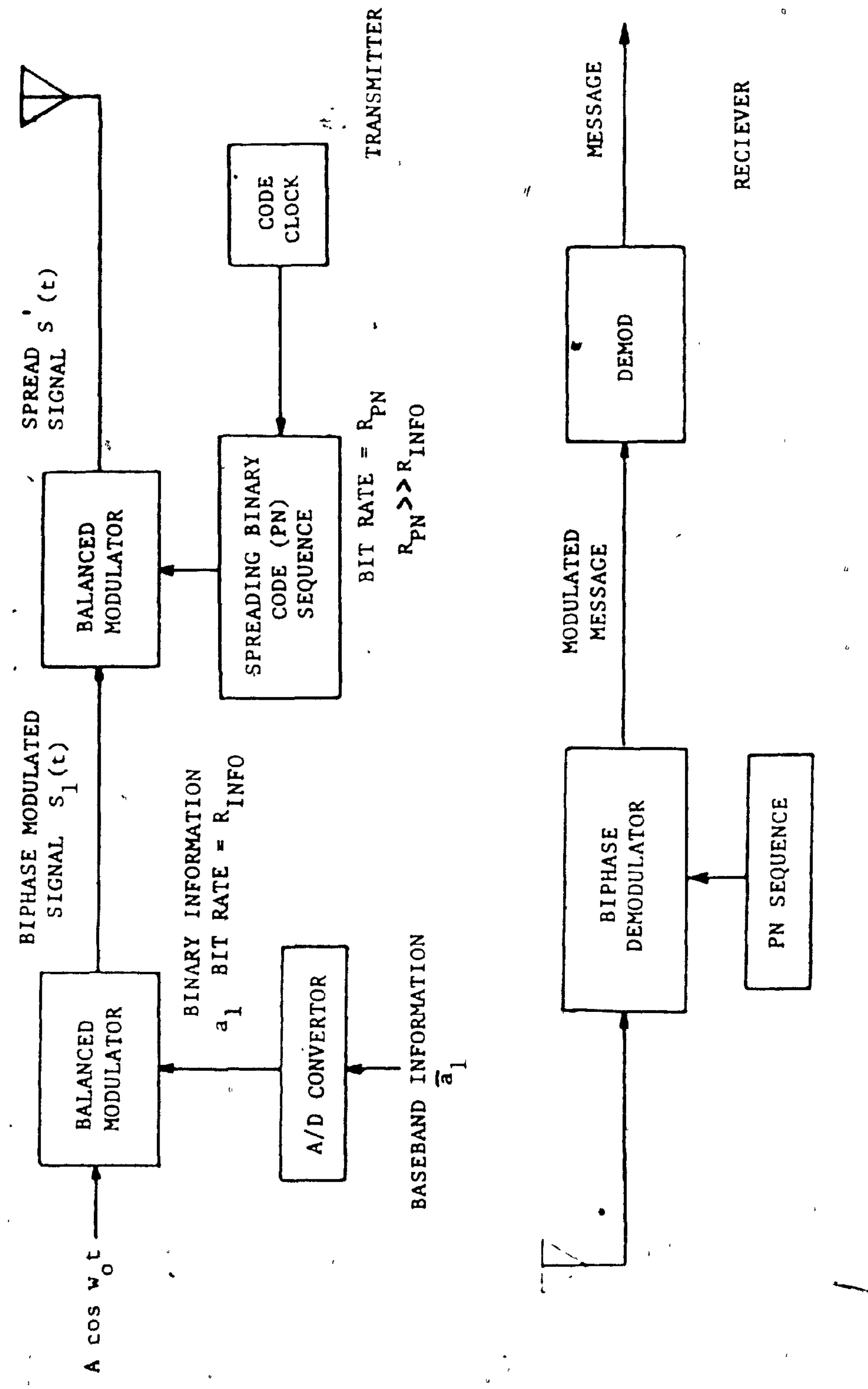


FIG. 1.3: BASIC DIRECT SEQUENCE SPREAD SPECTRUM SYSTEM

(i) it is difficult to detect the suppressed carrier produced when biphase balanced modulation is used

(ii) more power can be used to send useful information since the carrier is suppressed.

(iii) efficiency of the transmitted power is maximized for the bandwidth used, since the signal has a constant envelope.

(iv) the simplicity of the biphase modulator (consisting of one pair of transformers and a few diodes).

More commonly, transmission is achieved by using one balanced modulator to modulate the MOD-2 sum of the binary information and the PN code sequence, as shown in Figs. (1.4 and 1.5). The bits of the sequence formed by the MOD-2 addition of the binary information and the PN code sequence are referred to as "chips".

At the receiver, Fig. 1.6, the received signal passes through a wideband filter and the signal is then synchronized, (synchronization to be discussed later). A locally generated PN code sequence is used to demodulate the received signal in order to recover the modulated message. Further demodulation allows recovery of the data. Without knowledge of the PN code sequence, it is impossible to recover the transmitted signal.

Interference from users using a different PN code sequence will be minimal, Fig. 1.7, provided the codes are chosen to minimize the cross-correlation properties (previously mentioned in section 1.2).

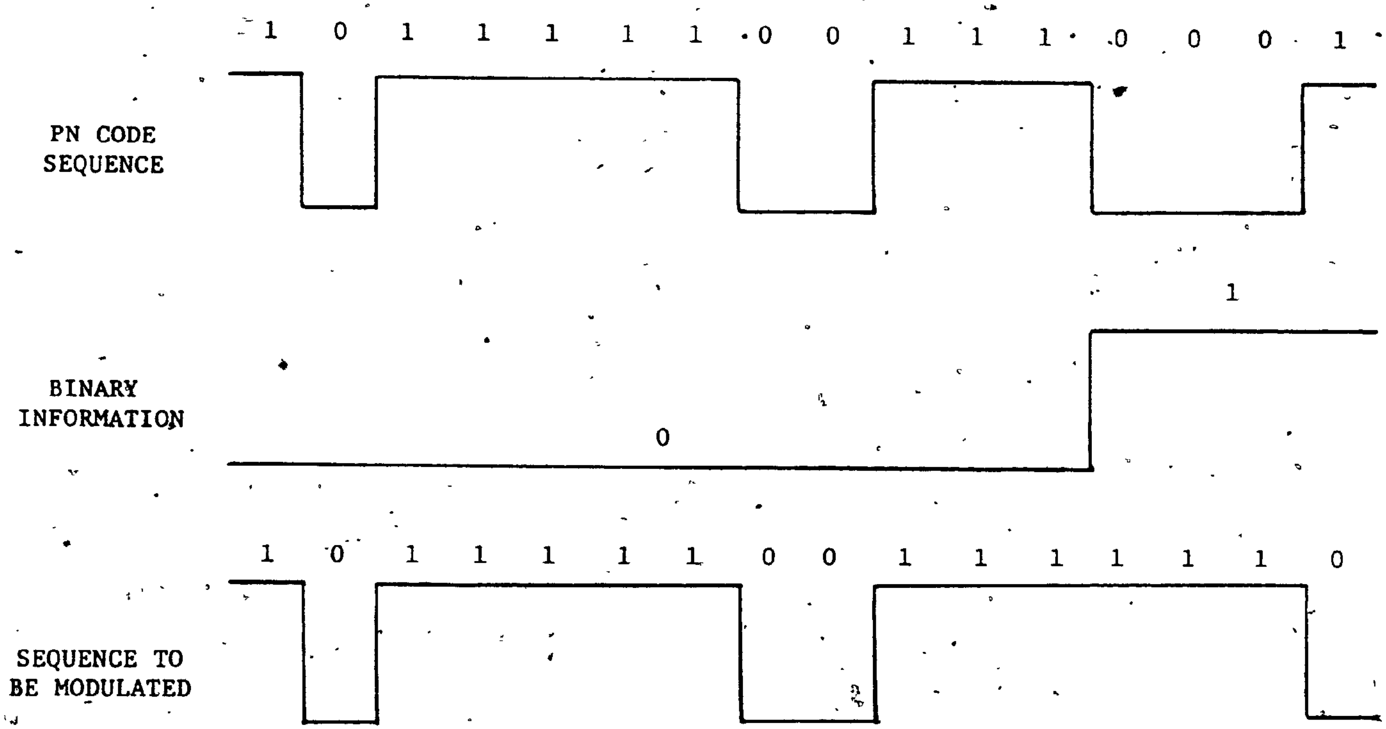


FIG. 1.4: MODULO-2 SUM OF BINARY INFORMATION AND THE PN CODE BEFORE MODULATION REPRESENTING THE FORMATION OF A "CHIP"

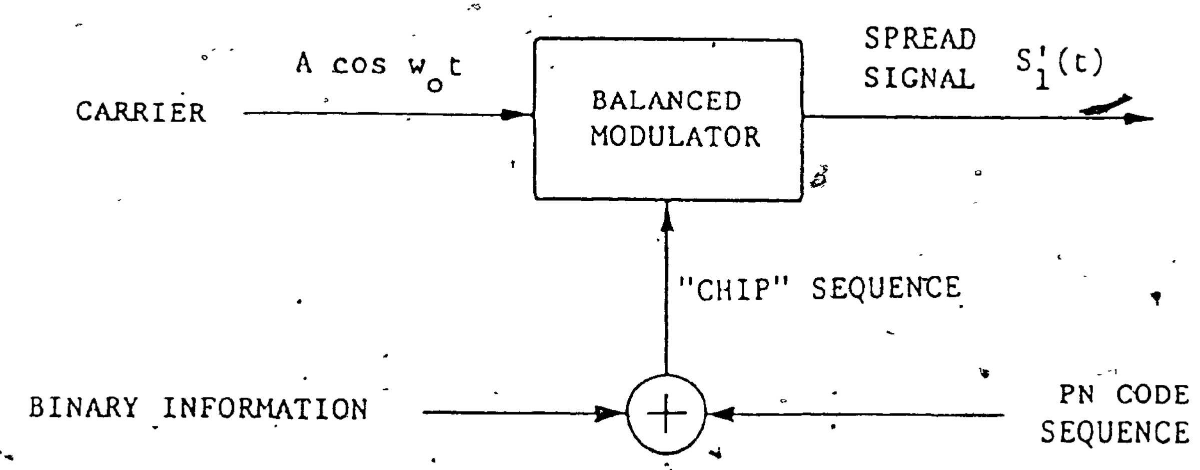


FIG. 1.5: SIMPLIFIED DIRECT SEQUENCE TRANSMITTER CONFIGURATION

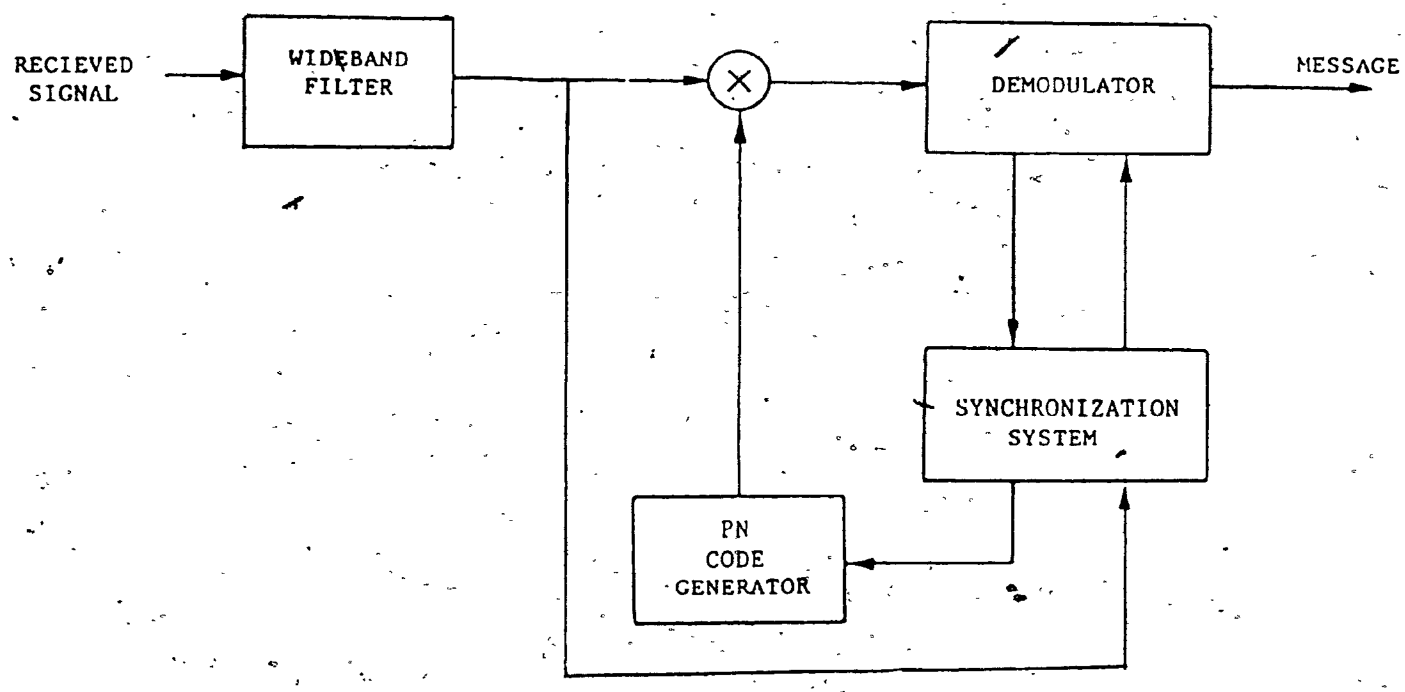
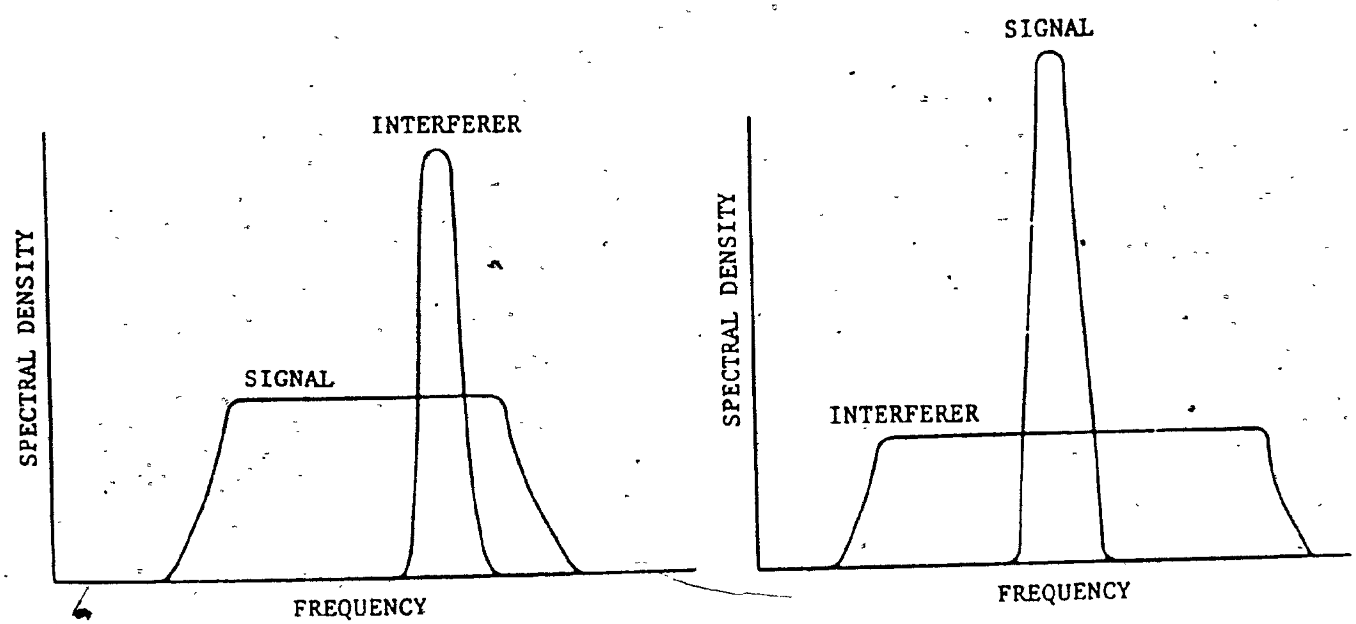


FIG. 1.6: POSSIBLE DIRECT SEQUENCE RECIEVER



(a) RECIEVED SIGNAL SPECTRA BEFORE DEMODULATION

(b) SIGNAL SPECTRA AFTER DEMODULATION WITH THE LOCALLY GENERATED PN CODE SEQUENCE

FIG. 1.7: EFFECT OF INTERFERER WITH ALTERNATE PN SEQUENCE

A key consideration in DS systems is the radio frequency (RF) bandwidth. The DS systems bandwidth is related to the code bit rate, and reflects itself in the processing gain, G_p, (the processing gain is a function of the RF bandwidth and the information bit rate). Quite often it is necessary to put limitations on the bandwidth, thereby reducing the processing gain of the system.

It is assumed, that the RF bandwidth is typically that of the main lobe of the direct sequence spectrum, or twice the rate of the clock used to generate the spreading code. As an example consider a system with a code clock rate or 15 Mbps. and an information rate of 1.5 Kbps, the processing gain would be $(2\text{x}15\text{x}10^6)/(1.5\text{x}10^3) = 2\text{x}10^4$ or 43db. It would appear that the processing gain can be infinite by choosing a sufficiently fast code clock, thereby increasing the RF bandwidth. However, physical limitations on current technology restrict code rates to 200-300 Mbps. As well, increasing the code rate would require a lower bit error rate. Presently, high-speed logic circuits, have a large sensitivity to noise making them susceptible to errors, as well, they consume large amounts of current and dissipate a large quantity of power. A further increase of processing gain is possible by decreasing the information (data) bit rate. However, reducing the data rate too much (10 bps) can cause significant errors, either through local oscillator phase noise or instability in the propagation medium [3]. Current limitations restrict direct sequence processing gain to about 75dB.

1.4 Frequency Hopping (FH) Spread Spectrum Systems

Frequency Hopping (FH) is another popular technique used in spread spectrum systems. The FH system is primarily composed of a frequency synthesizer, and a code generator, used to select a pseudorandom frequency pattern, Fig. 1.8.

Theoretically, the frequency spectrum for a FH system should be a band of sharp spikes, Fig. 1.9. However, the resulting spectrum is usually composed of desired frequencies, sidebands resulting from hopping, and other extraneous frequencies.

At the receiver a locally generated PN code sequence, is used to create a replica of the frequency pattern produced by the transmitter. After synchronization is achieved, the wideband filtered received signal is demodulated with the locally generated frequency pattern, to recover the modulated information. Further demodulation leads to recovery of the transmitted information.

Any signal that has an alternate PN code sequence is rejected as noise, similar to the direct sequence (DS) system. In general, DS and FH are identical. However, the two systems differ in some ways.

Processing gain for the FH system is identical to the DS system

$$G_p = \frac{BW_{RF}}{R_{INFO}} \tag{1.4}$$

where the frequencies (channels) are contiguous. If the frequencies (channels) are not contiguous, the processing gain is more readily defined as

G_p = the total number of available frequency choices = ℓ'

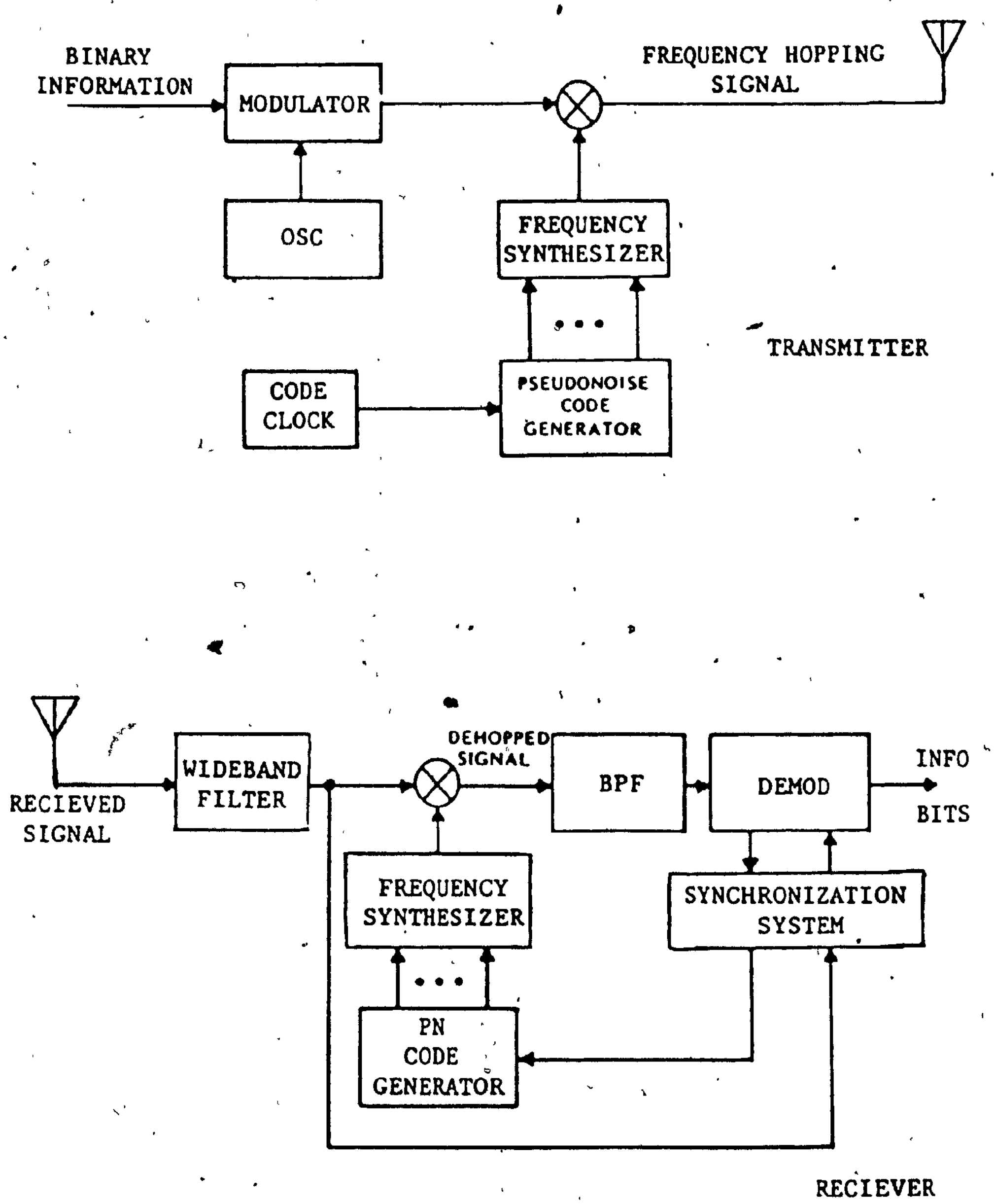


FIG. 1.8: BASIC FREQUENCY HOPPING SYSTEM

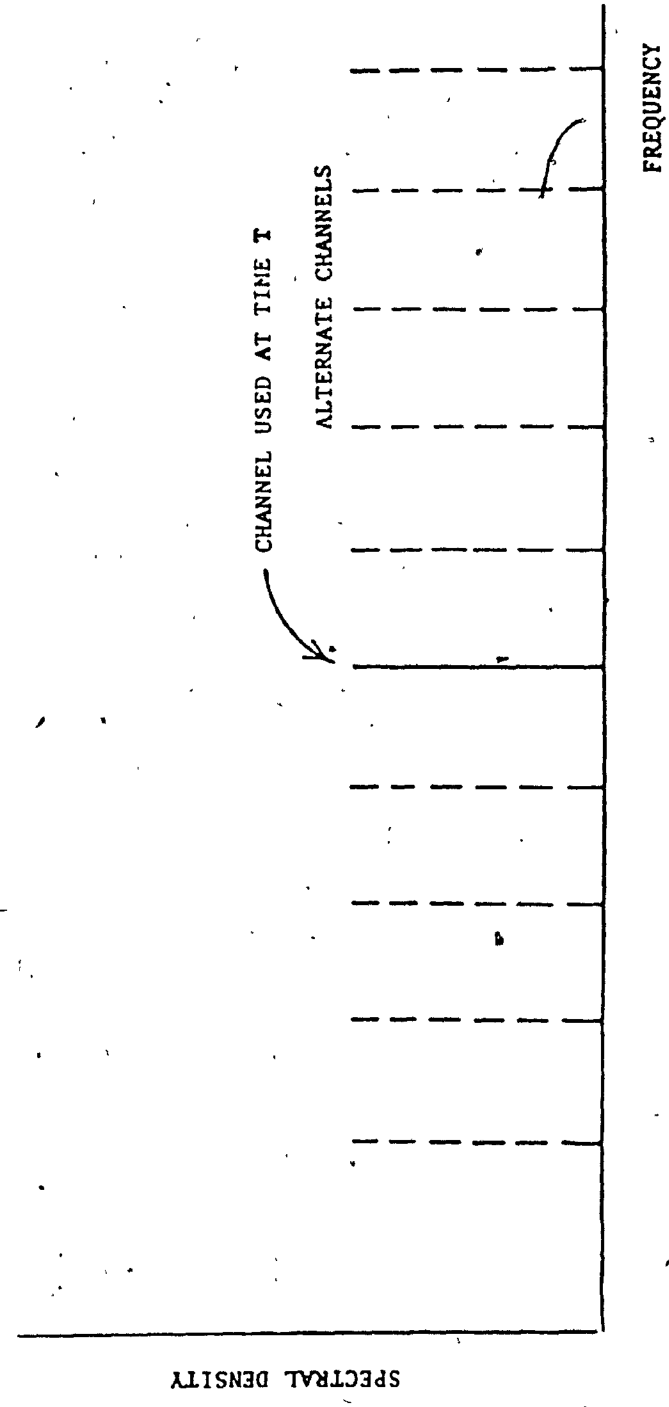


FIG. 1.9: IDEAL SPECTRUM OF A FH SYSTEM

The FH system, should then have a large number of available frequencies for increased processing gain. The number of frequencies required however, depends on the tolerable error rate and effects of jamming (to be discussed later). For a FH system without any error correction capability, the error rate is J/ℓ', where J is the number of channels the interferer can hit, and ℓ' is the number of available channels to choose from.

If binary FSK modulation, Fig. 1.10, is used for the FH system, each code bit (symbol) is transmitted using one of two frequencies. One frequency represents a logical one (mark), and the other, a logical zero (space). With each hop the pair of frequencies used for transmission changes. The channel (frequency) that is used to transmit the code bit is the transmission channel. The channel (frequency) that would be occupied if the logical state of the transmitted bit was reversed, is the complementary channel. From [3], it is shown that binary FSK has an error rate of

$$P_e = \sum_{x=r}^{c} \binom{c}{r} \nu^x q^{c-x} \tag{1.5}$$

where

ν = error probability on a single trial,
= J/ℓ',

J = number of jammed channels,

ℓ' = number of channels available to frequency hopper,

q = probability of no error for a single trial
= $1-\nu$,

c = number of chips (frequencies) sent per information bit,

r = number of incorrect chip decisions necessary to cause a bit error.

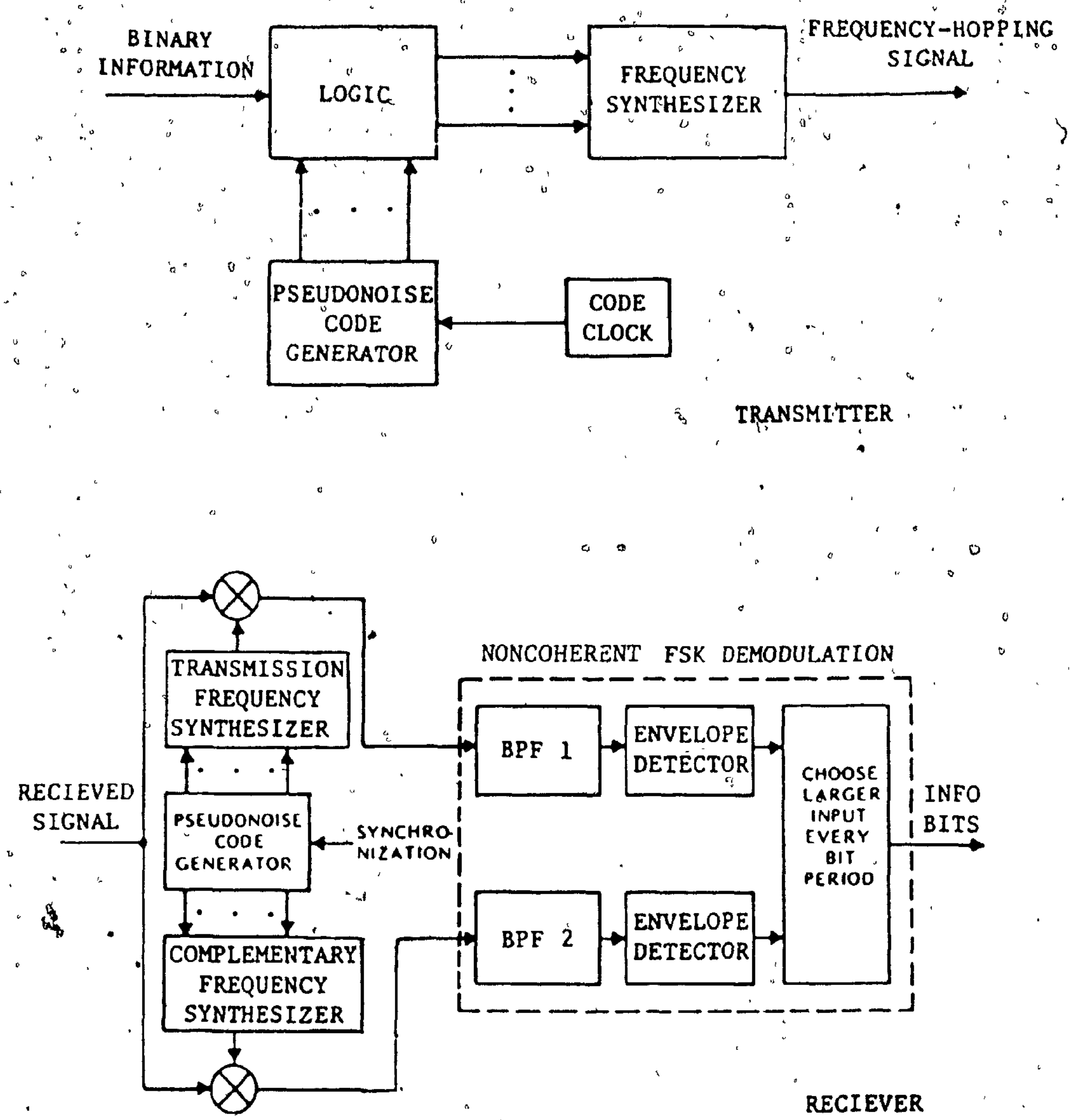


FIG. 1.10: FREQUENCY HOPPING SYSTEM WITH FSK MODULATION

When the interferer causes the power in the "space" channel to exceed the power of the "mark" channel, a chip error will result. A large increase in chip error, will make the error code increase as well. It is evident from (1.5) that increasing the number of chips sent per bit, c, will result in a lower bit error rate. Thus, the hopping rate, and RF bandwidth must increase in proportion. Restricting the number of frequencies available or the bandwidth will result in trade offs between the chip per information bit rate and the number of available frequencies. To take full advantage of a frequency hopping system, an error correcting code should be used to maximize the number of tolerable errors, r.

Frequency hopping rates usually take one of two forms. Slow frequency-hopping is the transmission of two or more symbols in the time between each frequency hop. While fast-frequency hopping is the transmission of one symbol between each frequency hop.

In slow frequency hopping, the c bits of the code word must be interleaved over c hopping periods to ensure each bit is transmitted in a different channel (i.e. less susceptible to jamming). The block code used for slow frequency hopping with bit interleaving, will provide correction capabilities equal to that of a fast frequency hopping system with the same block code.

1.5 Spread Spectrum Signal Acquisition

The most important aspect of a spread spectrum system is its ability to synchronize the received signal with the locally generated pseudonoise code. The problem of signal synchronization results from various conditions, such as Doppler effect, clock discrepencies, etc.

All of the aforementioned conditions produce an uncertainty at the frequency and time of the incoming signal commonly referred to as the uncertainty region. Code synchronization is necessary since it allows the receiver to despread the received signal and recover the modulated information (data).

Synchronization is divided into two components, (i) acquisition or course alignment, to bring the receiver code to within one chip of the transmitted code, (ii) and tracking or fine alignment, in which errors of alignment are further reduced or at least maintained within one chip. However, only acquisition will be considered throughout this section and the remainder of the thesis. Quite often, acquisition time is used as a performance measure of a spread spectrum acquisition technique.

Several techniques are available for spread spectrum signal acquisition, although two schemes are primarily used; (i) serial search or "sliding correlator", and (ii) matched filtering.

In a serial search acquisition scheme, a search is made over all relative phase shifts between the received signal and the locally generated PN code sequence. At the receiver the locally generated PN code sequence is operated at a rate different than that of the transmitter code. The product of the locally generated PN code and the received signal is then bandpass filtered and is integrated over a fixed period (dwell time), τ, to determine the power contained in the signal. When the locally generated code aligns itself with the received signal, a preset power threshold will be exceeded, and the receiver will move to tracking. This method of acquisition is referred to as single dwell serial search.

A faster acquisition time can be obtained by using a multiple dwell serial search scheme. This scheme is similar to the single-dwell acquisition technique, except integration is performed over several short dwell periods. Each short integration will continue in succession as long as the threshold is exceeded. If the power contained in the signal falls below the threshold, the integration process will return to the first integration.

Serial search acquisition can be applied to both direct sequence (DS), and frequency hopped (FH) spread spectrum systems, Figs. (1.11 a,b).

Matched filtering is another scheme that can be used for both direct sequence and frequency hopped spread spectrum signal acquisition. For a DS spread spectrum system a delay line is commonly used as a matched filter, Fig. 1.12(a). The binary representation of the received signal is shifted into the delay line and multiplied chip-by-chip by the stored PN code sequence. Acquisition is declared when integration (summing) over M chips produces a value in excess of a preset threshold, where M is the number of taps on the delay line.

Matched filtering for a FH spread spectrum system varies from that of a DS system. The filter is designed to match to a specific frequency hopped pattern, Fig. 1.12(b). In the K^{th} arm of the matched filter, the received signal is mixed with a tone $f_K = f_i + f_o$. The bandpass filter is centered around f_o, thus if the received frequency is f_i that arm will contribute significantly to the overall summation. The fixed delay allows the uncertainty region to be searched in real time, and provides a sliding window to the arriving signal.

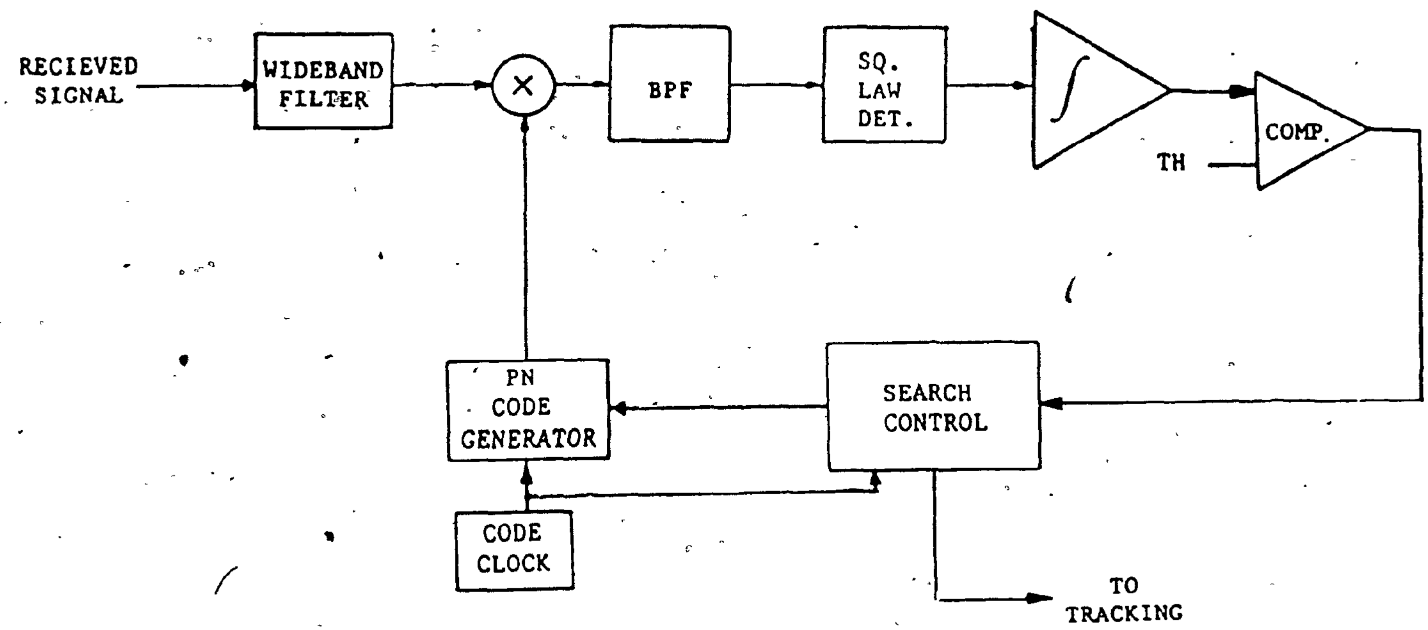


FIG. 1.11(a): SERIAL SEARCH ACQUISITION SCHEME FOR A DS SPREAD SPECTRUM SYSTEM

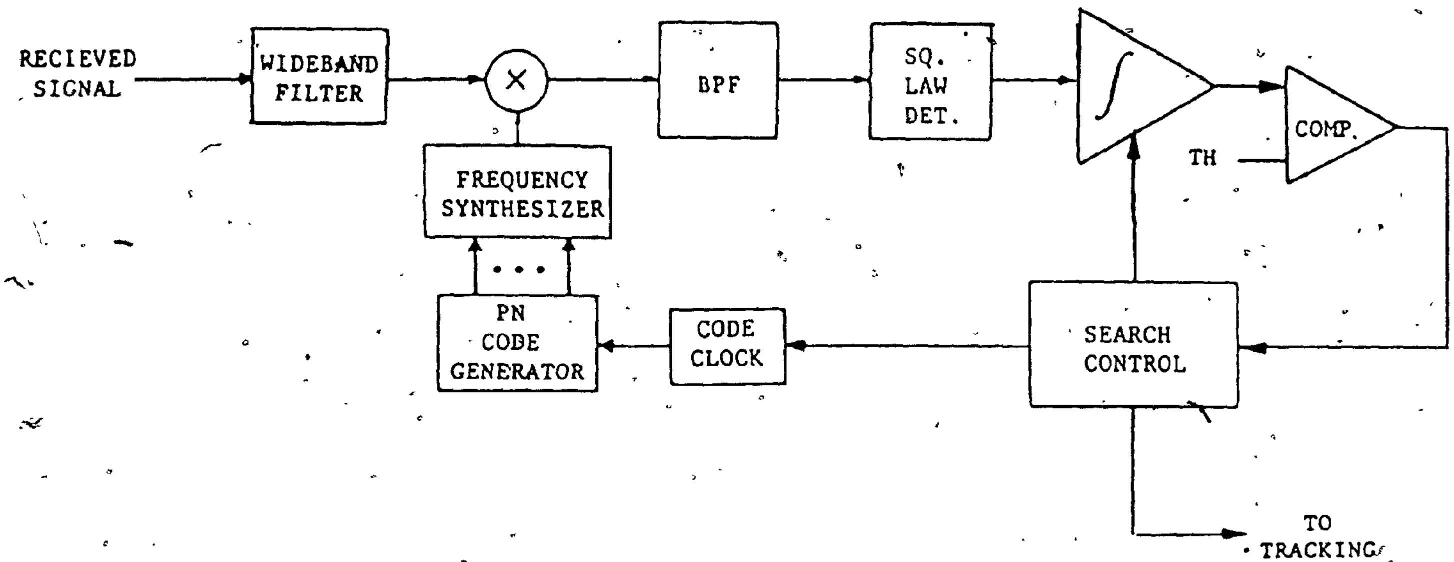


FIG. 1.11(b): SERIAL SEARCH ACQUISITION SCHEME FOR A FH SPREAD SPECTRUM SYSTEM

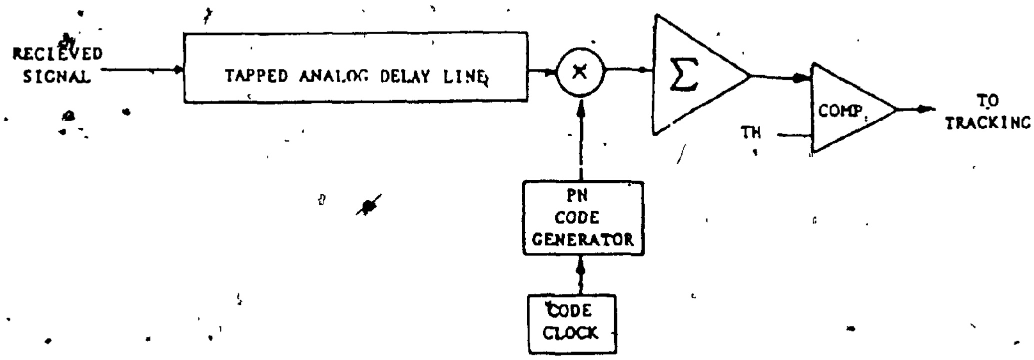


FIG. 1.12(a): MATCHED FILTER ACQUISITION SCHEME FOR A DS SPREAD SPECTRUM SYSTEM

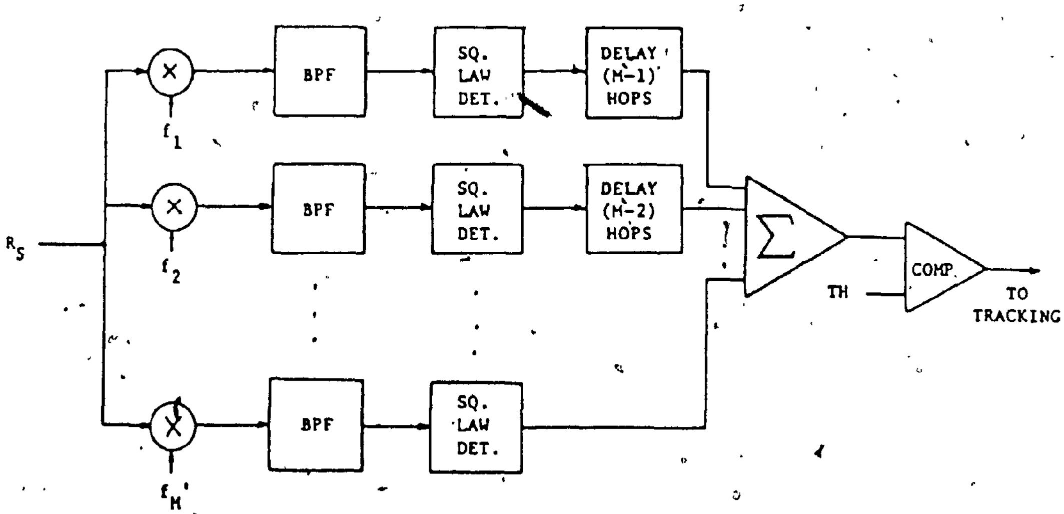


FIG. 1.12(b): MATCHED FILTER ACQUISITION SCHEME FOR A FH SPREAD SPECTRUM SYSTEM

1.6 Jamming

Jamming is the deliberate interference imposed on a transmitted signal. There are two basic ways by which the jammer can disrupt the communication systems. Either the synchronization system can be effected or the number of bit errors can be increased beyond correction capability. Several jamming techniques can be implemented to achieve either form of jamming mentioned above.

In a direct sequence (DS) spread spectrum system, the ability to reject jamming depends upon the integrity of the pseudonoise (PN) code. If the jammer is unable to access the PN code, then tone jamming at the center frequency of the PN spectrum is the most effective form of jamming. When the carrier frequency and chip rate of the PN code can be approximated, the jammer can achieve effective forms of jamming, by generating an alternate PN code at that particular carrier frequency and chip rate. Finally, it is important to note, that a DS spread spectrum system becomes more susceptible to jamming when the PN codes used are permanent (non-programmable).

In a frequency-hopped (FH) spread spectrum system, partial-band jamming and repeater jamming are among the most popular jamming techniques.

The partial band jamming technique is commonly used if the jammer's power is less than the product of the transmitter power and the number of hopping channels. In partial band jamming, the jammer concentrates his jamming power in only a portion of the overall hopping band. If the jamming power is limited, the jammer must reduce the jamming strength per channel in order to increase the total number of jammed channels, (i.e. jam a larger portion of the hopping band). The rate of bit errors is greatest, and hence the jamming most effective, in a partial-band jamming system, when the jamming power in a jammed channel is approximately equal to the signal power.

Another common form of jamming is repeater jamming. Repeater jamming is most effective in a slow frequency hopping system, because of the way jamming result. In repeater jamming, the transmitted signal is captured and analyzed to determine the hopping frequency being used. The analyzed signal is then jammed and transmitted, by the jammer, with the same hopping frequency. The jammed signal must reach the system receiver, before the receiver hops to a new set of channel frequencies.

1.7 Spectral Splatter

It should be noted, that the acquisition time will be largely increased, if the bit error is also increased. Both jamming and an effect refered to as spectral splatter, can produce a significant increase in bit error.

Spectral splatter, refers to the interfering spectrum overlap in adjacent channels, produced by a transmitted pulse. The effect of spectral splatter on bit error is dependent on several factors: hopping rate, frequency seperation between channels, transmitter-receiver deployment, and the transmitted signals spectrum.

In a fast-frequency hopped system, the spectrum of the transmitted signal and the channel availability are strongly effected by the hopping rate. For a slow-frequency hopped system, the switching time (i.e. the time interval when the frequency synthesizer is not operating between frequency hops), effects the signal spectrum. It is possible to

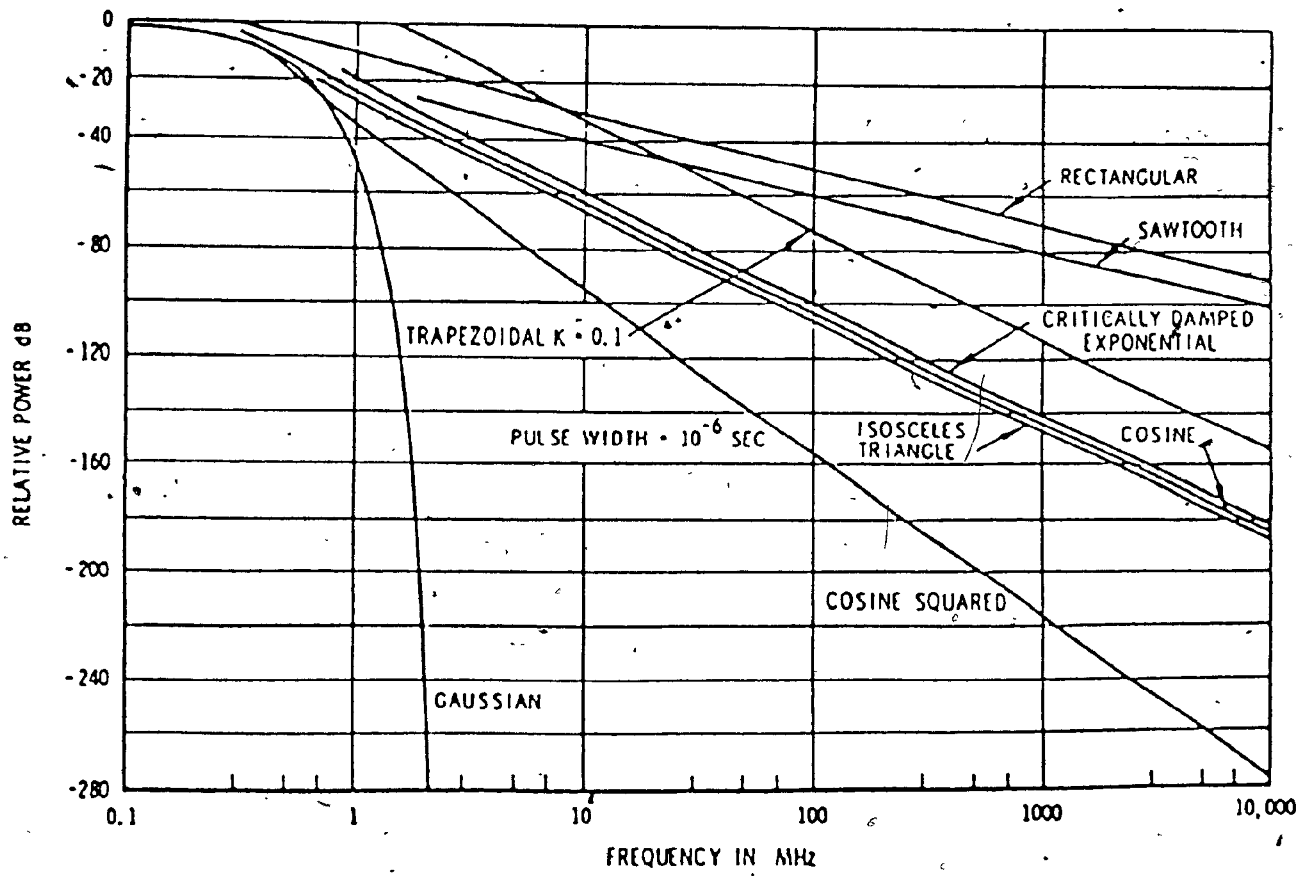


FIG. 1.13: SPECTRAL ENVELOPE FOR VARIOUS WAVEFORM PULSES

increase the frequency seperation, thereby reducing the splatter effect. However, by reducing the number of available channels, (i.e. increasing frequency seperation) repetition of the same transmission channel results more frequently, making the user more susceptible to interference (jamming).

Various pulses can reduce the effects of spectral splatter. Figure 1.13, from [4], shows the relative spectrum splatter effect for various transmission pulses. It is evident, that the spectral sidelobes are strongly reduced for Gaussian (a Gaussian approximation can be generated, since a perfect Gaussian shaped pulse is impossible to generate) and squared cosine (or raised cosine) pulses. Ideally, transmission of a constant envelope signal is desired, however, with pulse shaping (approximate Gaussian and raised cosine pulses), the nonlinearities of the final transmitter power amplifier and the propagation channel produce added spectral sidelobes. With the use of continuous phase frequency shift keying, CPFSK, the spectral sidelobes can be minimized (ie. sharp roll-off characteristics).

1.8 Review of Some Recent Work on Signal Acquisition

The problem of signal acquisition has been the focus of much research work in the area of spread spectrum communications. A variety of signal acquisition schemes for both direct sequence and frequency hopping spread spectrum systems, have been proposed and analyzed. Some of the more relevant signal acquisition research work [5 - 16], will be examined in this section.

Much of the work in spread spectrum signal acquisition tends to give a biased preference to one technique over the other, (i.e. direct sequence versus frequency hopping). However, in [5], an unbiased comparison between frequency hopping and direct spread pseudonoise (direct sequence) techniques are based on several parameters: processing gain, signal-to-noise performance, low probability of intercept, ranging, antenna tracking and propagation and acquisition. Although, we will only concentrate on the acquisition aspect of the comparison, the effect of the other parameters will also be considered. The acquisition performance comparison is based on the assumption that the amount of timing offset between the transmitted signal and stored reference code is identical for both techniques, and the processing gain for both techniques is also identical. Based on this premise, the uncertainty region for a frequency hopping technique is somewhat smaller than for the direct spread PN technique, and the frequency hopping technique tends to provide acquisition in a shorter period of time. However, this frequency hopping advantage exists only when a serial search method is used to acquire the signal. If matched filtering is used for acquisition, neither technique has an added advantage. Spellman also demonstrates how the direct spread PN technique is advantageous in, providing protection against processing gain attacks, providing good signal-to-noise performance and has a low probability of intercept, amongst other benefits. The frequency hop technique provides an extremely large nominal processing gain, and the ability to provide short acquisition times.

A more practical approach to the spread spectrum signal acquisition problem is given in [6] by Rappaport and Grieco. Rappaport et al provide insight into some of the current technologies available for spread spectrum systems and the application of the technology to direct sequence, frequency hopping, and hybrid acquisition techniques. Several schemes for signal acquisition are outlined, including matched filtering serial search, variable dwell time, estimation and two-level schemes. The choice of scheme is dependent on several factors, but primarily on the spreading modulation used and the anticipated application for the system. Matched filters, correlators, and convolvers are the primary devices used in these schemes, with four major technologies tending to dominate for these devices: Surface Acoustic Wave (SAW) matched filters; Charged-Coupled Device (CCD) matched filters; Digital matched filters and SAW monolithic convolvers. Rappaport et al, conclude that SAW convolvers will find significant application where high frequencies (60 MHz) are required, while digital matched filters will find application in systems operating below 25 MHz. The SAW matched filter may be used in applications where the operating frequency of the system lies between 25 MHz and 60 MHz, while the future of CCD matched filtering does not look optimistic.

In [7], three frequency hopped spread spectrum signal acquisition techniques are compared on the basis of miss probability, mean search time, and complexity. The performance of serial search, matched filter, and two-level acquisition tchniques is determined for an adverse operating environment. Putman et al [7], found that the serial search scheme provides good detection capability and is easy to implement, but the acquisition time is long, as compared to the matched filter or two-level scheme. Because of its relatively quick acquisition time, the matched filter scheme, although somewhat complex to implement, is prefered for acquisition systems operating in a less adverse environment. The two-level scheme combines the merits of both the serial search and matched filter schemes, and would be best utilized in systems requiring acquisition of a long code in a short period of time.

In [8], Holmes et al, analyzes serial search acquisition techniques with the aid of Markov models representing each acquisition scheme. Holmes et al, develop expressions for the statistical behaviour of the acquisition time for both a single dwell and double dwell PN type serial search acquisition scheme. Consideration is also given to the effect of Doppler on the system performance. The somewhat systematic approach to analysis, of the acquisition technique, lends itself quite easily to other models.

An alternate system peformance expression is derived in [9] by DiCarlo and Weber. Using the generating function derived in [8], DiCarlo and Weber develop an expression for the probability of successful synchronization for a single dwell serial synchronization system. The system performance is evaluated using the probabilty expression, for various dwell times, false alarm probabilties and detection probabilities, thereby providing insight into the tradeoffs involved in minimizing synchronization time.

Quite often, a more rapid acquisition technique, commonly referred to as multiple dwell-time system, is employed at the cost of increased complexity. In [10], DiCarlo and Weber, again apply the analytical techniques of Holmes et al, to determine the mean and variance acquisition time expressions for a multiple dwell serial search scheme. DiCarlo and Weber determined that the improvement in performance for the multiple dwell scheme is significant when two dwells are chosen over a single dwell. However, increasing the number of dwells beyond two only produces a slight reduction in mean acquisition time.

Hopkins [11], analyzes the complete synchronization problem, by developing a relationship between coarse acquisition and fine acquisition, and presenting a technique for determining the mean acquisition time, of a multiple dwell acquisition scheme. A functional representation of the acquisition scheme is proposed, including a delay lock loop system for tracking, a search/lock strategy, modelled as a Markov chain, for synchronization control, and a correlation detector for coarse acquistion. The acquisition technique is demonstrated for a particular system, to reveal performance characteristics under various signal-to-noise ratios. It is interesting to note that Hopkins determined that the mean acquisition time, for the multiple dwell scheme, was effected (increased) more rapidly by a decreasing signal-to-noise ratio with a small false alarm probability, then by a decreasing signal-to-noise ratio with a large false alarm probability.

Another serial search approach used for acquisition, is the variable dwell-time scheme. In [12], Braun presents a numerical approach for computing the performance of this scheme. The acquisition performance of the variable dwell-time scheme is compared to a fixed dwell-time system and an optimum detection algorithm; for a high SNR the multiple dwell technique slightly outperforms the fixed dwell technique, while for a low SNR, the multiple dwell technique nears the performance of an optimum sequential detector.

In [13], Braun analyzes an expanding search acquisition scheme that assumes some a priori information on the PN code approach is available. In the scheme, a search is made over a small assumed uncertainty region, the region is continuously increased until acquisition results. An approximation for the characteristic function expression of the acquistion scheme is compared with results from [9] illustrating the validity of the approximation. As well, an expression for the mean acquisition time is developed for the expanding search scheme, allowing for comparison with other acquisition schemes. A key feature of Braun's analysis is that the reuslts can be applied to both fixed and variable dwell-time systems. We note the expanding search technique is particularly effective for re-acquiring signals that have been temporarily lost.

El-Hakeem et al [14], present and analyze new acquisition schemes for hybrid spread spectrum systems. By using a technique referred to as ASEAT (autoregressive spectral estimation acquisition technique), outlined in [15], the signal acquisition time was significantly reduced for a frequency hopping system with moderate hopping rates and processing gain and a low carrier-to-noise ratio. Further, simulation results indicated that acquisition times for hybrid spread spectrum systems (FH/TH and FH/DS), were much less when the ASEAT was used over other acquisition techniques.

In [16], Baier et al outlines and analyzes a somewhat complex spread spectrum synchronization (acquisition and tracking) scheme in which SAW tapped delay lines are used as matched filters. The technique employs a modulation cancellor to improve the synchronization characteristics. However, simulation and experimentation results indicate that the mean acquisition time increases with the use of the modulation cancellor, and in fact the modulation cancellor should be disabled during the acquisition stage.

Another point of interest that will be examined in this thesis is the effect of multi-user interference on a spread spectrum system. In [17] the problem of multi-user interference on pseudonoise, time division multiple access/PN, synchronous and asynchronous frequency hopping spread spectrum systems is considered. Simulation results revealed that the probability of bit error for each system depended on the location and number of the interferers, and the level of the signal-to-noise ratio. The synchronous FH spread spectrum system was able to tolerate the greatest number of interferers but required more complexity than the other systems.

The acquisition schemes that will be presented in this thesis employ some of the analysis techniques in [5] to [17]. However, it will be shown that the performance of these new spread spectrum acquisition techniques, in terms of mean acquisition time, is better than other proposed acquisition techniques, and requires a minimum of complexity to implement.

1.9 Outline of the Thesis

The thesis is divided into four chapters. Chapter 1, provides some background information into spread spectrum communication theory. As well, some recent papers in spread spectrum signal acquisition are examined.

In Chapter 2, the new direct sequence spread spectrum signal acquisition scheme is proposed, and a flow graph representation of the new acquisition scheme is obtained. The mean and the standard deviation of the acquisition time are determined for the new acquisition technique, under various detection and false alarm probabilities.

In Chapter 3, the new frequency hopping spread spectrum signal acquisition scheme is proposed. The expression for the mean acquisition time is generated by showing the analogy between the acquisition scheme outlined in Chapter 2, and the frequency hopping technique of Chapter 3. Finally, the peformance characteristics are examined for the frequency hopping acquisition technique, under various operating environments.

Chapter 4 contains the summary, concluding remarks and recommendations.

CHAPTER 2

A NEW SIGNAL ACQUISITION TECHNIQUE FOR A DIRECT SEQUENCE (DS) SPREAD SPECTRUM SYSTEM

2.1 Introduction

Several approaches [6] are possible for signal acquisition. The technique presented in this paper is based on the use of tapped analog delays as matched filters for the acquisition of a direct sequence (DS) signal. The acquisition scheme could find application in a multi-user environment, where long acquisition codes are required to incorporate several users. A bank of TADs is used, where each TAD is uniquely matched (by proper weighting) to a specific segment (subdivision) of the overall code. Subdivision of a long code is possible without loss of such desired characteristics as large autocorrelation and small cross correlation, by using Gold codes, for instance as in [2]. For analysis, it is assumed that a code of length ℓ is used for signal acquisition, and the code is composed of n maximal length codes having the desired properties of JPL codes (i.e. large auto-correlation and small cross-correlation properties). Further, each maximal length code sequence is generated by a linear feedback shift register (LFSR), with a unique feedback connection, Fig. 2.1. Thus, a code of length ℓ, is the result of several LFSR's of equal length, each with a unique feedback connection. Since, the code of length ℓ is subdivided into n sections (n maximal length code sequences of equal length), sequence detection circuits (SDC) can be used to detect when all or a portion of the received signal is acquired. Clearly, from Fig. 2.1, the sequence $1 \rightarrow 2 \rightarrow 3 \rightarrow \ldots \rightarrow n$ must occur (i.e. the received signal must match with the

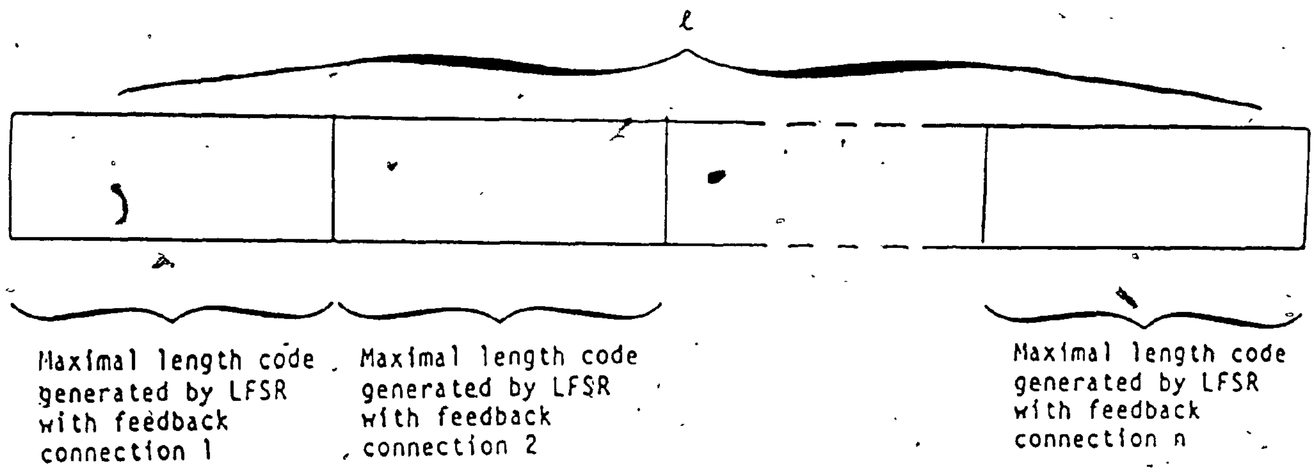


FIG 2.1: REPRESENTATION OF THE CODE USED FOR ACQUISITION WITH DIRECT SEQUENCE TRANSMISSION

TAD's and be detected) for the code of length to be acquired. By using several SDCs in parallel, each designed to detect a shorter unique sequence, it will be shown that the mean acquisition time can be significantly reduced.

A flow graph representation of the signal acquisition technique is illustrated with consideration given to detection probability, false alarm probability and probability of alignment. The generating function of the acquisition time is determined from the flow graph of the new technique, and the mean and standard deviation of the acquisition time for various detection, false alarm, and alignment probabilities are calculated.

2.2 The Acquisition Algorithm

The proposed acquisition system is composed of a bank of matched filters and SDC systems. Each filter is assumed to be matched to a unique segment of a direct sequence (DS) code, further, it is assumed that M-chip tapped analog delays (TAD), are used to make up each matched filter. Hence, the overall DS code with codelength ℓ, is contained entirely in n TADs, and each TAD can be matched to M chips, such that

$$M = \ell/n \text{ (in chips)} \tag{2.1}$$

Clearly, the number of TADs will increase substantially as the codelength ℓ increases, (for a fixed value of M). For analysis purposes, it will be assumed that M is an unconstrained variable; this is justified in practice, since it is possible to cascade several TADs together to form a larger codelength M.

By using TADs, each matched to M chips, it is possible to match to the entire codelength ℓ, with n TADs, provided each TAD is weighted to a unique segment of the codelength. By adjusting the weights of the TAD taps, such that TAD_1 is matched to the first M chip segment, TAD_2 is matched to the second M chip segment, etc. It becomes evident that the TADs will peak in sequence if the proper segments of codelength are received in the correct order. Thus, sequence detection circuitry is required, Fig. 2.2. Once it has been established that a TAD is matched with the received signal, the sequence detection circuits (SDCs) must determine which TAD has peaked, and whether the peaking TAD is correct in sequence, Fig. 2.3. Every SDC is designed such that it is in a waiting state, and begins operating only once the correct "start-up" state occurs, (for SDC_1 this means that TAD_1 matches). Once the SDC begins operation, it will wait for the next M chips to be sampled and shifted in, then a decision must be made to determine if the next peaking TAD is in correct sequence. The SDC will continue to operate as long as the peaking TAD sequence is correct, however, if the peaking TAD is out of sequence, the SDC will return back to the waiting state. If the entire sequence is correctly received, acquisition will be declared, and the receiver will move to the tracking phase of synchronization.

By using several SDCs, Fig. 2.4, all working simultaneously and each designed to begin operating for a unique TAD peak, the acquisition time can be reduced, (i.e. by keeping ℓ constant and increasing n, the acquisition time will decrease). The effect of increasing SDCs will be examined later. It is important to note that n refers to the number of

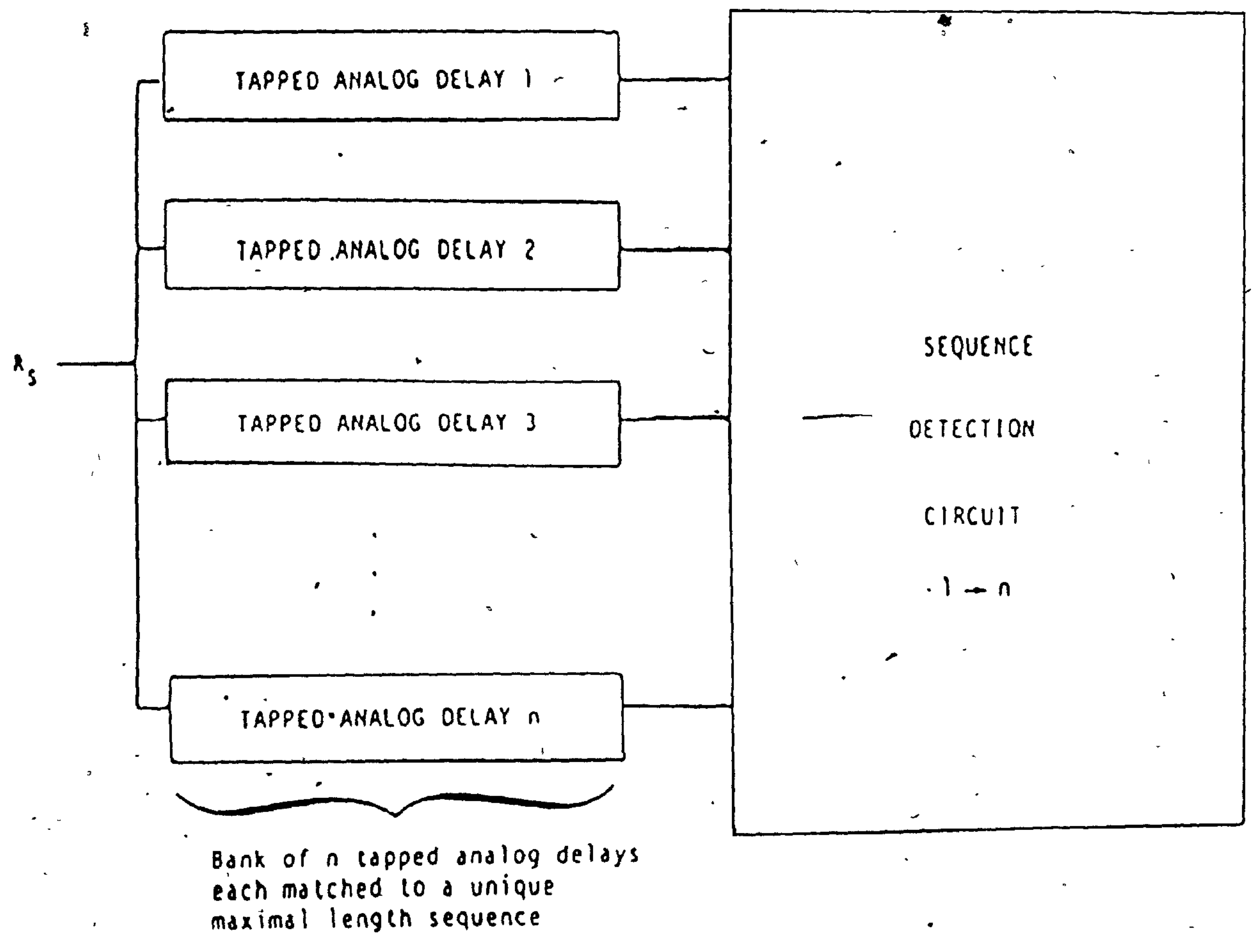


FIG. 2.2: CONFIGURATION OF n-TAPPED ANALOG DELAYS AND SEQUENCE DETECTION CIRCUIT

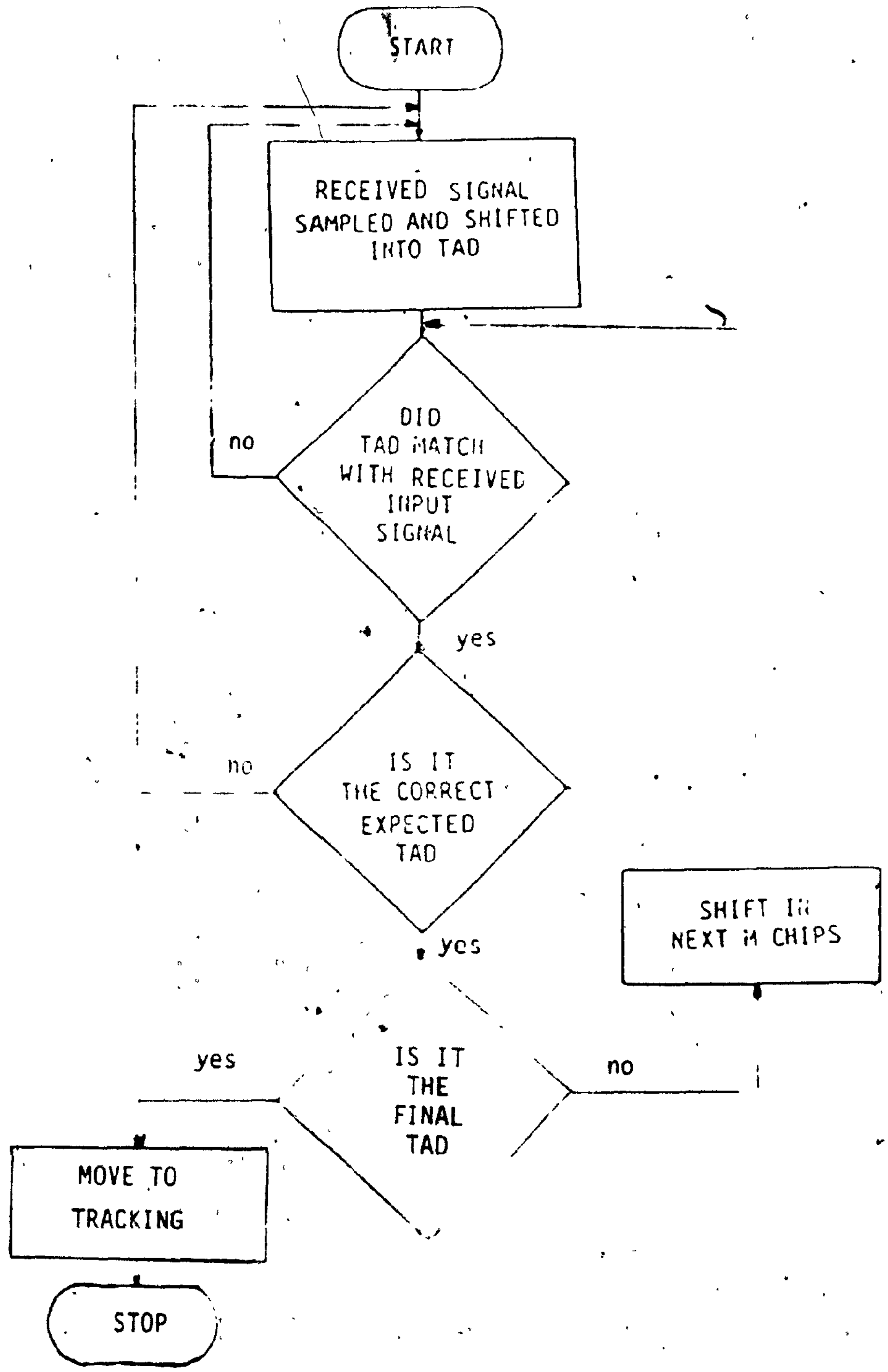


FIG. 2.3: FLOWCHART OF THE ACQUISITION TECHNIQUE
(Action taken by one of the sequence detection circuits)

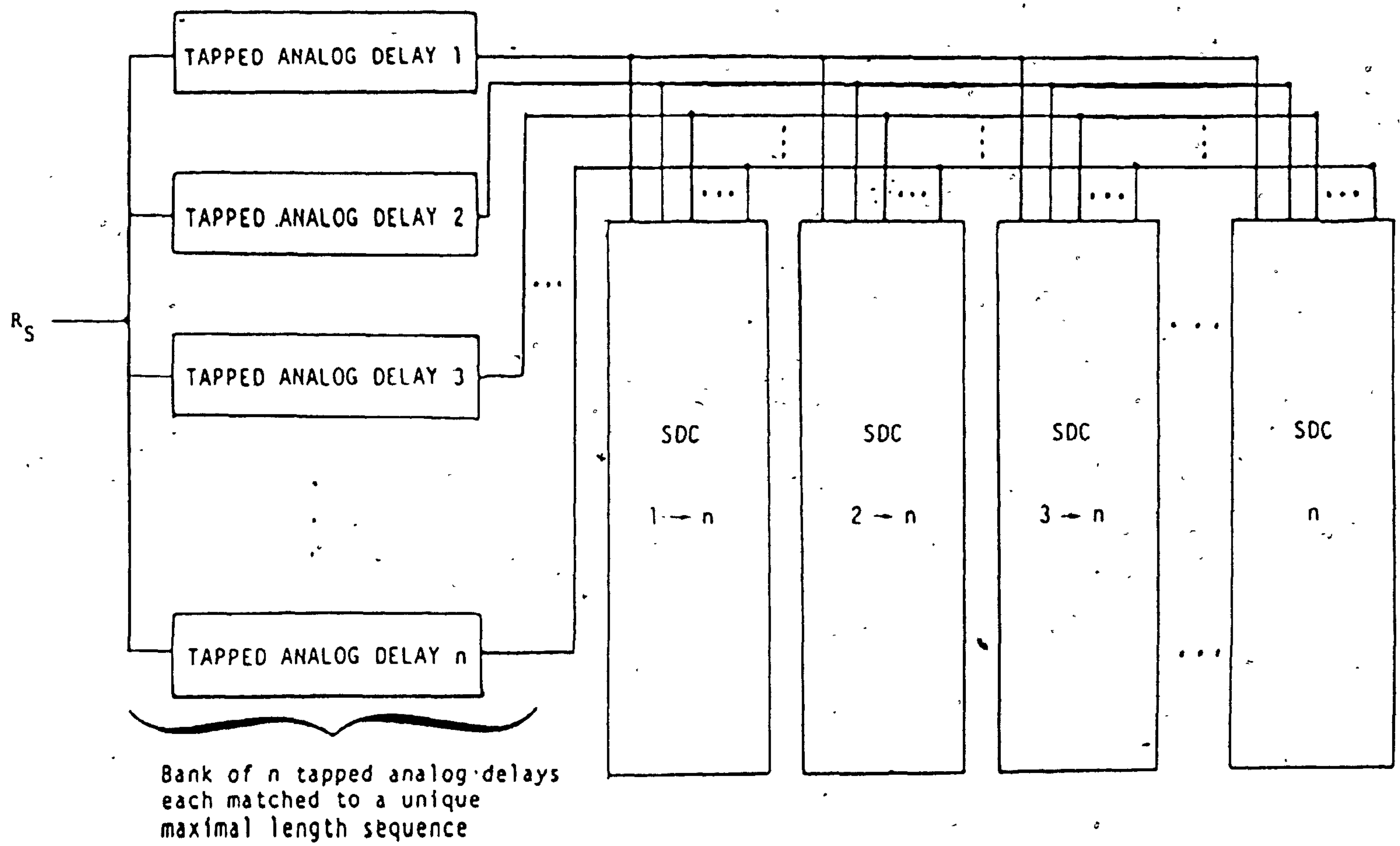


FIG. 2.4 : CONFIGURATION OF n-TAPPED ANALOG DELAYS AND n-SEQUENCE DETECTION CIRCUITS

SDCs and to the number of divisions of the codelength ℓ. For n equal to unity the codelength is not divided, and hence there is no need to use a SDC.

The flow diagram, to be presented, is based on the assumption that a multiple of SDCs, operating in parallel are used to acquire the input, where each SDC is designed to detect one of the following unique sequences:

$$
\begin{array}{r}
1 \rightarrow 2 \rightarrow 3 \rightarrow \dots \rightarrow n \\
2 \rightarrow 3 \rightarrow \dots \rightarrow n \\
\vdots \quad \\
n-1 \rightarrow n \\
n
\end{array}
$$

1, 2, 3,...,n-1, n refers to TAD_1, TAD_2, TAD_3,...,TAD_{n-1}, TAD_n correctly matching to the received input signal.

If one or more of these sequences is correctly detected, acquisition will be declared and the system will move to tracking.

2.3 Flow Diagram Representation of the n-SDC Acquisition Technique

To aid in analyzing this multi-SDC acquisition scheme, a flow diagram representation of the acquisition technique is presented in Fig. 2.5. By using this flow diagram approach, it is possible to obtain the statistical behaviour (mean and standard deviation), of the acquisition time, as in [8] and [10]. Since there are n SDCs, working in parallel to acquire a code of length ℓ, the probabilty of alignment can be expressed, using (2.1), as

$$P_{align} = n/\ell = \frac{1}{m} \tag{2.2}$$

Thus, the probability of misalignment is,

$$(1-P_{align}) = \frac{m-1}{m} \tag{2.3}$$

We also define probability of detection or no detection, and the probability of false alarm or no false alarm as μ_d, λ_m, μ_f and λ_r respectively.

If the input signal is aligned with one of the weighted TADs, with probability P_{align}, then there is a probability that the aligned signal may not be detected (missed) due to interference, noise, jamming, etc. This is denoted as

λ_m = probability that the aligned signal is not detected

However, if the signal is aligned and is detected acquisition will be declared and the system will move to tracking. This probability, of detection is denoted as

μ_d = probability the aligned signal is detected

If the received signal is misaligned, which occurs with probability $(1-P_{align})$, there is a probability that the SDCs will correctly indicate the received signal is mialigned with the TADs, denoted as

λ_r = probability that misalignment is recognized

However, also associated with misalignment is a probability that the misaligned received signal is falsely declared to be aligned, resulting in false acquisition, denoted as

μ_f = probability of false declaration of acquisition

The probabiltiies λ_m, μ_d, λ_r, μ_f will be examined in detail later.

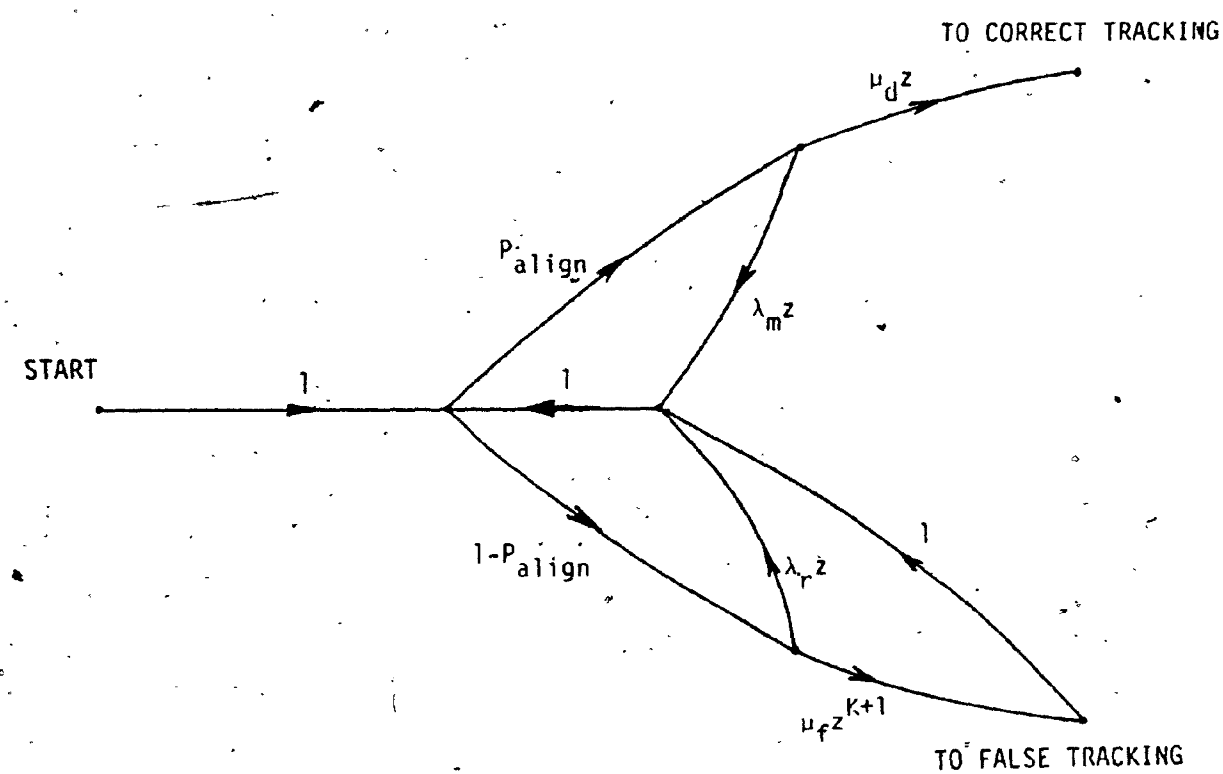


FIG. 2.5: FLOW DIAGRAM REPRESENTATION OF THE ACQUISITION SCHEME

In Fig. 2.5, z is used to represent a delay of one codelength. The additional delay, z^K, associated with false acquisition probability, μ_f, is a delay penalty resulting from the time spent trying to track a misaligned input signal, and K is an integer multiple of the codelength. For example K=2, indicates a penalty of two codelengths.

Consider a probability tree, Fig. 2.6, for a properly aligned input signal. It is possible to determine an expression for the probability of miss λ_m and the probability of detection μ_d. Clearly, there is a probability at each node of detection, μ_{d_j} or no detection $(1-\mu_{d_j})$, for the j^{th} SDC, (both μ_{d_j} and $(1-\mu_{d_j})$ will be discussed in detail later). Since each SDC is working seperately, we assume independence. Thus a miss will result with the following probability,

$$P_{miss}=\lambda_m=(1-\mu_{d_1})(1-\mu_{d_2})(1-\mu_{d_3})\dots(1-\mu_{d_n}) \qquad (2.4)$$

While the detection probability is given by

$$\begin{aligned} P_{detection} &= \mu_d = 1-\lambda_m \\ &= \mu_{d_1}+(1-\mu_{d_1})\mu_{d_2}+(1-\mu_{d_1})(1-\mu_{d_2})\mu_{d_3}+\dots+ \\ &\quad (1-\mu_{d_1})(1-\mu_{d_2})(1-\mu_{d_3})\dots(1-\mu_{d_{n-1}})\mu_{d_n} \end{aligned} \qquad (2.5)$$

If the received signal is misaligned then from Fig. 2.7, the expressions for the probabiltiy of recognition λ_r and the probability of false acquisition λ_f can be obtained. At each node there is a probability of false acquisition μ_{f_j} or no false acquisition $(1-\mu_{f_j})$ for the j^{th} SDC. (μ_{f_j} and $(1-\mu_{f_j})$ to be discussed in detail later).

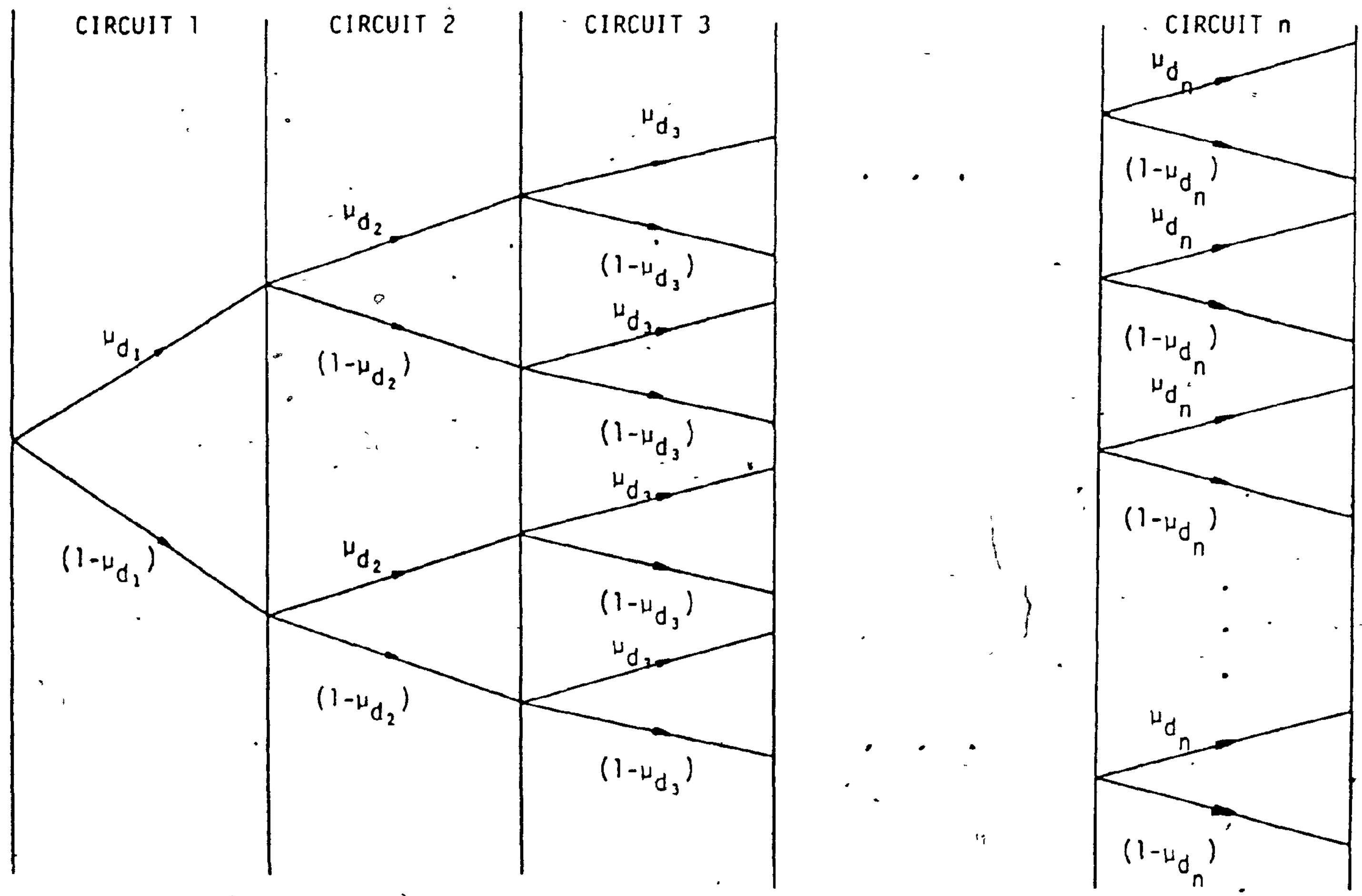


FIG 2.6: PROBABILITY TREE FOR ALIGNMENT

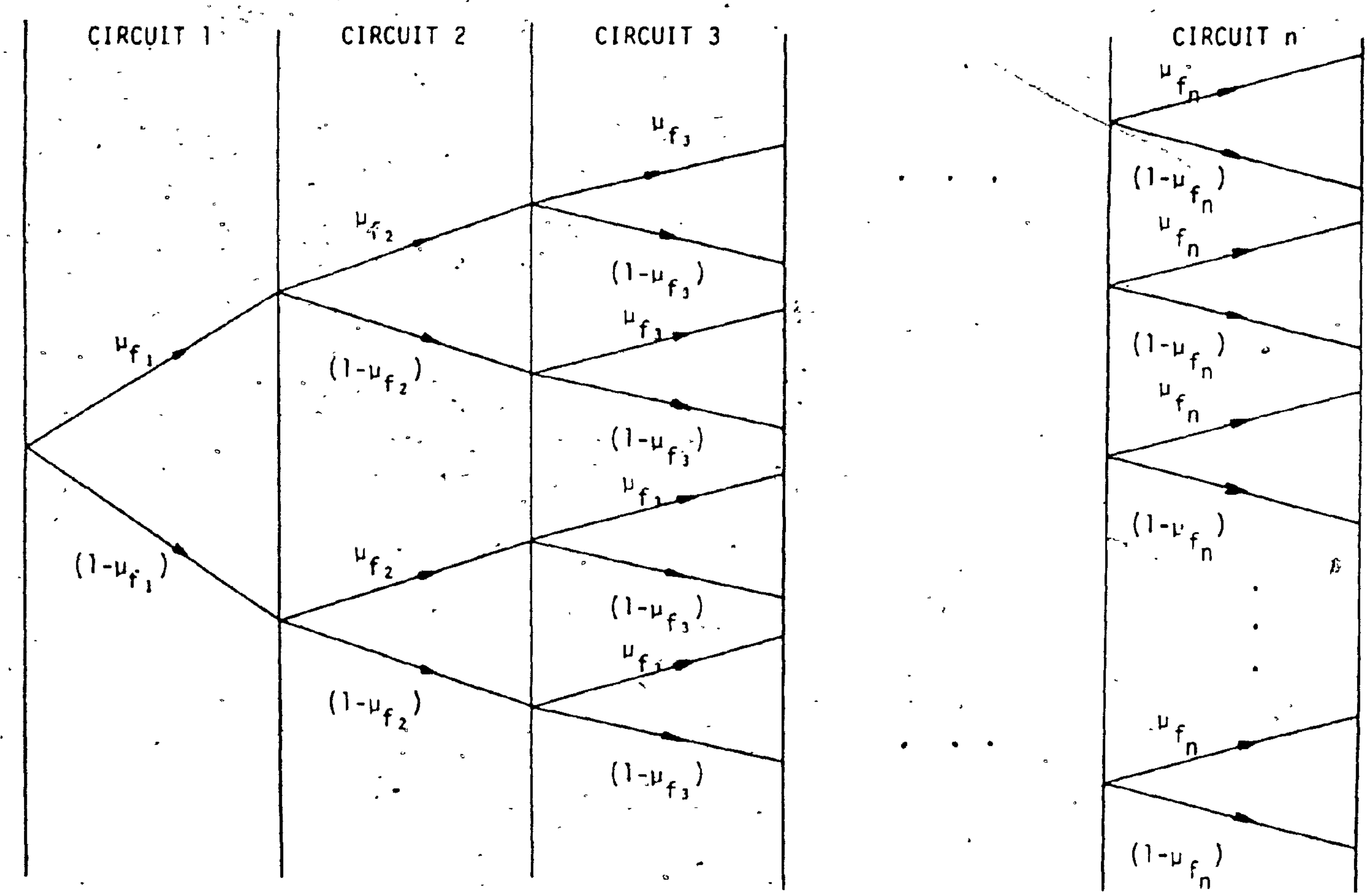


FIG. 2.7: PROBABILITY TREE FOR MISALIGNMENT

Assuming independence, the probability of recognition (no false alarm) is

$$\lambda_r=(1-\mu_{f_1})(1-\mu_{f_2})(1-\mu_{f_3})\ldots(1-\mu_{f_n}) \tag{2.6}$$

and probability of false acquisition is given by

$$\begin{aligned}\mu_f &= 1-\lambda_r \\ &= \mu_{f_1}+(1-\mu_{f_1})\mu_{f_2}+(1-\mu_{f_1})(1-\mu_{f_2})\mu_{f_3}+\ldots+ \\ &\qquad (1-\mu_{f_1})(1-\mu_{f_2})(1-\mu_{f_3})\ldots(1-\mu_{f_{n-1}})\mu_{f_n}\end{aligned} \tag{2.7}$$

The branch μ_d, in Fig. 2.5, indicates the probability of correct detection, and is composed of several subsets μ_{d_j}, j=1,2,3,...,n, where each μ_{d_j} represents the probability of detection for a specific SDC. The branch μ_d is shown in Fig. 2.8. Associated with each subset μ_{d_j}, in Fig. 2.8, is a transition probability denoted as α_i, where

$$\alpha_i = P_{d_i} \text{(the probability that the correctly peaked TAD}_i\text{ will be detected)} \tag{2.8}$$

$$(i=1,2,3,\ldots,n)$$

For a movement to occur along any subset branch μ_{d_j}, the TAD corresponding to that node must peak. As an example, consider μ_{d_1} (corresponding to the SDC for detecting sequence $1\to2\to3\to\ldots\to n$), in Fig. 2.8. If TAD_1 peaks and is detected, SDC moves to State 2, this happens

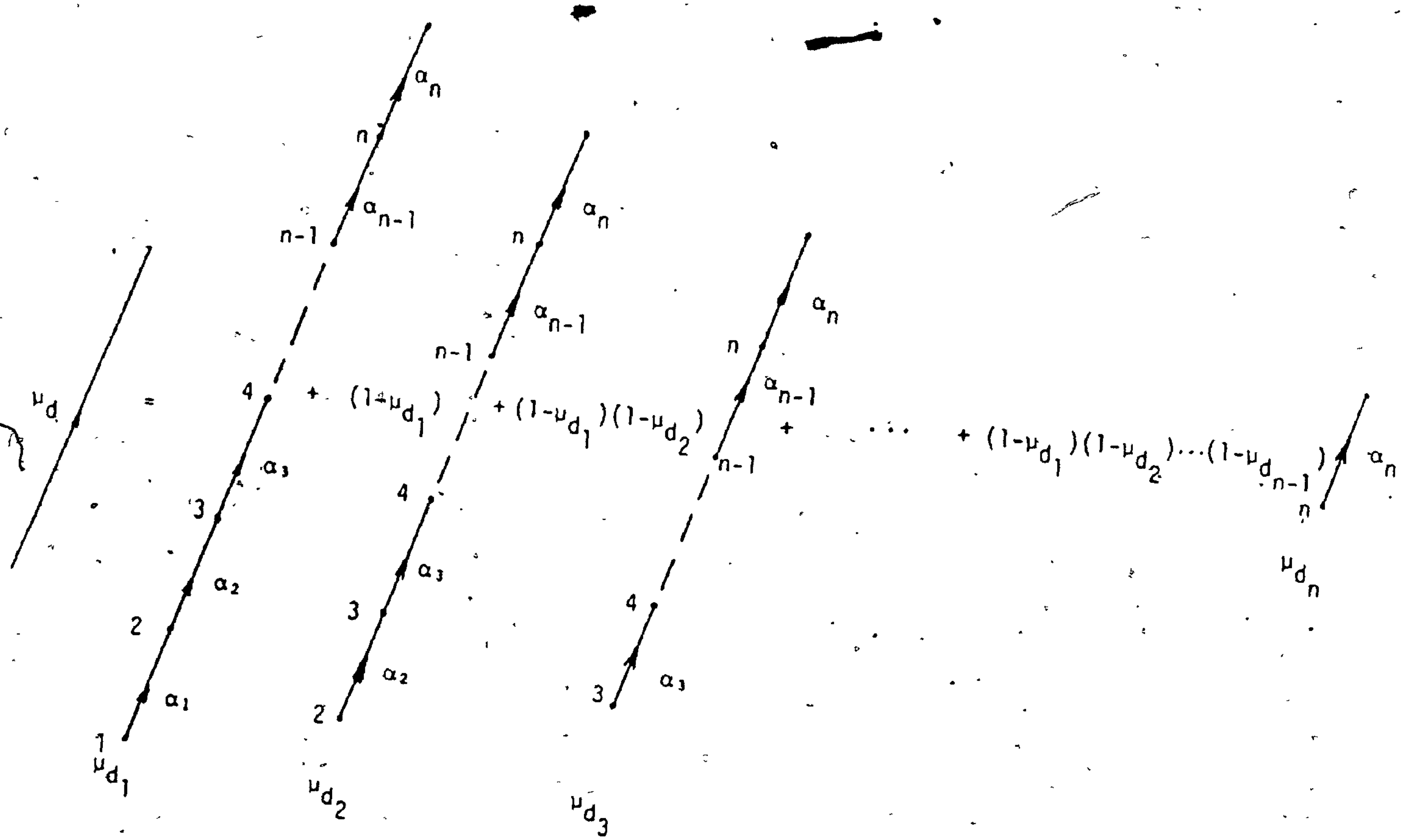


FIG. 2.8: REPRESENTATION OF THE μ_d BRANCH

with probability α_1, where

$\alpha_1 = P_{d_1}$ (probability of correct peaking TAD_1 being detected)

Once the next M chips are received, a decision is made to determine which, if any, TAD has peaked, if TAD_2 peaks a transition to state 3 results with probability α_2. This process proceeds in a similar manner for subsequent transitions. When the correct sequence is detected, (i.e. all transistions have occured) for any of the subsets μ_{d_j}, j=1,2,3,...,n, then acquisition is declared and tracking begins.

If for some reason (noise, jamming, etc.) a TAD does not peak, even though the code is propertly aligned, the SDCs will declare misalignment (miss) of the correct sequence, this occurs with the probability λ_m. The branch λ_m is a probability composed of several subset probabilities $\lambda_{m_j}=(1-\mu_{d_j})$ j=1,2,3,...,n, as shown in Fig. 2.9. Clearly, each subset will eventually lead to a decision that the threshold was not exceeded by one of the TADs, denoted by a branch ε_i, where ε_i is defined by,

$$\varepsilon_i=(1-\alpha_i) = \text{(the probability that the } i^{th} \text{ TAD will be aligned but will not peak)} \quad (i=1,2,3,\ldots,n) \tag{2.9}$$

The value of α_i remains the same as in (2.8). As an example, consider the subset $\lambda_{m_1}=(1-\mu_{d_1})$ in Fig. 2.9. At node 1, SDC is waiting for TAD_1 to peak, however if TAD_1 doesn't peak a transition along ε_i will result, this happens with probability $(1-P_{d_1})$. If TAD_1 does peak, a movement along branch α_1 will result, with probability P_{d_1}. Eventually, there will be a transition along the ε_i (i=1,2,3,...,n) branch, indicating the

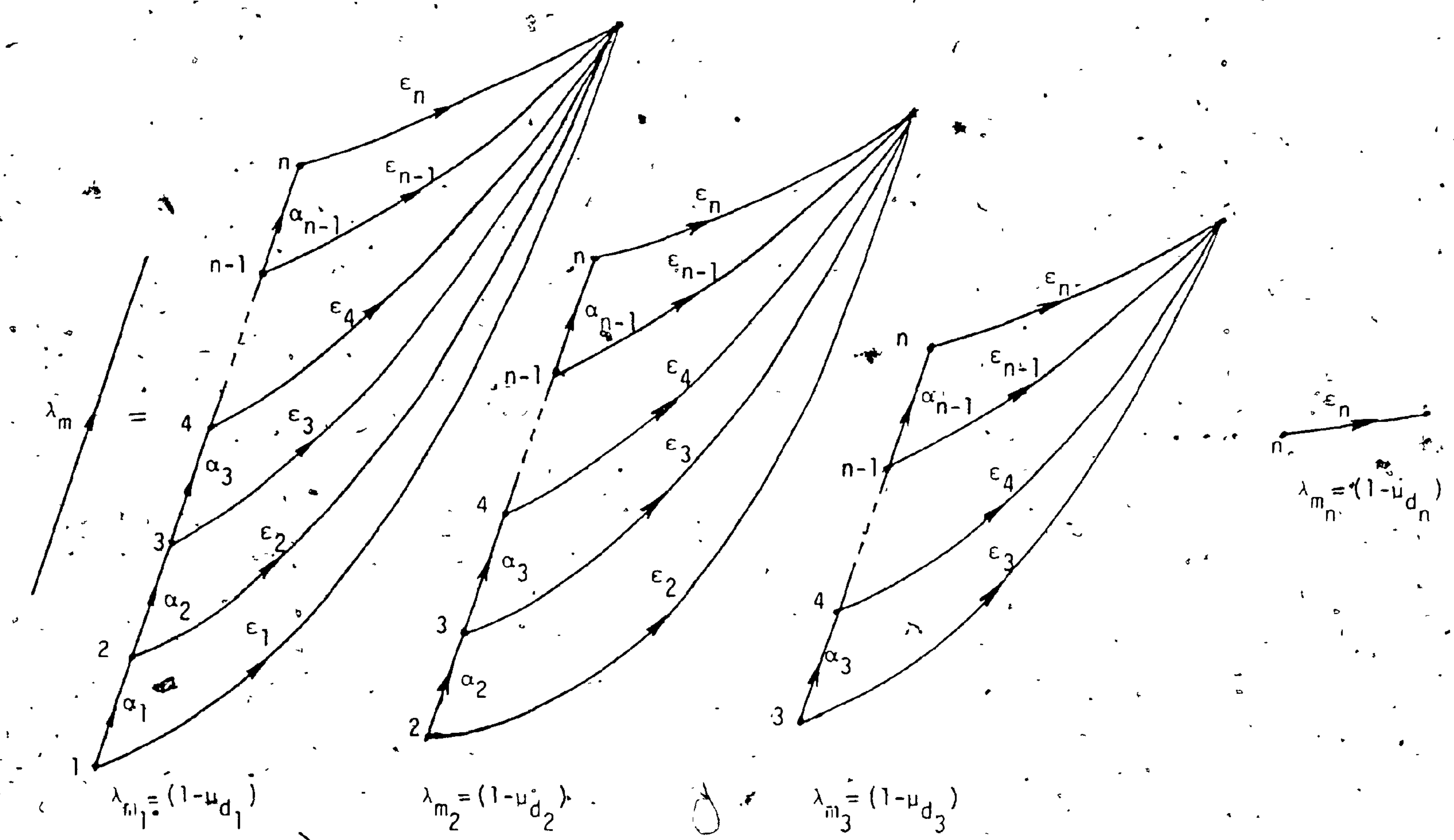


FIG. 2.9: REPRESENTATION OF THE λ_m BRANCH

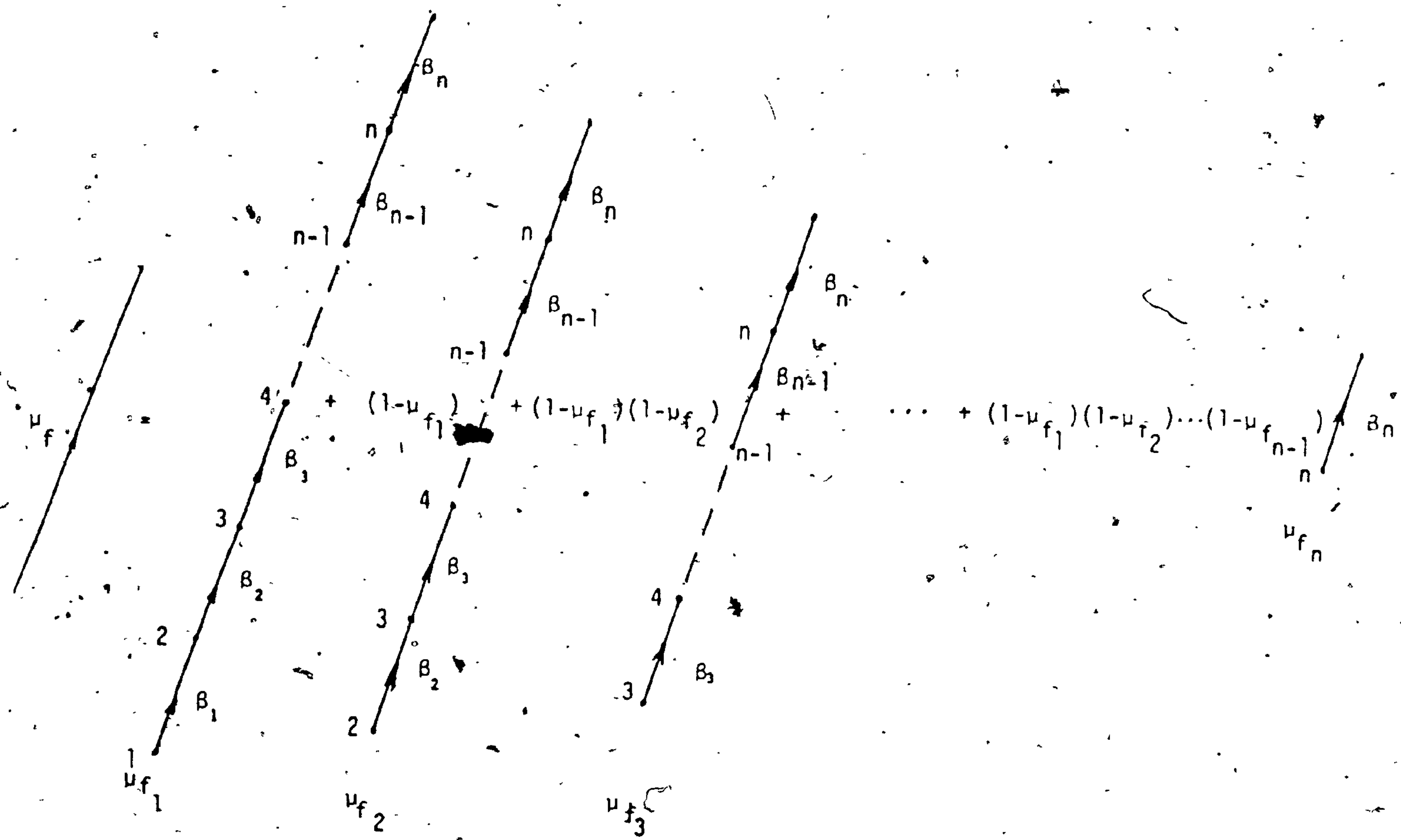


FIG. 2.10: REPRESENTATION OF THE μ_f BRANCH

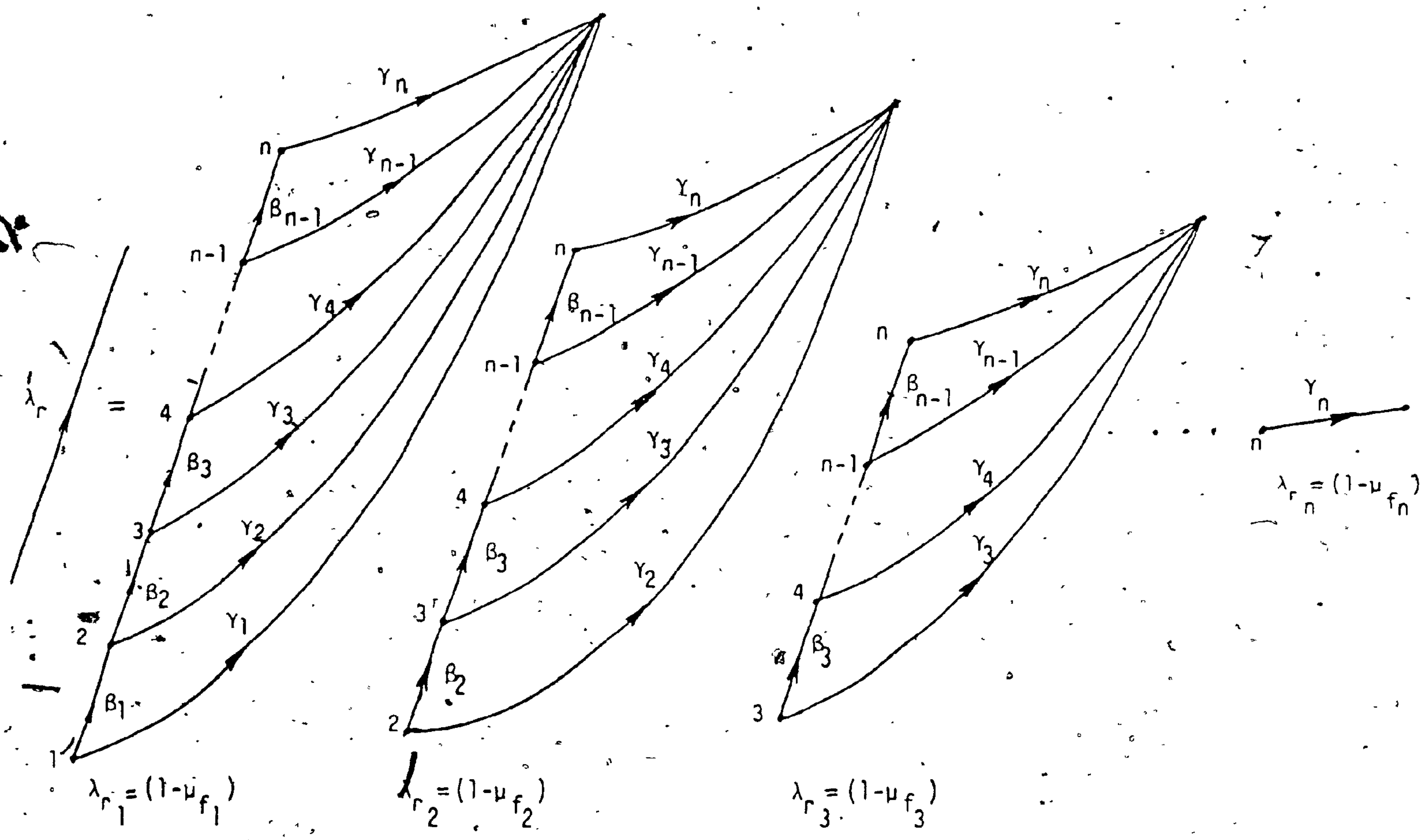


FIG. 2.11: REPRESENTATION OF THE λ_r BRANCH

i^{th} TAD did not peak, (i.e. detection missed).

The branch marked μ_f in Fig. 2.5, is the probability of declaring false acquisition, it is composed of several subsets μ_{f_j}, j=1,2,3,...,n, where each μ_{f_j} represents the probability that the j^{th} SDC will falsely declare acquisition, shown in Fig. 2.10. The probability associated with a transition in any subset is denoted by β_i, where,

$$\beta_i = P_{f_i} \quad (1=1,2,3,\ldots,n) \text{ (the probability that the } i^{th} \text{ TAD will falsely peak)} \tag{2.10}$$

The branch λ_r is the probability that the receiver will recognize that the input signal is misaligned. λ_r is composed of several subsets $\lambda_{r_j}=(1-\mu_{f_j})$, j=1,2,3,...,n, shown in Fig. 2.11, where λ_{r_j} is the probability the j^{th} SDC will recognize misalignment. For recognition of misalignment to occur, all subsets of λ_r must indicate misalignment.

This corresponds to a transition along a branch marked γ_i where

$$\gamma_i = (1-\beta_i) \text{ (the probability the } i^{th} \text{ TAD will not falsely peak)} \tag{2.11}$$

$$(i=1,2,3,\ldots,n)$$

The parameter β_i remains the same as in (2.10).

2.4 Probability Expressions of the Flow Diagram Representation

The parameters λ_m, μ_d, λ_r and μ_f are various probabilities associated with miss, detection, recognition and false acquisition respectively, of the systems. In this section, expressions for these four probabilities will be determined. The branch μ_d, represents the probability that an SDC will correctly detect alignment, and move to tracking. From Fig. 2.8, the probability μ_{d_1} is given by,

$$\begin{aligned}\mu_{d_1} &= \alpha_1 \cdot \alpha_2 \cdot \alpha_3 \cdot \dots \cdot \alpha_n \\ &= \prod_{i=1}^{n} \alpha_i\end{aligned} \tag{2.12}$$

The probability μ_{d_2} is given by

$$\mu_{d_2} = \prod_{i=2}^{n} \alpha_i \tag{2.13}$$

and in general

$$\mu_{d_j} = \prod_{i=j}^{n} \alpha_j \tag{2.14}$$

Hence the probability μ_d from (2.5) is

$$\begin{aligned}\mu_d &= \sum_{j=1}^{n} \mu_{d_j} \prod_{\substack{A=1 \\ j\neq 1}}^{j-1} (1-\mu_{d_A}) \\ &= \sum_{j=1}^{n} \prod_{i=j}^{n} \alpha_i \prod_{\substack{A=1 \\ j\neq 1}}^{j-1} (1- \prod_{I=A}^{n} \alpha_I)\end{aligned} \tag{2.15}$$

The probability of miss λ_m, is composed of several subsets $\lambda_{m_j} = (1-\mu_{d_j})$, $j=1,2,3,\dots,n$. Each subset represents the probability that the j^{th} SDC will miss detection. We note that for a miss to occur, in the whole system, all SDCs should miss. The probability λ_{m_1} is given by

$$\lambda_{m_1} = (1-\mu_{d_1}) = \varepsilon_1 + \alpha_1\varepsilon_2 + \alpha_1\alpha_2\varepsilon_3 + \ldots + \varepsilon_n \prod_{i=1}^{n-1} \alpha_i$$
$$= \sum_{g=1}^{n} \varepsilon_g \prod_{i=1}^{g-1} \alpha_i \tag{2.16}$$

where the expression $\prod_{i=1}^{g-1} \alpha_i$, in (2.16) equals unity for g=1.

The probability λ_2 is given by,

$$\lambda_{m_2} = (1-\mu_{d_2}) = \varepsilon_2 + \alpha_2\varepsilon_3 + \alpha_3\alpha_3\varepsilon_4 + \ldots + \varepsilon_n \prod_{i=2}^{n-1} \alpha_i$$
$$= \sum_{g=2}^{n} \varepsilon_g \prod_{i=2}^{g-1} \alpha_i \tag{2.17}$$

where the expression $\prod_{i=2}^{g-1} \alpha_i$, in (2.17), is unity.

In general,

$$\lambda_{m_h} = (1-\mu_{d_h}) = \sum_{g=h}^{n} \varepsilon_g \prod_{i=h}^{g-1} \alpha_i \tag{2.18}$$

where the expression $\prod_{i=h}^{g-1} \alpha_i$, in (2.18) is unity for g=h.

Now λ_m is given by (2.4),

$$\lambda_m = \sum_{g=1}^{n} \varepsilon_g \prod_{i=1}^{g-1} \alpha_i \cdot \sum_{g=2}^{n} \varepsilon_g \prod_{i=2}^{g-1} \alpha_i \cdot \sum_{g=3}^{n} \varepsilon_g \prod_{i=3}^{g-1} \alpha_i \cdot$$
$$\ldots \cdot \sum_{g=n}^{n} \varepsilon_g \prod_{i=n}^{g-1} \alpha_i \tag{2.19}$$
$$= \prod_{h=1}^{n} \sum_{g=h}^{n} \varepsilon_g \prod_{i=h}^{g-1} \alpha_i$$

where the expression $\prod_{i=h}^{g-1}$, in (2.19), is unity for g=h. The probability

μ_f is the probability of falsely declaring acquisition, as in Fig. 2.10, where

$$\mu_{f_1} = \beta_1 \cdot \beta_2 \cdot \beta_3 \cdot \dots \cdot \beta_n = \prod_{i=1}^{n} \beta_i \tag{2.20}$$

The probability of μ_{f_2}, is

$$\mu_{f_2} = \beta_2 \cdot \beta_3 \cdot \beta_4 \cdot \dots \cdot \beta_n = \prod_{i=2}^{n} \beta_i \tag{2.21}$$

and in general

$$\mu_{f_c} = \prod_{i=c}^{n} \beta_i \tag{2.22}$$

Thus, the probability of μ_f, given by (2.7) is

$$\mu_f = \sum_{c=1}^{n} \mu_{f_c} \prod_{\substack{B=1 \\ C\neq 1}}^{k-1} (1-\mu_{f_B})$$

$$= \sum_{c=1}^{n} \prod_{i=c}^{n} \beta_i \prod_{\substack{B=1 \\ C\neq 1}}^{c-1} (1- \prod_{I=B}^{n} \beta_I) \tag{2.23}$$

Finally, the branch λ_r, the probability of recognizing misalignment, is composed of the product of subsets $\lambda_{r_j} = (1-\mu_{f_j})$, j=1,2,3,...,n. The probability represented by the first subset $\lambda_{r_1}=(1-\mu_{f_1})$ is given by

$$\lambda_{r_1} = (1-\mu_{f_1}) = \gamma_1+\beta_1\gamma_2+\beta_1\beta_2\gamma_3+\dots+\gamma_n \prod_{i=1}^{n-1} \beta_i = \sum_{q=1}^{n} \gamma_q \prod_{i=1}^{q-1} \beta_i \tag{2.24}$$

where the expression $\prod_{i=1}^{q-1} \beta_i$, in (2.24), is unity for q=1.

For the probability λ_{r_2},

$$\lambda_{r_2} = (1-\mu_{f_2}) = \gamma_2+\beta_2\gamma_3+\beta_2\beta_3\gamma_4+\dots+\gamma_n \prod_{i=2}^{n-1} \beta_i$$
$$= \sum_{q=2}^{n} \gamma_q \prod_{i=2}^{q-1} \beta_i \tag{2.25}$$

where the expression $\prod_{i=2}^{q-1} \beta_i$, in (2.25), is unity for q=2.

In general,

$$\lambda_r = \sum_{q=s}^{n} \gamma_q \prod_{i=s}^{q-1} \beta_i \tag{2.26}$$

where the expression $\prod_{i=s}^{q-1} \beta_i$, in (2.26), is unity for q=s.

Thus the probability λ_r, given by (2.26) is

$$\lambda_r = \sum_{q=1}^{n} \gamma_q \prod_{i=1}^{q-1} \beta_i \cdot \sum_{q=2}^{n} \gamma_q \prod_{i=2}^{q-1} \beta_i \cdot \sum_{q=3}^{n} \gamma_q \prod_{i=3}^{q-1} \beta_i \cdot$$
$$\dots \cdot \sum_{q=n}^{n} \gamma_q \prod_{i=n}^{q-1} \beta_i \tag{2.27}$$
$$= \prod_{s=1}^{n} \sum_{q=s}^{n} \gamma_q \prod_{i=s}^{q-1} \beta_i$$

where the expression $\prod_{i=s}^{q-1} \beta_i$, in (2.27), is unity for q=s.

2.5 Statistical Behaviour of the Acquisition Technique

The flow diagram, Fig. 2.5, can be simplified to form a generating function, from which the mean and standard deviation of the acquisition time (in codelengths) can be obtained, similar to the technique in [8].

Examining Fig. 2.5, we see that the feedforward paths of the flow diagram are given by:

$$M_1 = P_{align}\,\mu_d z$$

and the single feedback paths

$$P_{11} = P_{align}\,\lambda_m z$$
$$P_{21} = (1-P_{align})(\mu_f z^{K+1}+\lambda_r z)$$

where z refers to a delay of one codelength.

Hence the generating function of the acquisition time is

$$GF(z) = \frac{P_{align}\,\mu_d z}{1-[(P_{align}\,\lambda_m z)+(1-P_{align})(\mu_f z^{K+1}+\lambda_r z)]} \tag{2.28}$$

It is now possible to determine the mean and standard deviation of the acquisition time, by evaluating the proper derivatives of (2.28).

The generating function can be defined as,

$$F_T(z) = \sum_{n=0}^{\infty} f(n)\, z^n \tag{2.29}$$

and the mean as

$$\begin{aligned}\bar{T} &= \sum_{n=0}^{\infty} n\, f(n) \\ &= \left.\frac{\partial F_T(z)}{\partial z}\right|_{z=1}\end{aligned} \tag{2.30}$$

Thus, the mean of (2.28) is,

$$\bar{T} = \left.\frac{\partial \; GF(z)}{\partial z}\right|_{z=1}$$

$$= \frac{\partial}{\partial z}\left[\frac{P_{align}\,\mu_d z}{1-[(P_{align}\,\lambda_m z)+(1-P_{align})(\mu_f z^{K+1}+\lambda_r z)]}\right]\Bigg|_{z=1} \tag{2.31}$$

Evaluating (2.31), with $(\lambda_m = 1-\mu_d)$ and $(\lambda_r = 1-\mu_f)$ yields

$$\bar{T} = \frac{1+K\;\mu_f(1-P_{align})}{(P_{align}\;\mu_d)} \tag{2.32}$$

Substituting (2.23) and (2.15) into (2.32),

$$\bar{T} = \frac{1 + K \cdot \sum\limits_{c=1}^{n} \prod\limits_{W=c}^{n} \beta_W \prod\limits_{\substack{B=1 \\ c\neq 1}}^{c-1} \left(1- \prod\limits_{X=B}^{n} \beta_X\right)\cdot(1-P_{align})}{P_{align} \cdot \sum\limits_{j=1}^{n} \prod\limits_{Y=j}^{n} \alpha_Y \prod\limits_{\substack{A=1 \\ j\neq 1}}^{j-1} \left(1- \prod\limits_{Z=A}^{n} \alpha_Z\right)} \tag{2.33}$$

where P_{align} is given by (2.2) and K is an integer value associated with the the penalty time.

The variance of the generating function, defined by (2.29) is given by,

$$\sigma^2_T = \sum_{n=0}^{\infty} n^2 f(n) - \sum_{n=0}^{\infty} n^2 f^2(n)$$

$$= \left[\frac{\partial^2 F_T(z)}{\partial z^2} + \frac{\partial F_T(z)}{\partial z} - \left(\frac{\partial F_T(z)}{\partial z}\right)^2\right]\Bigg|_{z=1} \tag{2.34}$$

Thus the variance of the acquisition time is given by

$$\sigma^2_T = \left[\frac{\partial^2 GF}{\partial z^2} + \frac{\partial GF}{\partial z} - \left(\frac{\partial GF}{\partial z}\right)^2\right]\Bigg|_{z=1} \tag{2.35}$$

After some calculations, Appendix A, the variance of (2.28) is found to be

$$\sigma^2_T = \frac{(1-P_{align}\,\mu_d)+[K\,\mu_f(1-P_{align})]^2+K\,\mu_f(1-P_{align})(2+K\,\mu_d\,P_{align})}{(P_{align}\,\mu_d)^2} \qquad (2.36)$$

where μ_d and μ_f are given by (2.15) and (2.23) respectively. P_{align} is given in (2.2) and K is an integer value associated with penalty time.

With expressions obtained for the mean (2.32) and variance (2.35) of the acquisition time, it is possible to determine the effect of varying detection probability, false alarm probability, n (number of SDCs), and alignment probability, on the mean, and standard deviation, where

$$\sigma_T = (\sigma^2_T)^{1/2}$$

$$= \frac{[(1-P_{align}\,\mu_d)+[K\,\mu_f\cdot(1-P_{align})]^2+K\,\mu_f(1-P_{align})(2+K\,\mu_d\,P_{align})]^{1/2}}{P_{align}\,\mu_d} \qquad (2.37)$$

represents the standard deviation of the acquisition time.

2.6 Results

The mean and standard deviation values of the acquisition time are evaluated, using equations (2.33) and (2.37) respectively, for various alignment probabilities, detection and false alarm probabilities, and sequence detection circuit configurations.

The probability of detection and the probability of false alarm are varied for a fixed codelength ℓ=1000 chips and a fixed number of tapped analog delays (or equivalently a fixed number of sequence detection circuits). The results presented are in units of codelength (a mean acquisition time of 7.32 would represent 7.32 codelengths). The probability of false alarm and probability of detection are the probabilities associated with each tapped analog delay, (β_i and α_i respectively). As in [8], it is assumed that $\beta_i = \beta$ (for all i), and $\alpha_i = \alpha$ (for all i).

Tables 1-5 show the effect of varying the probability of detection, α, and the probability of false alarm, β, on the mean and standard deviation of the acquisition time. From Table 1 we see some very interesting results. First, we note that increasing the number of sequence detection circuits, n, (or the number of tapped analog delays), increases the probability of alignment, P_{align}. Thus, one would expect, the mean acquisition time to decrease for an increasing n, however, this is only true for some of the results. For a low probability of detection α, (α = 0.5), the mean acquisition time tends to increase as n, the number of sequence detection circuits increases. As the detection probability, α, increases, (α=0.9), the mean acquisition time decreases, until n reaches a set limit. By comparing the results for n=1 and n=10 where α=0.9 and $\beta=0.5\times10^{-4}$, (it is known that the SNR during acquisition is very low, thus justifying the pessimistic values selected for β and α), we observe that the mean acquisition time for n=10 (318.68 codelengths) is significantly less than the mean time for n=1 (1234.63 codelengths). However, if we compare the results for n=10 and n=20 (α=0.9, $\beta=0.5\times10^{-4}$) we notice that the mean acquisition time for n=20 is 456.98 codelengths, while the mean acquisition time for n=10 is 318.68 codelengths. Thus, it is evident that increasing the number of sequence detection circuits n, is beneficial only up to a certain limit. Fig 2.12, shows the effect of increasing the sequence detection circuits, n, on the mean acquisition time, for various detection and false alarm probabilities, (β=0.5x10^{-4}).

As expected, increasing the probability of false alarm, β, results in an increased mean acquisition time, (all other parameters constant), although the increase is insignificant, Fig. 2.13. The effect of other paramters dominates the effect of the false alarm probability β, for this case. However, there is a major reduction in the mean acquisition time for an increasing detection probability, α, (all other parameters constant), Fig. 2.14. It is important to note, that the standard deviation of the acquisition time is approximately equal to the mean acquisition time. Thus, at worst, the signal will be acquired in a period equal to twice the mean, or in some instances the signal can be obtained within a few code length periods. As is typical with other acquisition techniques, the highest efficiency (i.e. minimum acquisition time), occurs when the detection probability, α, is maximized.

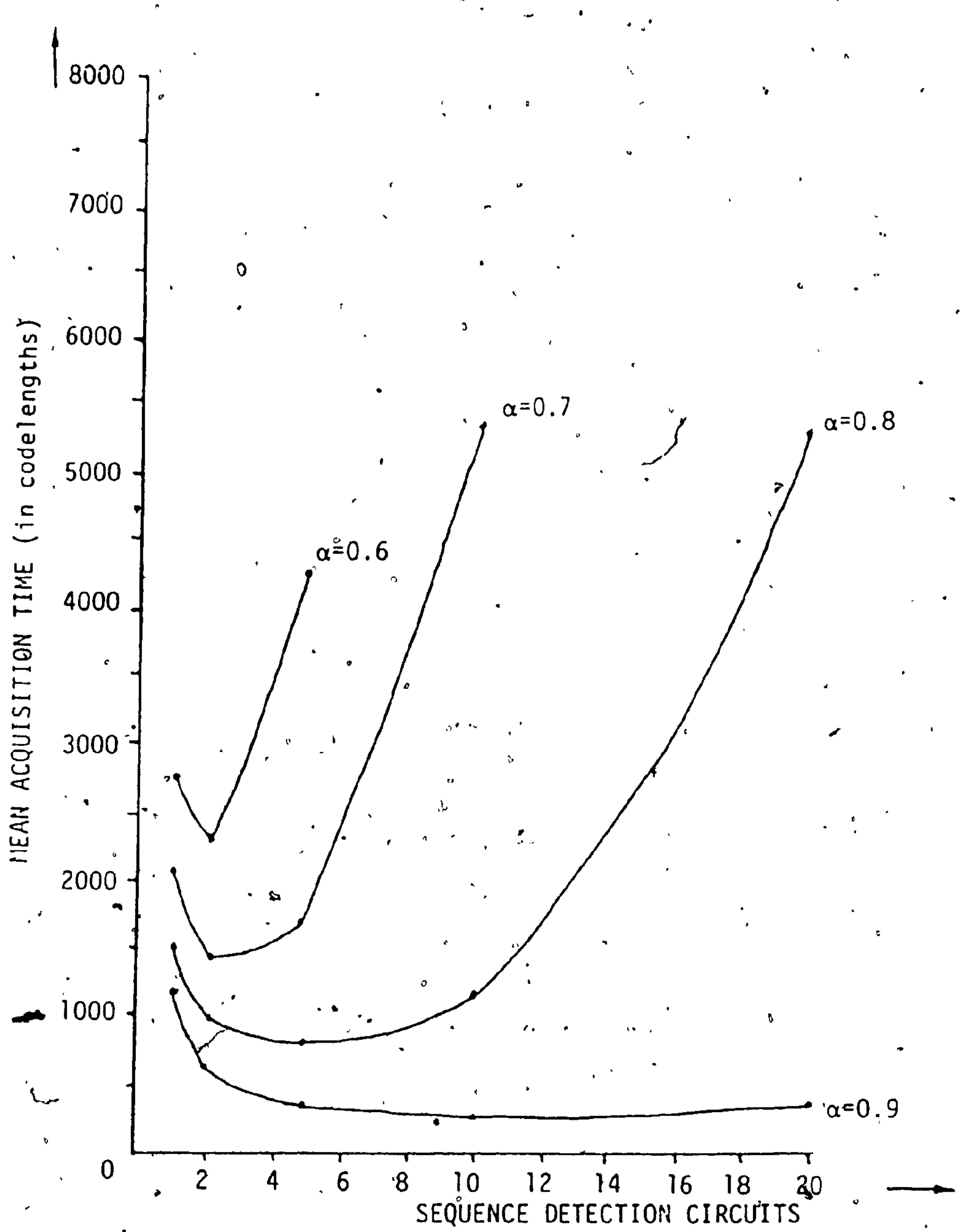


FIG 2.12 : MEAN ACQUISITION TIME VERSUS SEQUENCE DETECTION CIRCUITS FOR VARYING PROBABILITY OF DETECTION (α)
(PROBABILITY OF FALSE ALARM $\beta = 0.5 \times 10^{-4}$)
(CODELENGTH = 1000 chips)

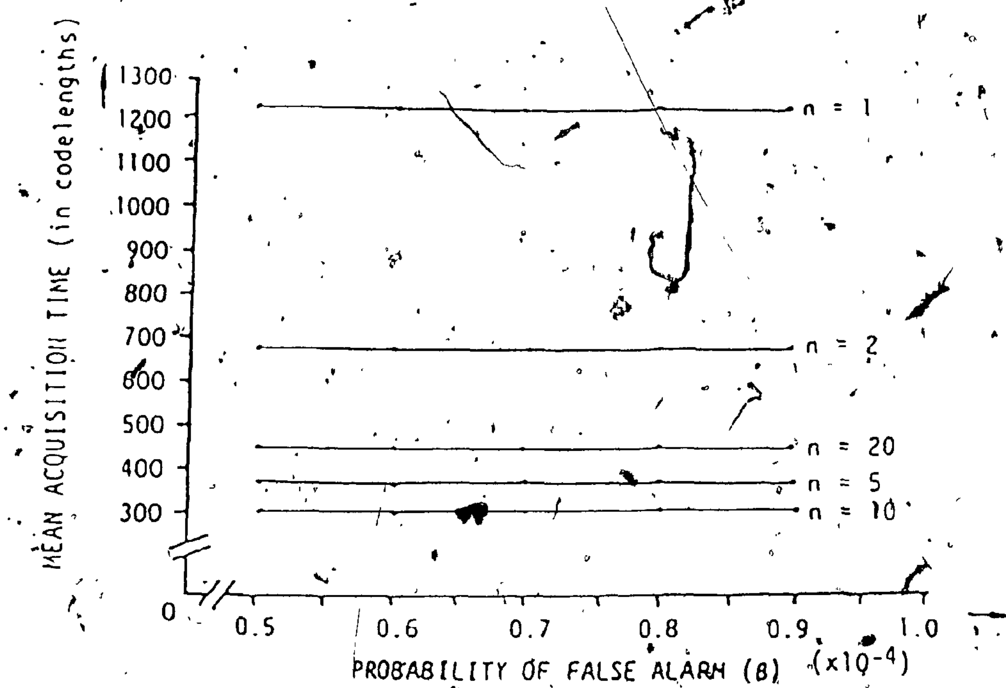


FIG. 2.13: MEAN ACQUISITION TIME VERSUS PROBABILITY OF FALSE ALARM (β) FOR A VARYING NUMBER OF SEQUENCE DETECTION CIRCUITS (n)
(PROBABILITY OF DETECTION α = 0.9)
(CODELENGTH = 1000 chips)

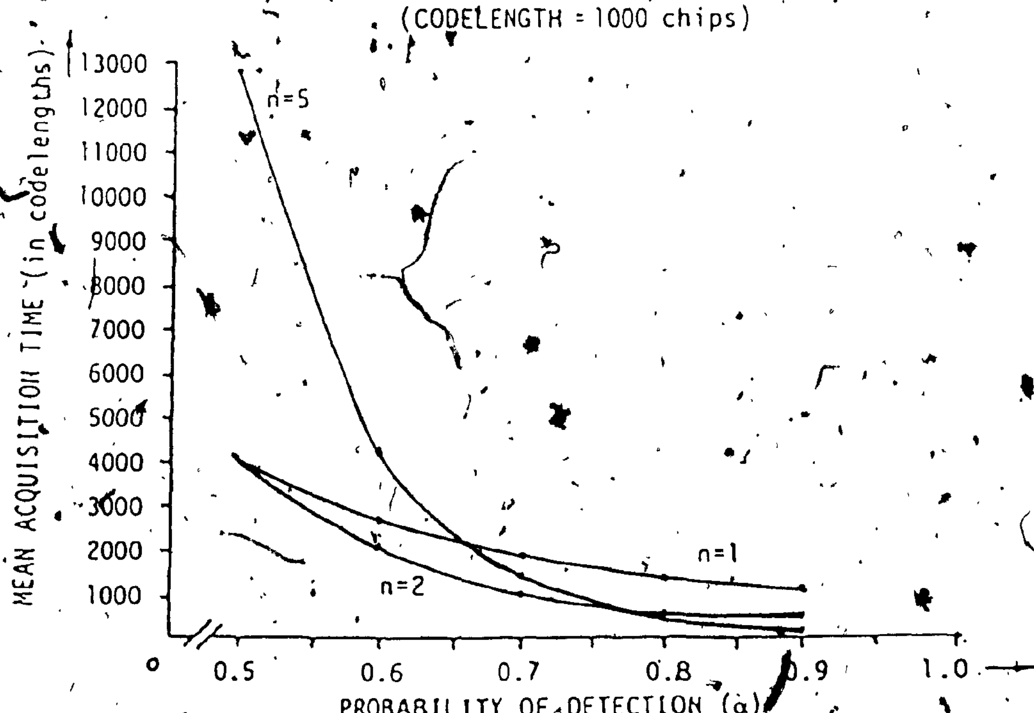


FIG 2.14: MEAN ACQUISITION TIME VERSUS PROBABILITY OF DETECTION (α) FOR A VARYING NUMBER OF SEQUENCE DETECTION CIRCUITS (n)
(PROBABILITY OF FALSE ALARM β = 0.5 x 10^{-4})
(CODELENGTH = 1000 chips)

TABLE 1: MEAN AND STANDARD DEVIATION UNDER VARIOUS DETECTION PROBABILITIES USING 1 SDC

α \ $\beta(\times 10^{-4})$	0.5	0.6	0.7	0.8	0.9
0.5	4000.20 3999.70	4000.24 3999.74	4000.28 3999.78	4000.32 3999.82	4000.36 3999.86
0.6	2777.92 2777.42	2777.94 2777.44	2777.97 2777.47	2778.00 2777.50	2778.03 2777.53
0.7	2040.92 2040.42	2040.94 2040.43	2040.96 2040.46	2040.98 2040.48	2041.00 2040.50
0.8	1562.58 1562.08	1562.59 1562.09	1562.61 1562.11	1562.62 1562.12	1562.64 1562.14
0.9	1234.63 1234.13	1234.64 1234.14	1234.65 1234.15	1234.67 1234.17	1234.68 1234.18

$n = 1$ $\ell = 1000$ chips

$P_{align} = n/\ell = 0.001$

MEAN (units of codelengths)
STD DEVIATION (units of codelengths)

α: probability of detection (for each TAD)
β: probability of false alarm (for each TAD)
n: number of sequence detection circuits
λ: codelength

TABLE 2: MEAN AND STANDARD DEVIATION UNDER VARIOUS DETECTION PROBABILITIES USING 2 SDC

α \ $\beta(\times 10^{-4})$	0.5	0.6	0.7	0.8	0.9
0.5	4000.20 3999.70	4000.24 3999.74	4000.28 3999.78	4000.32 3999.82	4000.36 3999.86
0.6	2314.93 2314.43	2314.95 2314.45	2314.98 2314.48	2315.00 2314.50	2315.02 2314.50
0.7	1457.80 1457.30	1457.81 1457.31	1457.83 1457.33	1457.84 1457.34	1457.86 1457.36
0.8	976.61 976.11	976.62 976.12	976.63 976.13	976.64 976.14	976.65 976.15
0.9	685.91 685.41	685.91 685.41	685.92 685.42	685.93 685.43	685.93 685.43

$n = 2$ $\ell = 1000$ chips.

$P_{align} = n/\ell = 0.002$

MEAN (units of codelengths)
STD DEVIATION (units of codelengths)

α: probability of detection (for each TAD)
β: probability of false alarm (for each TAD)
n: number of sequence detection circuits
λ: codelength

TABLE 3: MEAN AND STANDARD DEVIATION UNDER VARIOUS DETECTION PROBABILITIES USING 5 SDC

α \ β(x10^{-4})	0.5	0.6	0.7	0.8	0.9
0.5	12800.64	12800.76	12800.89	12801.02	12801.15
	12800.14	12800.26	12800.39	12801.52	12801.65
0.6	4286.91	4286.95	4286.99	4287.04	4287.08
	4286.41	4286.45	4286.49	4286.54	4286.58
0.7	1700.06	1700.07	1700.09	1700.11	1700.12
	1699.56	1699.57	1699.59	1699.61	1699.62
0.8	762.98	762.99	762.99	763.00	763.01
	762.48	762.48	762.49	762.50	762.51
0.9	376.35	376.36	376.36	376.37	376.37
	375.85	375.86	375.86	375.86	375.87

$n = 5$ $\ell = 1000$ chips.

$P_{align} = n/\ell = 0.005$

MEAN (units of codelengths)
STD DEVIATION (units of codelengths)

α: probability of detection (for each TAD)
β: probability of false alarm (for each TAD)
n: number of sequence detection circuits
ℓ: codelength

TABLE 4: MEAN AND STANDARD DEVIATION UNDER VARIOUS DETECTION PROBABILITIES USING 10 SDC

α \ β(x10^{-4})	0.5	0.6	0.7	0.8	0.9
0.5	204810.14	204812.17	204814.19	204816.22	204818.25
	204809.64	204811.67	204813.69	204815.72	204817.75
0.6	27564.98	27565.26	27565.53	27565.[illegible]0	27566.08
	27564.48	27564.76	27565.03	27565.30	27565.58
0.7	5057.58	5057.63	5057.68	5057.73	5057.78
	5057.08	5057.13	5057.18	5057.23	5057.28
0.8	1164.21	1164.22	1164.23	1164.25	1164.26
	1163.71	1163.72	1163.73	1163.75	1163.76
0.9	318.68	318.68	318.69	318.69	318.69
	318.18	318.18	318.19	318.19	318.19

$n = 10$ $\ell = 1000$ chips.

$P_{align} = n/\ell = 0.010$

MEAN (units of codelengths)
STD DEVIATION (units of codelengths)

α: probability of detection (for each TAD)
β: probability of false alarm (for each TAD)
n: number of sequence detection circuits
ℓ: codelength

TABLE 5: MEAN AND STANDARD DEVIATION UNDER VARIOUS DETECTION PROBABILITIES USING 20 SDC

α \ $\beta(\times 10^{-4})$	0.5	0.6	0.7	0.8	0.9
0.5	104862738.29 104862[illegible]7.78	104863765.99 104863765.50	104864793.73 104864793.23	104865821.49 104865820.79	104866849.27 104866848.77
0.6	2279371.05 2279370.55	2279393.38 2279392.86	2279415.72 2279415.22	2279438.06 2279437.56	2279460.41 2279459.91
0.7	89522.55 89522.05	89523.43 89522.93	89524.30 89523.80	89525.18 89524.68	89526.06 89525.56
0.8	5421.28 5420.78	5421.33 5420.83	5421.38 5420.88	5421.44 5420.94	5421.49 5420.99
0.9	456.98 456.48	456.99 456.49	456.99 456.49	456.99 456.49	457.00 456.50

$n = 20 \quad \ell = 1000$ chips.

$P_{align} = n/\ell = 0.020$

MEAN (units of codelengths)
STD DEVIATION (units of codelengths)

α: probability of detection (for each TAD)
β: probability of false alarm (for each TAD)
n: number of sequence detection circuits
ℓ: codelength

CHAPTER 3

A NEW SIGNAL ACQUISITION TECHNIQUE FOR A FREQUENCY HOPPING (FH) SPREAD SPECTRUM SYSTEM

3.1 Introduction

In Chapter 2, it was assumed that the n-SDC acquisition technique was used to acquire a pseudonoise (PN), direct sequence (DS) spread spectrum (SS) signal. However, with slight modifications the n-SDC acquisition scheme can be used to acquire received signals that are transmitted by frequency hopping.

In this chapter, an n-SDC acquisition scheme is presented for frequency hopped spread spectrum signal transmissions. The values of detection probability, α_i, from (2.3) and false alarm probability, β_i, from (2.10), are determined for the system under various operating environments, and the corresponding mean acquisition time is evaluated using (2.33). Since communication in a frequency hopping network is not always isolated (i.e. a single transmitter-reciever network), the effect of like user interference and intentional jamming will also be considered.

The results of [2], for spectral splatter effects will be presented and used to determine the performance of the new acquisition scheme in a more realistic operating environment.

In [7], performance characteristics are given for various schemes of coarse synchronization of frequency-hopped spread spectrum signals. The analysis provided in this chapter will be based on a matched filter scheme similar to that presented by Putman et al. However, the acquisition scheme will have a degree of flexibility, in that the total number of available frequencies, ℓ', (equivalent to the codelength in Chapter 2) and the number of frequencies per matched filter bank M', (equivalent to the number of taps per TAD in Chapter 2) can be varied by the designer. Thus, tradeoffs will be discussed for choosing the optimum number of total channels and the number of frequencies per filter bank, under various operating environments.

3.2 The Frequency Hopping (FH) Acquisition Technique

It is assumed, for analysis, that the acquisition scheme for frequency hopped signals is composed of n' banks of matched filters. Figure 3.1, shows one bank of the matched filtering technique. The bank represents a standard matched filter scheme, where the filters are matched to a specific frequency hopping pattern consisting of M' frequencies. Further, it is assumed that the frequency hopping pattern is generated by a linear feedback shift register (LFSR) and a frequency synthesizer. Thus, the M' frequency pattern is determined by a unique feedback connection or the LFSR. If the total number of available frequencies ℓ', is small say $\ell' = M'$, then the user is susceptible to both like user interference (assuming a like users share the same ℓ' frequencies), and jamming interference. To minimize the susceptibility to external interference, the number of frequencies ℓ' is increased. This increase in frequency results in the need for additional banks of matched filtering. It is assumed, that each bank of the n'-bank system will be matched to a unique frequency hopping sequence composed of M'/n' frequencies, and none of the M' available frequencies will repeat, (i.e. each of the ℓ' frequencies is used only once).

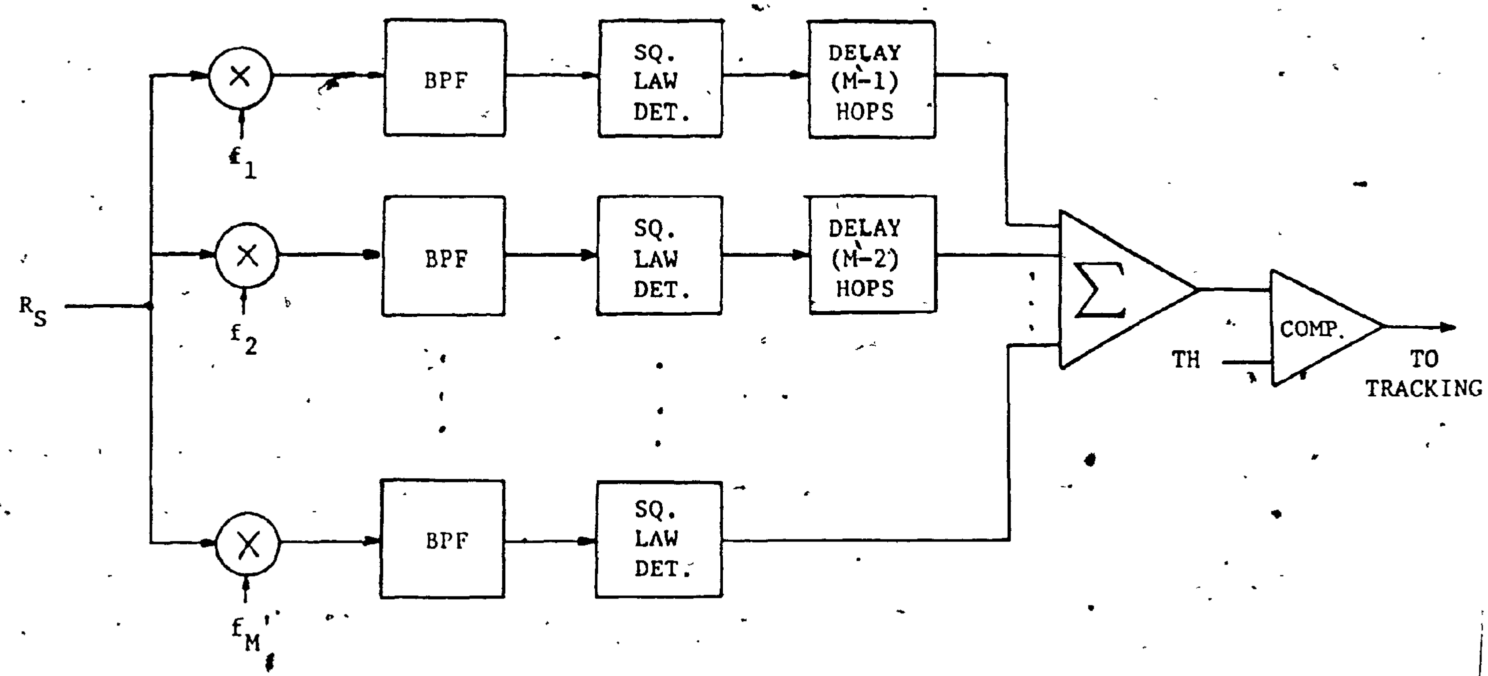


FIG. 3.1: SINGLE BANK OF A MATCHED FILTER ACQUISITION SCHEME
FOR FREQUENCY HOPPED SIGNALS

The analogy can now be drawn between the FH acquisition scheme and the direct sequence (DS) acquisition scheme presented in Chapter 2. Each matched filter bank in the FH system can be thought of as a tapped analog delay (TAD), where each tap is represented by the bandpass filter, square law detector, and delay, Fig. 3.1. When the proper frequency sequence is received, and the threshold exceeded, the corresponding bank will peak. Since n' banks of matched filters are used to acquire the received signal, sequence detection circuits (SDC) are required. The SDCs are used to detect the correct sequence 1→2→3→...→n' (where 1→2→3→...→n' refers to matched filter bank 1→ matched filter bank 2→ matched filter bank 3→...→matched filter bank n' peaking in correct order), as well, smaller sequences can also be detected 2→n', 3→n', etc., Fig. 3.2. The analysis presented for DS acquisition can now be applied to a FH acquisition scheme, and the resulting detection and false alarm probabilities can be determined under various operating conditions, to yield the mean acquisition time.

3.3 Bit Error Probability for the Frequency Hopping Acquisition Scheme

In the analysis of a frequency hopped spread spectrum system, consideration is often given to the operating environment. The operating environment, consists of; the number of users, the presence of intentional jamming, the type of jamming used, the number of availalbe frequencies (channels), the transmission power, the jamming power, spectral splatter effects, etc. Analysis of the new frequency hopping spread spectrum acquisition technique, is based on the assumption that

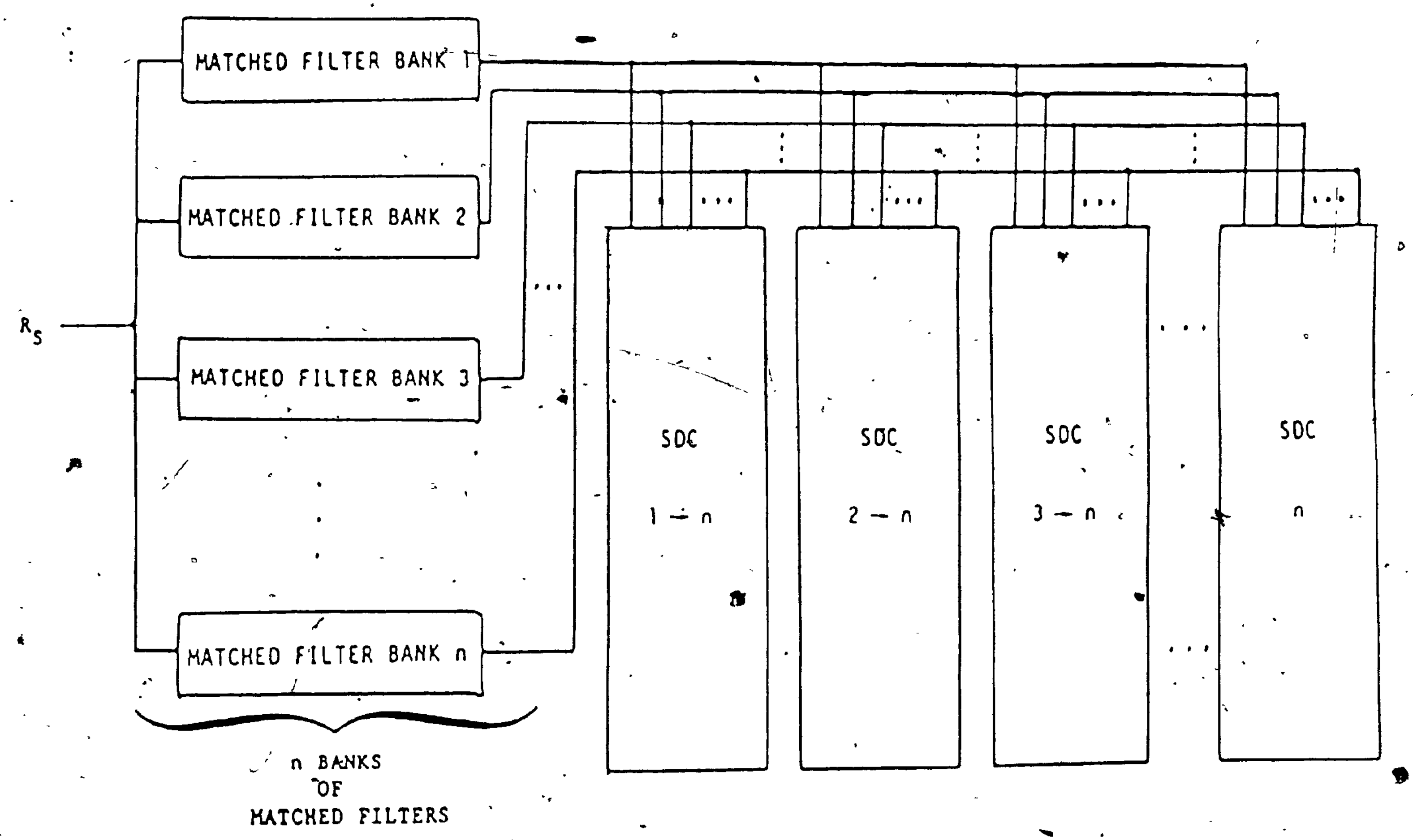


FIG. 3.2: CONFIGURATION OF n MATCHED FILTER BANKS AND n SEQUENCE DETECTION CIRCUITS

signal transmission occurs in the presence of partial band jamming. As well, the effect of spectral splatter is examined in the analysis under various signal-to-noise ratios, total jamming power and channel availability conditions.

To determine the mean acquisition time using (2.33) for a frequency hopped system, the detection and false alarm probabilities for the i^{th} filter bank, α_i and β_i, respectively, must be evaluated from the bit error value produced for a specified operating environment.

The equation for bit error probability, P_b, will be presented for the case where spectral splatter is significant only in the two channels adjacent to the transmission channel. The network is composed of independent frequency hopping systems each generating an equivalent amount output power, each user having the same total frequencies ℓ', each user is approximately stationary over one bit interval, and each user transmits similar waveforms.

To determine the bit error probability, P_b, several intermediate probabilities must first be defined. Initially, $P(E_{jk})$, the probability of the event E_{jk} occuring, must be determined. E_{jk} is defined as the event that j interferers use the transmission channel and k interferers use one of the two adjacent channels. The probability of bit error given E_{jk} is denoted as $P_{bE}(j,k)$ and thus the bit error probability is

$$P_b = \sum_{j=0}^{N} \sum_{k=0}^{N-j} P(E_{jk}) \, P_{bE}(j,k) \qquad (3.1)$$

where N is the total number of potentially interfering hoppers in the network.

Each interferer can transmit in any hopping channel with equal probability, and therefore the interferer can hit the transmission channel with probability d/ℓ'. The variable d is defined as the duty factor and represents the product of the probability that a significant portion of the interferer's waveform occurs during a bit interval and the probability that an interferer is transmitting. The probability that an interferer hits one of the two adjacent channels is $2d/\ell'$. It is assumed that the total number of available channels ℓ', is large enough to neglect the fact that the channels at the end of the bandwidth have only one adjacent channel instead of two.

The probability that the power does not hit the transmission channel or one of the adjacent channels is defined as $(1-3d/\ell')$. The number of ways in which the interferer can hit the transmission channel is $\binom{N}{j}$ where

$$\binom{N}{j} = \frac{N!}{j!\,(N-j)!} \tag{3.2}$$

If $j+k<N$ then the number of ways of hitting one of the adjacent channels is $\binom{N-j}{k}$. Thus, $P(E_{jk})$ is defined as

$$P(E_{jk}) = \binom{N}{j}\left(\frac{d}{\ell'}\right)^{j}\binom{N-j}{k}\left(\frac{2d}{\ell'}\right)^{k}\left(1-\frac{3d}{\ell'}\right)^{N-j-k} \tag{3.3}$$

substituting (3.3) into (3.1) yields

$$P_b = \sum_{j=0}^{N}\sum_{k=0}^{N-j} 2^k \binom{N}{j}\binom{N-j}{k}\left(\frac{d}{\ell'}\right)^{j+k}\left(1-\frac{3d}{\ell'}\right)^{N-j-k} P_{bE}(j,k) \tag{3.4}$$

To simplify the evaluation of (3.4), a limit L is chosen to reduce the number of integrations involved in calculating $P_{bE}(j,k)$.
If $N > L$, then the lower limit of bit error probability, P_L, is

$$P_L = \sum_{\substack{j=0 \\ j+k \leq L}}^{N} \sum_{k=0}^{N-j} 2^k \binom{N}{j}\binom{N-j}{k}\left(\frac{d}{\ell}\right)^{j+k}\left(1 - \frac{3d}{\ell}\right)^{N-j-k} P_{bE}(j,k) \quad (3.5)$$

and the upper limit of the bit error probability, for $j + k > L$ is

$$P_U = P_L + \frac{1}{2} \sum_{\substack{j=0 \\ j+k > L}}^{N} \sum_{k=0}^{N-j} 2^k \binom{N}{j}\binom{N-j}{k}\left(\frac{d}{\ell}\right)^{j+k}\left(1 - \frac{3d}{\ell}\right)^{N-j-k} \quad (3.6)$$

The conditional probability $P_{bE}(j,k)$, in (3.5) is defined by

$$P_{bE}(j,k) = \int_0^\infty \dots \int_0^\infty P\left(\sum_{i=1}^{j} x_i + K \sum_{i=1}^{k} y_i\right) \prod_{i=1}^{j} f(x_i) \prod_{i=1}^{k} f(y_i) dx_1 \dots dx_j \, dy_1 \dots dy_k \quad (3.7)$$

The expression x_i is the ratio of power from the i^{th} interferer using the transmission channel to the power of the desired signal at the demodulator. The expression y_i is used as a variable change, to simplify the expression (3.7), where $y_i = z_i/K$.

The variable z_i is the ratio of power from the i^{th} interferer using one of the adjacent channels to the power of the desired signal at the demodulator; K is the ratio at power from an adjacent channel interferer to the corresponding power that would be produced if the interferer were using the transmission channel. Basically, x_i and y_i are interference-to-signal ratios. The expression $P(\sum_{i=1}^{j} x_i + K \sum_{i=1}^{k} y_i)$ is the probability of bit error given j interferers in the transmission channel with interference-to-signal ratios x_1, x_2, x_3,....,x_j and k interferers in one of the two adjacent channels, with interference-to-signal ratios Ky_1, Ky_2, Ky_3,....,Ky_k, where the change of variable $y_i = z_i/K$ is substituted for simplification. By modelling each interferer as an independent zero-mean process, it is possible to treat the probability as a function of the sum of ratios x_i and y_i. For ease of analysis, the expression $P(\sum_{i=1}^{j} x_i + K \sum_{i=1}^{k} y_i)$ will be defined as follows

$$P(\sum_{i=1}^{j} x_i + K \sum_{i=1}^{k} y_i) = P(x') \qquad (3.8)$$

3.4 The Operating Environment: Jamming and Deployment

Thus far, it has been assumed that j interferers use the transmission channel and k interferers use one of the two adjacent channels. As well, each user shares the same ℓ' frequencies. We now consider the effect of jamming and the deployment of the transmitter and receiver.

If we assume the use of binary MSK, differential demodulation yields an expression for the bit error probability given in [18]

$$P(E) = \frac{1}{2} \exp\left(-\frac{\varepsilon R_s}{N_T}\right) \qquad (3.9)$$

where R_s is the desired signal power, N_T is the total white Gaussian noise power, and ε (approximately unity) is a variable depending on the demodulation. In the presence of other users and jamming, the noise N_T can be modelled as

$$N_T = N_{th} + N_j + R_s x' \qquad (3.10)$$

where N_{th} is the background and thermal noise, N_j is the power (noise) produced by the jammer, $N_j = 0$ when no jamming is present, and $R_s x'$ is the summation of the power resulting from the other users in the network.

If repeater jamming is used, and the repeater is sufficiently close, the probability of bit error (3.8), becomes

$$P(x') = \frac{1}{2} \exp\left(-\frac{\varepsilon R_s}{N_{th} + N_j + R_s x'}\right) \quad (3.11)$$

The results to be presented in a latter section, will be based on the assumption that the signal R_s is transmitted in the presence of partial-band jamming. The probability of bit error (3.8), for partial-band jamming is given by

$$P(x') = \left(1 - \frac{J}{\ell'}\right) S_0(x') + \frac{J}{\ell'} S_1(x') \quad (3.12)$$

J is defined as the number of channels the interferer hits out of the total ℓ' channels. $S_0(x')$ is the probability of bit error given x' and the event that the transmission channel is not jammed, defined as

$$S_0(x') = \frac{1}{2} \exp\left(-\frac{\varepsilon R_s}{N_{th} + R_s x'}\right) \quad (3.13)$$

We see that $S_0(x')$ is the probability of error caused only by background noise, thermal noise, or like user interference. The expression $S_1(x')$, is the probability of bit error given x' and the event that the transmission channel is jammed, defined as

$$S_1(x') = \frac{1}{2} \exp\left(-\frac{\varepsilon R_s}{N_{th} + N_j + R_s x'}\right) \quad (3.14)$$

The derivation of $S_0(x')$ and $S_1(x')$ is outlined in Appendix B.

In (3.7), $f(x_i)$ represents the probability density function for an

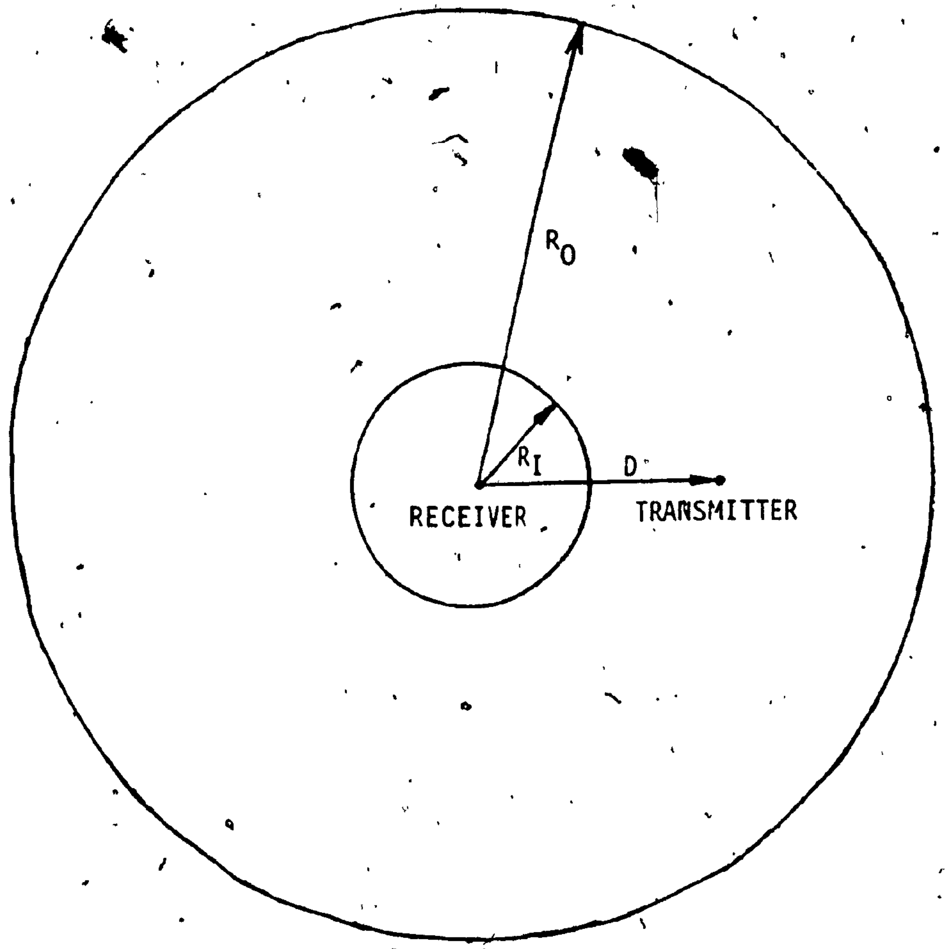


FIG. 3.3: BOUNDARY ALLOCATION FOR DEPLOYMENT OF INTERFERERS

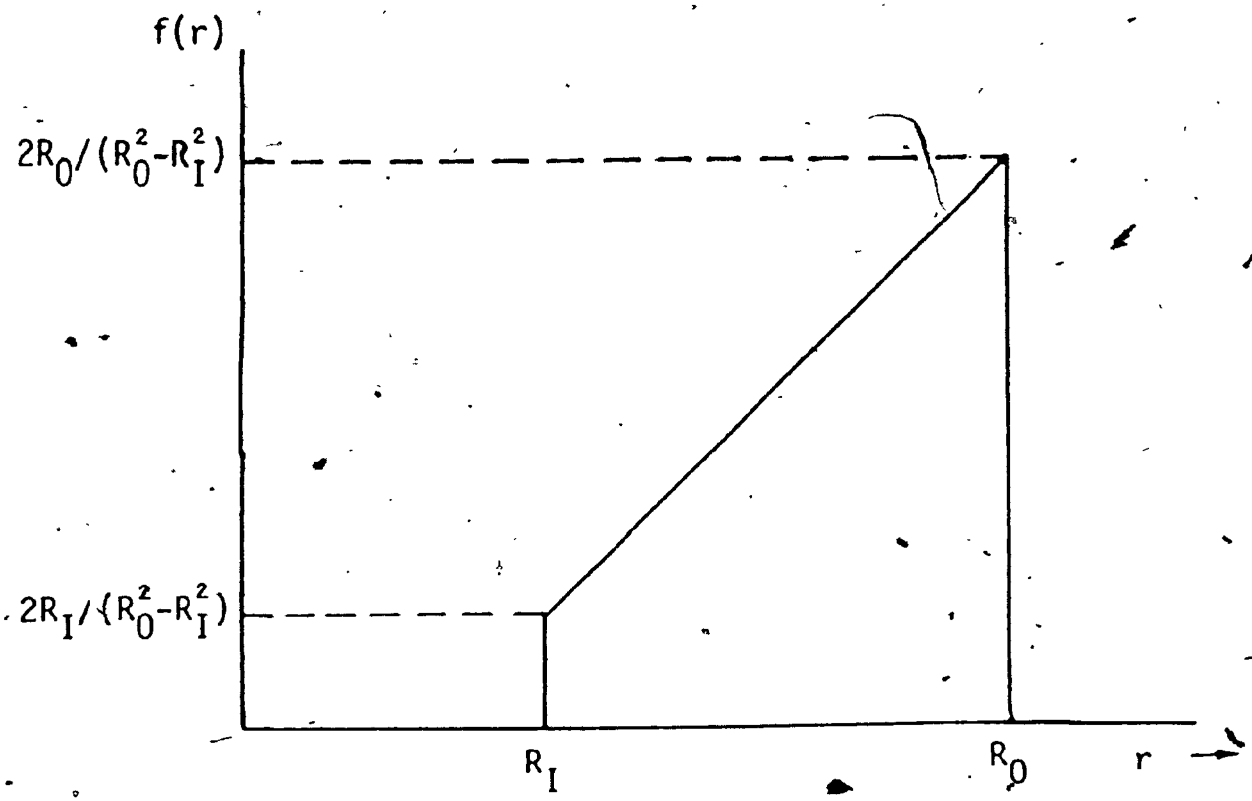


FIG. 3.4 : RADIAL DENSITY OF THE INTERFERERS WITH A UNIFORM DEPLOYMENT

interference-to-signal ratio given that the interference enters the transmission channel; and $f(y_i)$, represents the probability density function for an interference-to-signal ratio given that the interference enters one of the adjacent channels. The probability density functions are dependent on the deployment statistics.

The distance between a possible interferer and a receiver is significant in determing if the interference will cause a bit error. We assume the effect of the interferer is negligle past an outer radial distance R_0, and the minimum inner radial distance between the receiver and an interferer is R_I. Figure 3.3, shows the boundary allocation, where D is the distance between the receiver and the transmitter.

The radial density for a uniform location probability is given by,

$$f(r) = \begin{cases} \dfrac{2r}{R_0^2 - R_I^2} & R_I < r < R_0 \\ 0 & \text{elsewhere} \end{cases} \tag{3.15}$$

and is shown in Fig. 3.4.

By deriving an equation that relates deployment statistics to the interference-to-signal ratio, we can determine the density function for the interference-to-signal ratios, $f(x_i)$ and $f(y_i)$. If we define the interference-to-signal ratio as U, and assume the received power varies inversely with v^{th} power of the distance D, between receiver and transmitter, we obtain

$$U = \left(\frac{D}{r}\right)^v \tag{3.16}$$

It is posible to take the Jacobian of (3.15), since all the conditions, as in [19], are satisfied for its existence.

Taking the Jacobian of (3.15), with the relation given in (3.16), yields the expression for the density function of the interference-to-signal ratio f(U).

$$f(U) = f(r = DU^{-1/v}) \cdot \left|\frac{\partial r}{\partial U}\right| \qquad R_I < (r = DU^{-1/v}) < R_0$$

$$= \frac{2DU^{-1/v}}{(R_0^2 - R_I^2)} \left| -\frac{1}{v} DU^{-(1+v)/v} \right| \qquad \left(\frac{D}{R_I}\right)^v < U < \left(\frac{D}{R_0}\right)^v$$

$$= \begin{cases} \dfrac{2D^2\, U^{-(2+v)/v}}{v(R_0^2 - R_I^2)} & \left(\dfrac{D}{R_I}\right)^v < U < \left(\dfrac{D}{R_0}\right)^v \\ 0 & \text{elsewhere} \end{cases} \tag{3.17}$$

Typically v is equal to 4.

To evaluate the mean acquisition time using (2.33), we must first determine the detection probabilities per bank α_i, and the false alarm probabilities per bank β_i. To calculate the detection and false alarm probabilities, as in [8] we assume $\alpha_i = \alpha$ and $\beta_i = \beta$ (for all i). The detection probability per bank α is then defined by the bit error probability (i.e. the probability that the power in the complementary channel is stronger than the power in the transmission channel for a received signal R_s). If we assume each filter bank must exceed a set threshold limit, TH, Fig. 3.1, and the detection probability per bank is binomially distributed (due to the randomness of interferers), then the detection probability can be written as,

$$\alpha = \sum_{X=TH}^{m'} \binom{m'}{X} (1 - P_b)^X (P_b)^{m'-X} \tag{3.18}$$

P_b is defined by (3.1), and M' is the number of channels per matched filter bank.

Similarly, the probability of false alarm is defined by the bit error probability P_b given by (3.1). However, P_b is now the probability that the power in the transmission channel exceeds the power in the complementary channel, and is evaluated for $R_s = 0$ in (3.13) and (3.14).

Again, assuming the threshold, TH must be exceeded, and a binomial distribution, the probability of false alarm β becomes,

$$\beta = \sum_{Y=TH}^{m'} \binom{m'}{Y} (1- P_b)^{M'-Y} (P_b)^Y \qquad (3.19)$$

3.5 Results

The results are based on the operating environment outlined in section 3.4. The probability of bit error is calculated for a varying number of total available channels ℓ', a varying number of potential interferers N, a varying number of jammers J, and a varying number of channels per matched filter bank M', using the expression obtained in (3.12) to (3.19).

Setting, the limit L in (3.5) and (3.6), to two in order to minimize s computational time in evaluating P_{bE} in (3.7), (L = 2 implies we only consider the effect on mean acquisition time for two or fewer interferern hitting the transmission channel and/or adjacent channels),

and using the software routine outlined in Appendices C and D, the following results were obtained. Intuitively, we assume for L 2 the mea acquisition time should increase.

These results illustrate the behaviour of the frequency hopping acquisition technique, in terms of mean acquisition time, under various operating conditions. Comparing Figs. 3.5 - 3.7, we see that increasing the number of jamming tones causes the mean acquisition time to increase significantly, for a high threshold limit. The mean acquisition time is measured in units of the duration of a single frequency hop; since the mean acquisition time (2.33) is calculated in units of total available channels ℓ', we multiply the resulting value of the total number of available channels in order to represent the mean acquisition time in standard units, independent of ℓ'. If we compare Fig. 3.6 and Figs. 3.8 to 3.10, we see that increasing the number of potential interferers (like users), other parameters held fixed, has the same basic effect as increasing the number of jamming tones. Figures 3.15 and 3.16 illustrate the effect of increasing the number of sequence detection circuits, for a given ℓ', on the mean acquisition time. As the number of sequence detection circuits is increased for a set number of tolerable limits, the mean acquisition time is significantly reduced. If we compare the mean acquisition time for $\ell'=512$, in Fig. 3.16, for $n'=16$ and $n'=32$, we see that there is a noticable reduction in the mean acquisition time for the case of $n'=32$. This inverse relation between mean acquisition time and the number of sequence detection circuits is observed at $\ell'=128$ and $\ell'=256$, as well.

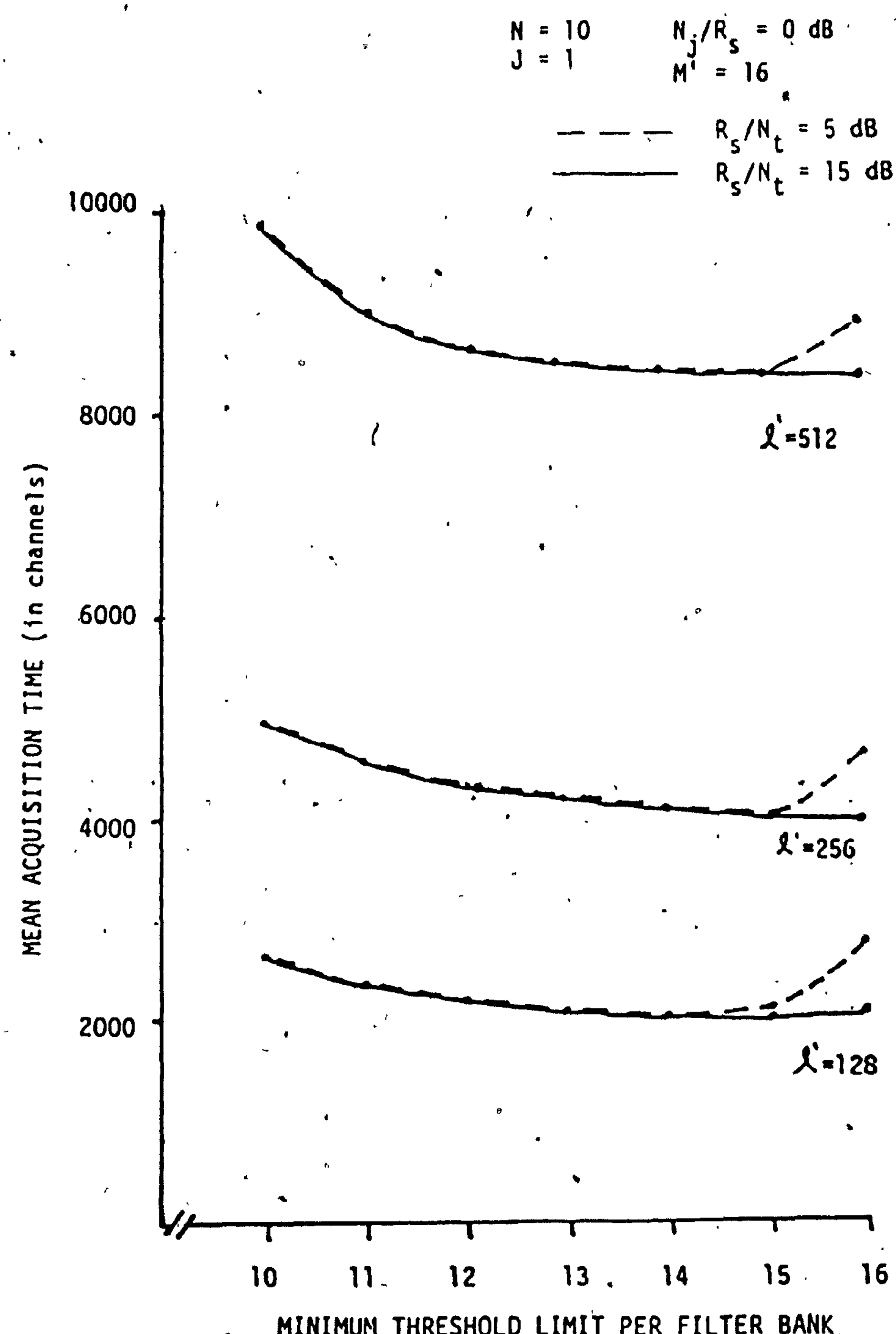


FIG. 3.5: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR VARIOUS TOTAL CHANNEL AVAILABILITY VALUES (ℓ') (POTENTIAL INTERFERS=10, JAMMING TONES=1, CHANNELS PER FILTER BANK=16)

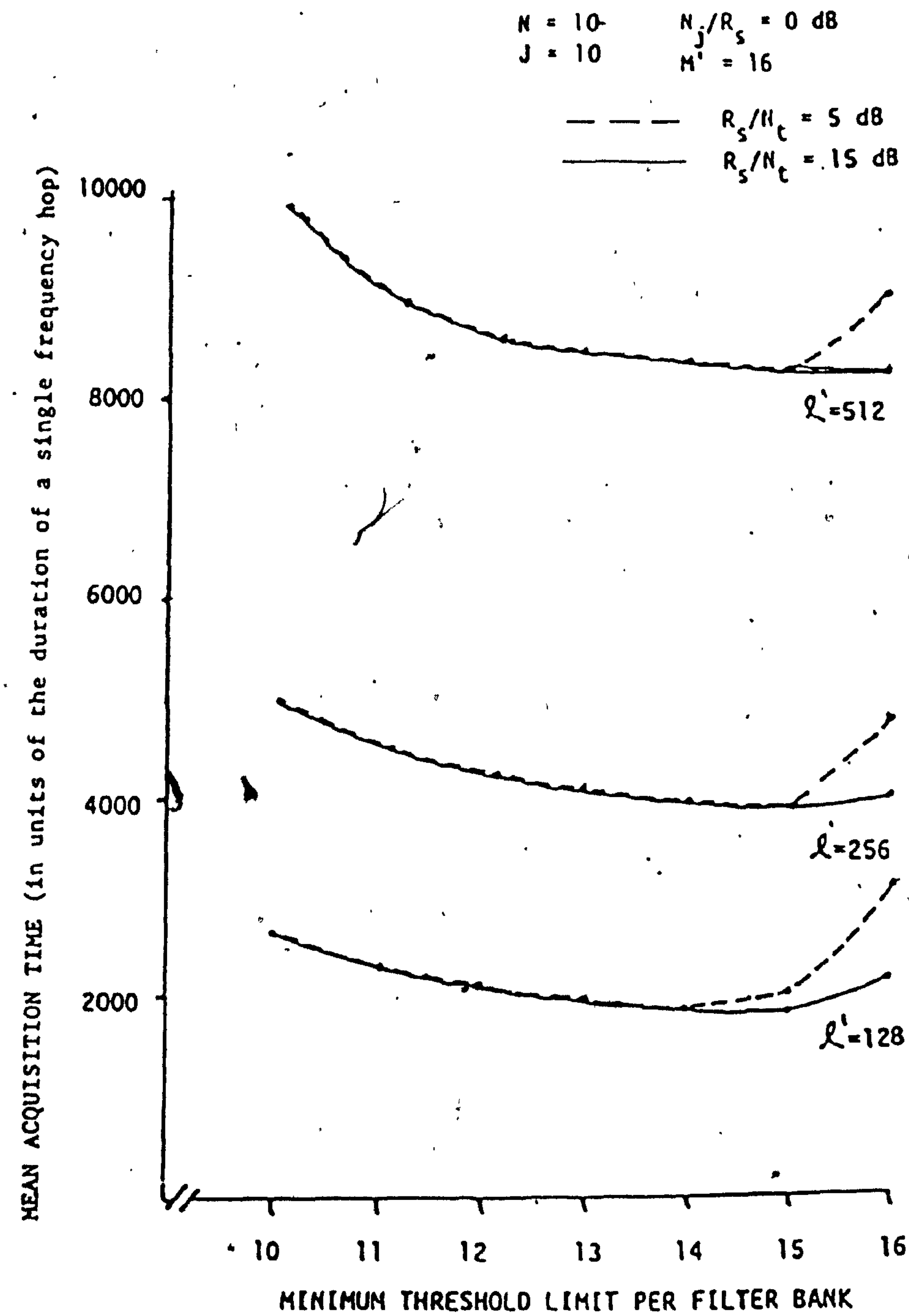


FIG. 3.6: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') (POTENTIAL INTERFERERS=10, JAMMING TONES=10, CHANNELS PER FILTER BANK=16)

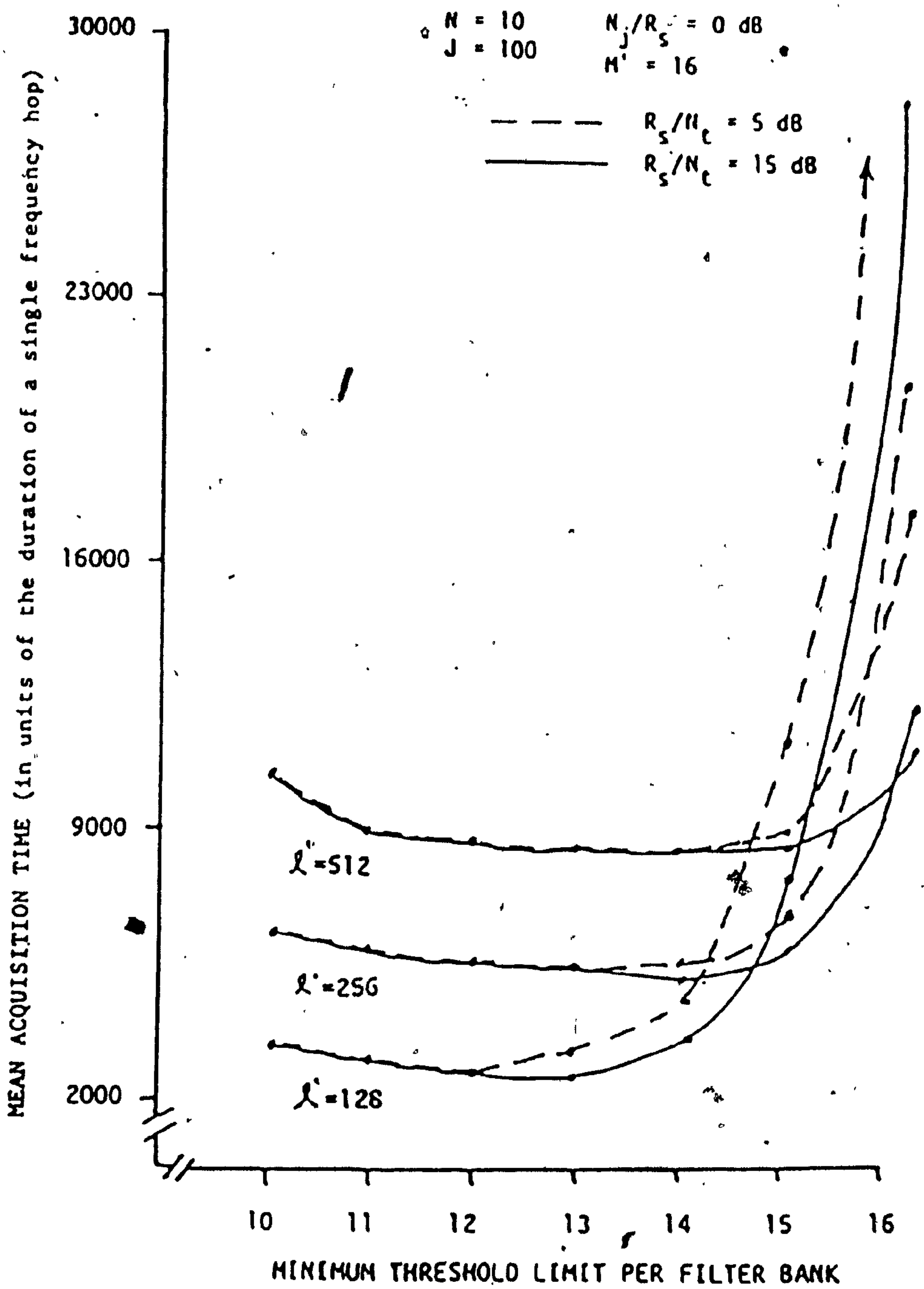


FIG. 3.7: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR VARIOUS TOTAL CHANNEL AVAILABILITY VALUES (ℓ') (POTENTIAL INTERFERERS=10, JAMMING TONES=100, CHANNELS PER FILTER BANK=16)

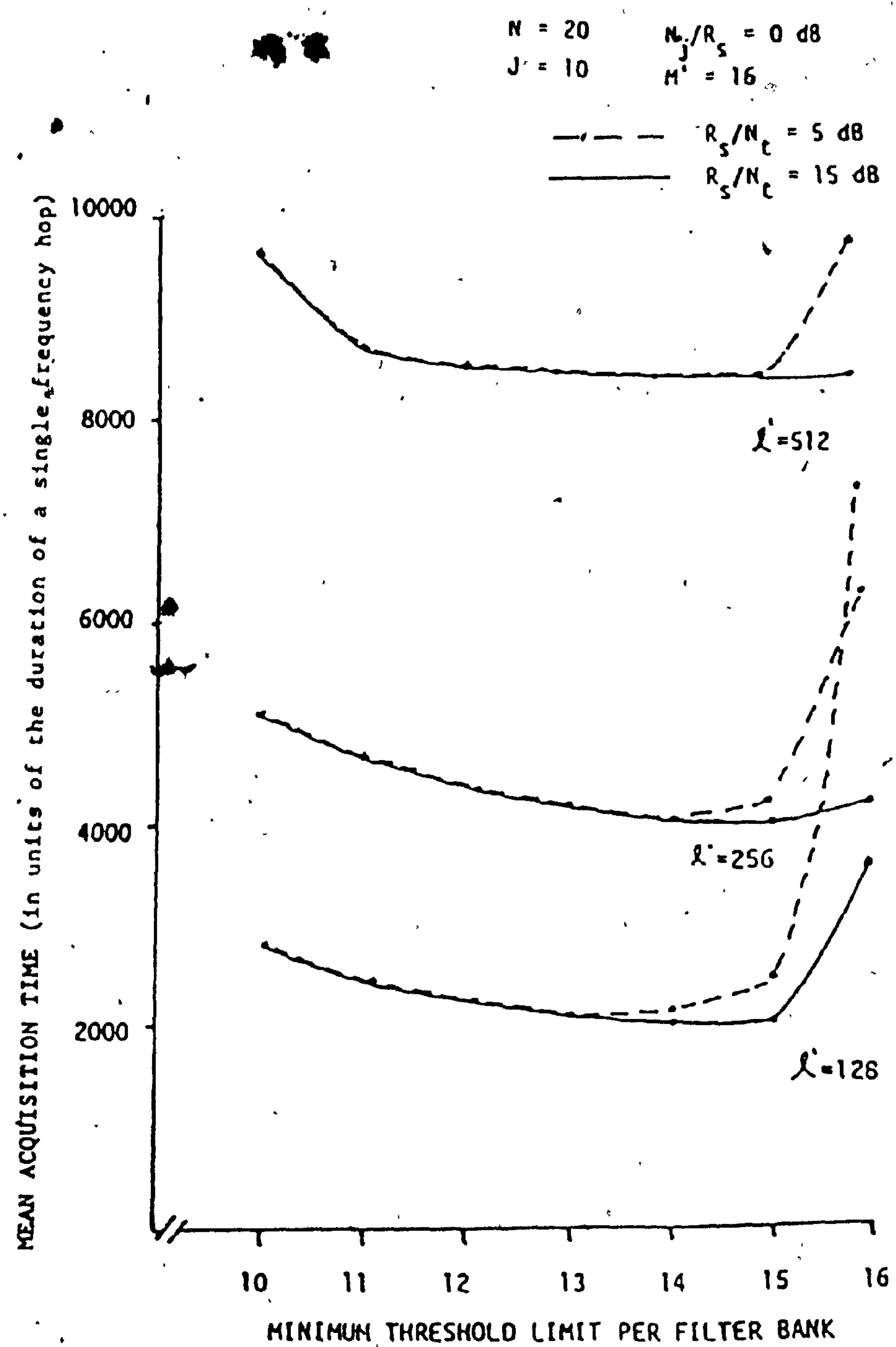


FIG. 3.8: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') (POTENTIAL INTERFERERS=20, JAMMING TONES=10, CHANNELS PER FILTER BANK=16)

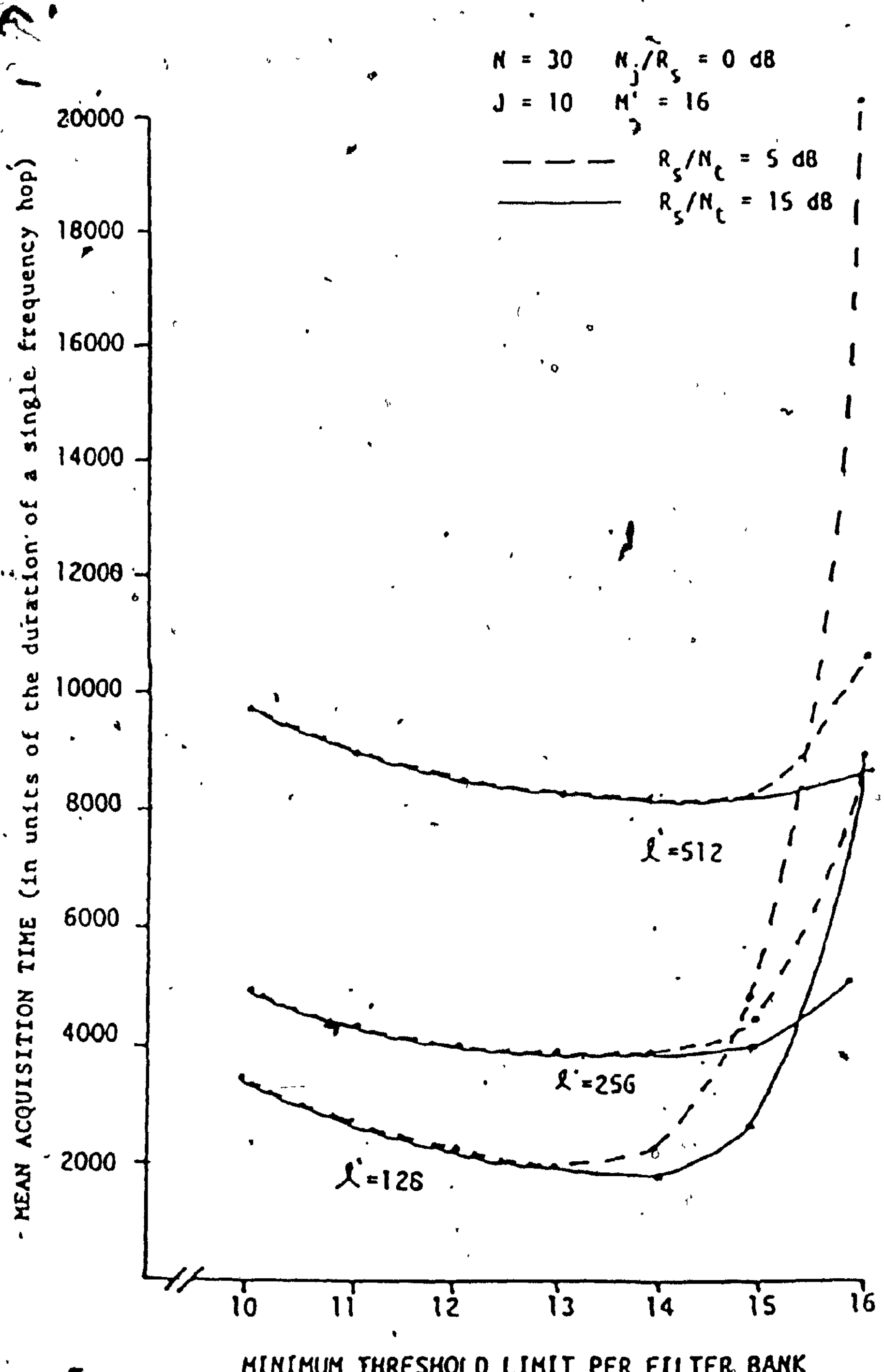


FIG. 3. 9 : EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR VARIOUS TOTAL CHANNEL AVAILABILITY VALUES (ℓ') (POTENTIAL INTERFERERS=30, JAMMING TONES=10, CHANNELS PER FILTER BANK=16)

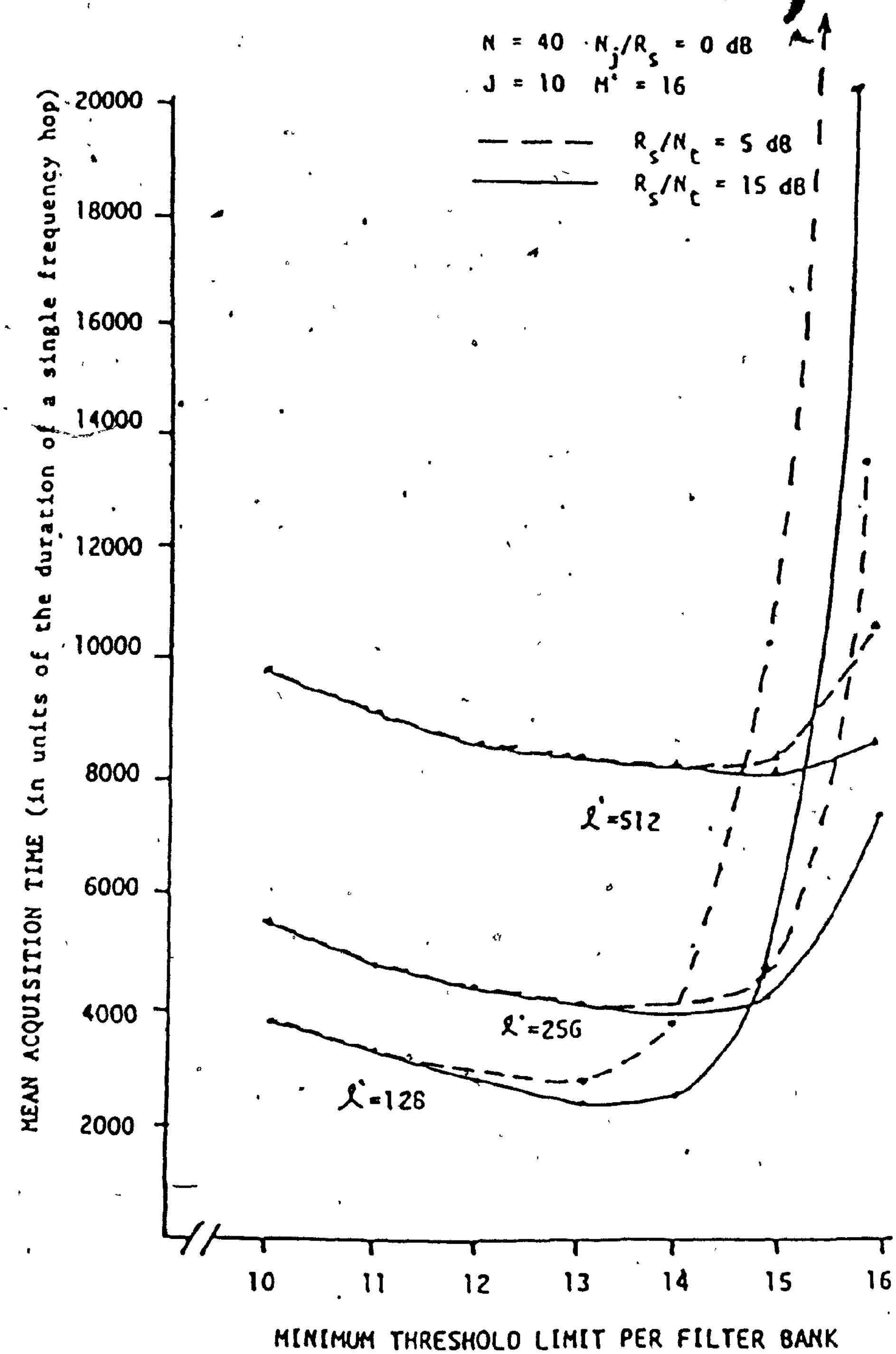


FIG. 3.10: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR VARIOUS TOTAL CHANNEL AVAILABILITY VALUES (ℓ') (POTENTIAL INTERFERERS=40, JAMMING TONES=10, CHANNELS PER FILTER BANK=16)

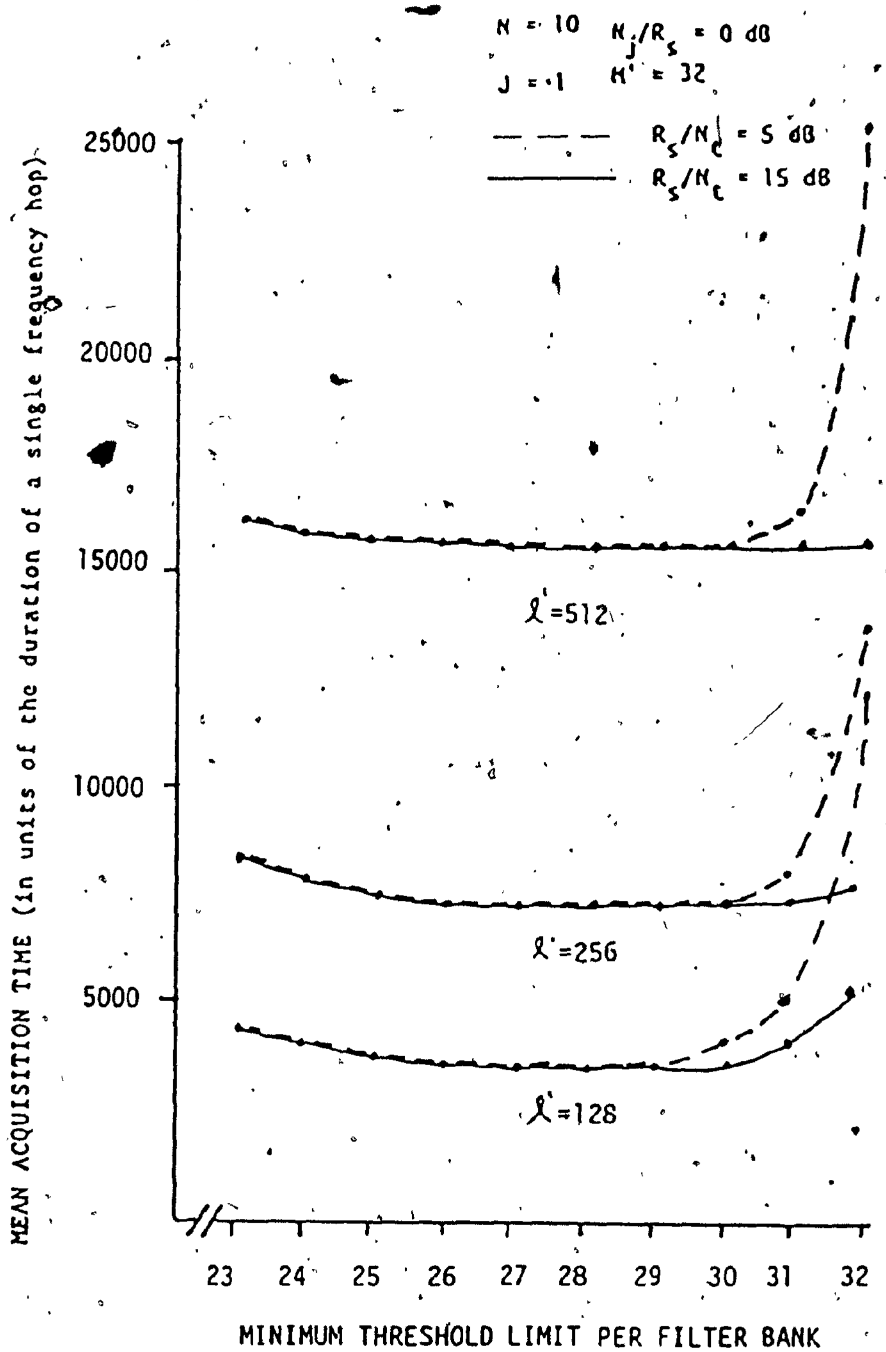


FIG. 3.11: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR VARIOUS TOTAL CHANNEL AVAILABILITY VALUES (ℓ') (POTENTIAL INTERFERERS=10, JAMMING TONES=1, CHANNELS PER FILTER BANK=32)

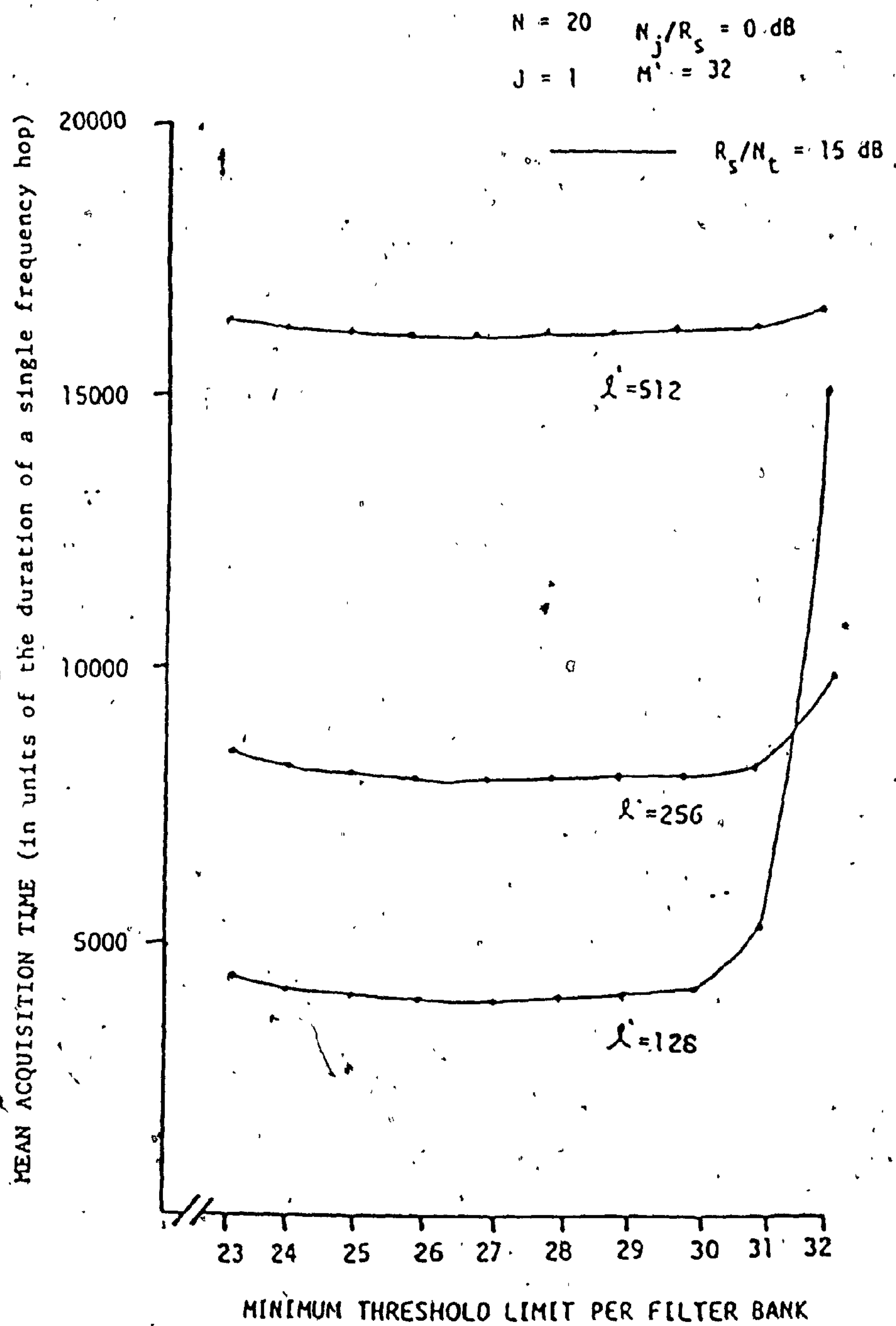


FIG. 3.12: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') (POTENTIAL INTERFERERS=20, JAMMING TONES=1, CHANNELS PER FILTER BANK=32)

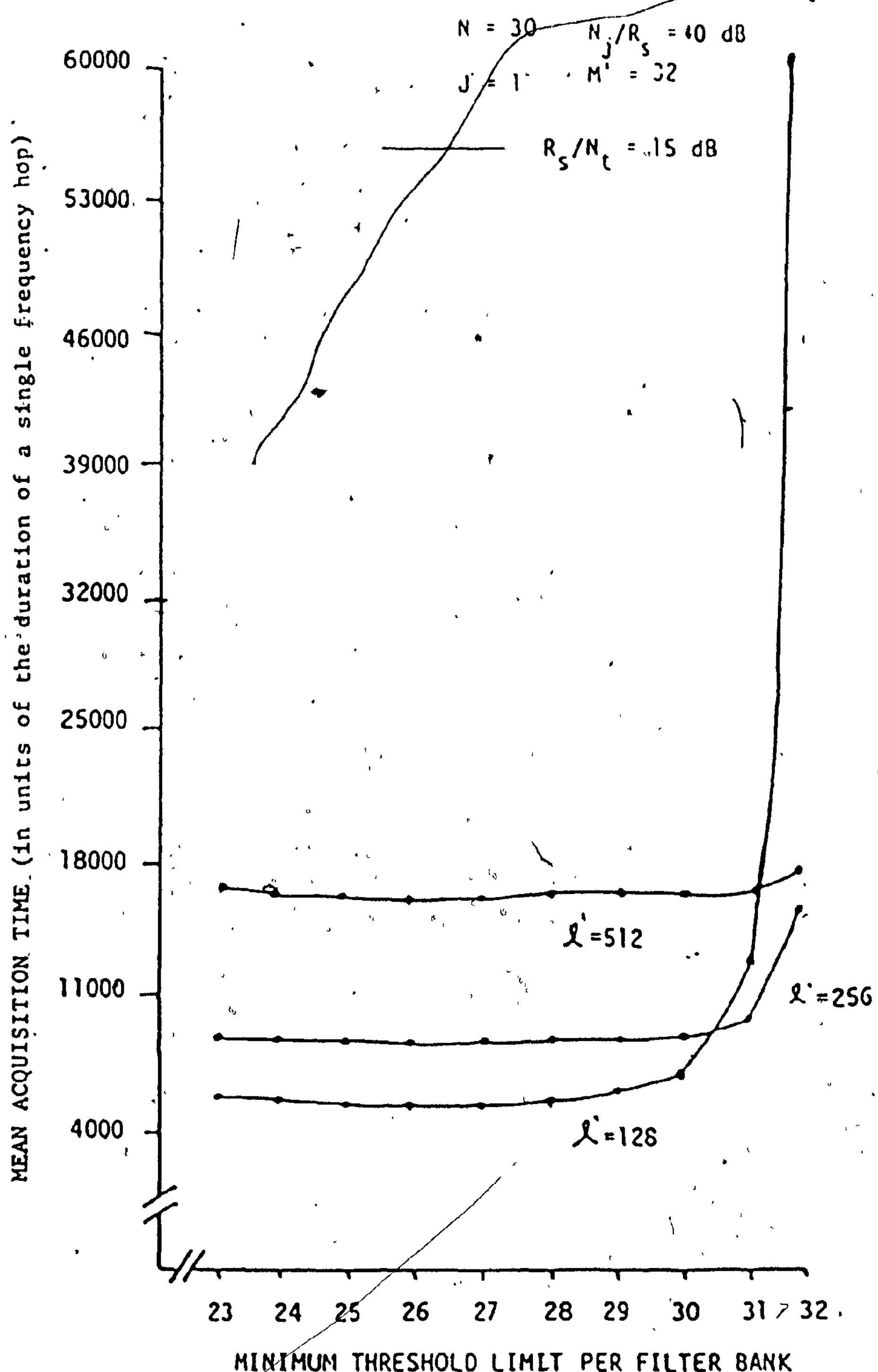


FIG. 3.13: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') (POTENTIAL INTERFERERS=30, JAMMING TONES=1, CHANNELS PER FILTER BANK=32)

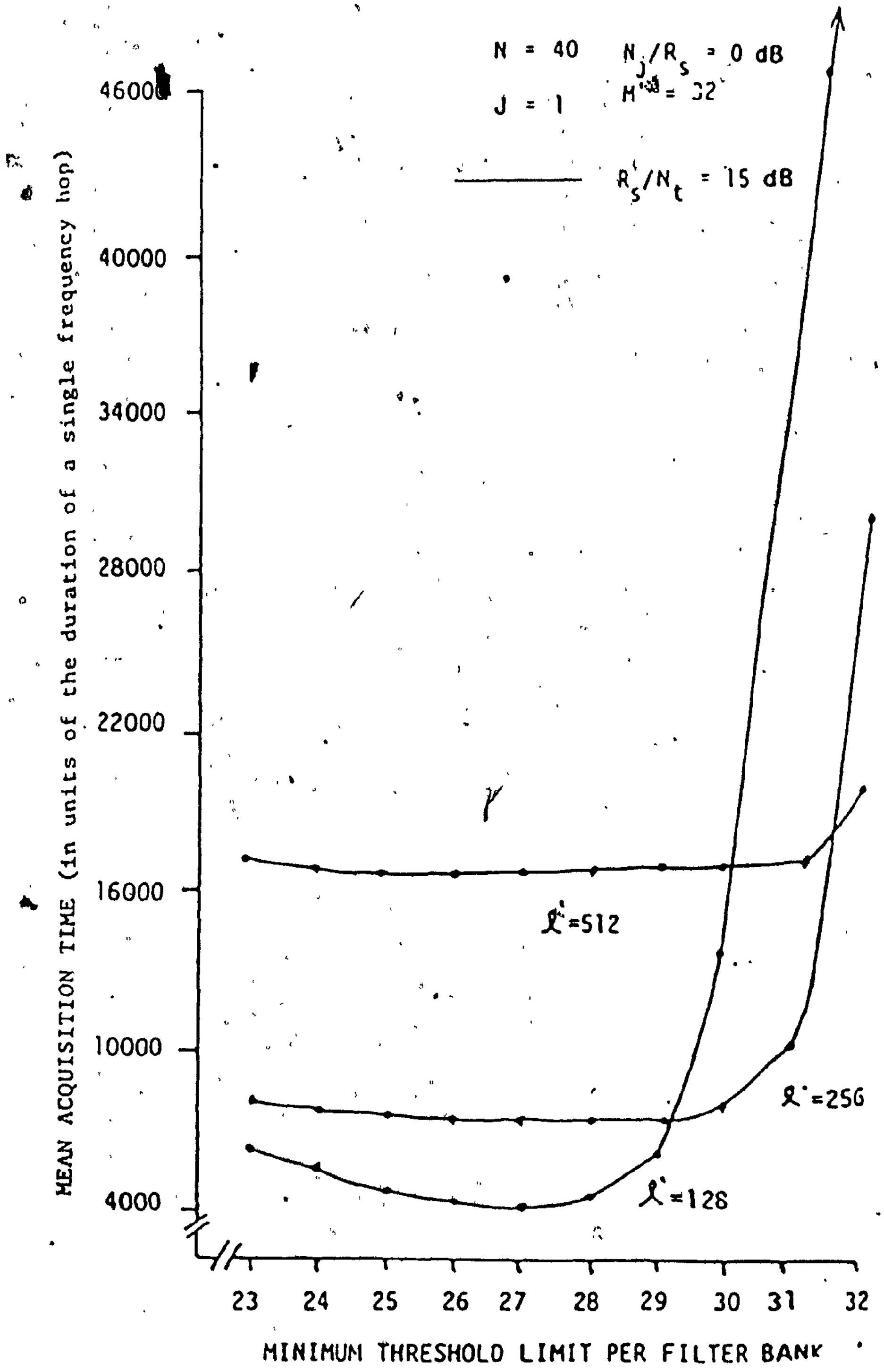


FIG. 3.14: EFFECT OF CHANGING THRESHOLD ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') (POTENTIAL INTERFERS=40, JAMMING TONES=1, CHANNELS PER FILTER BANK=32)

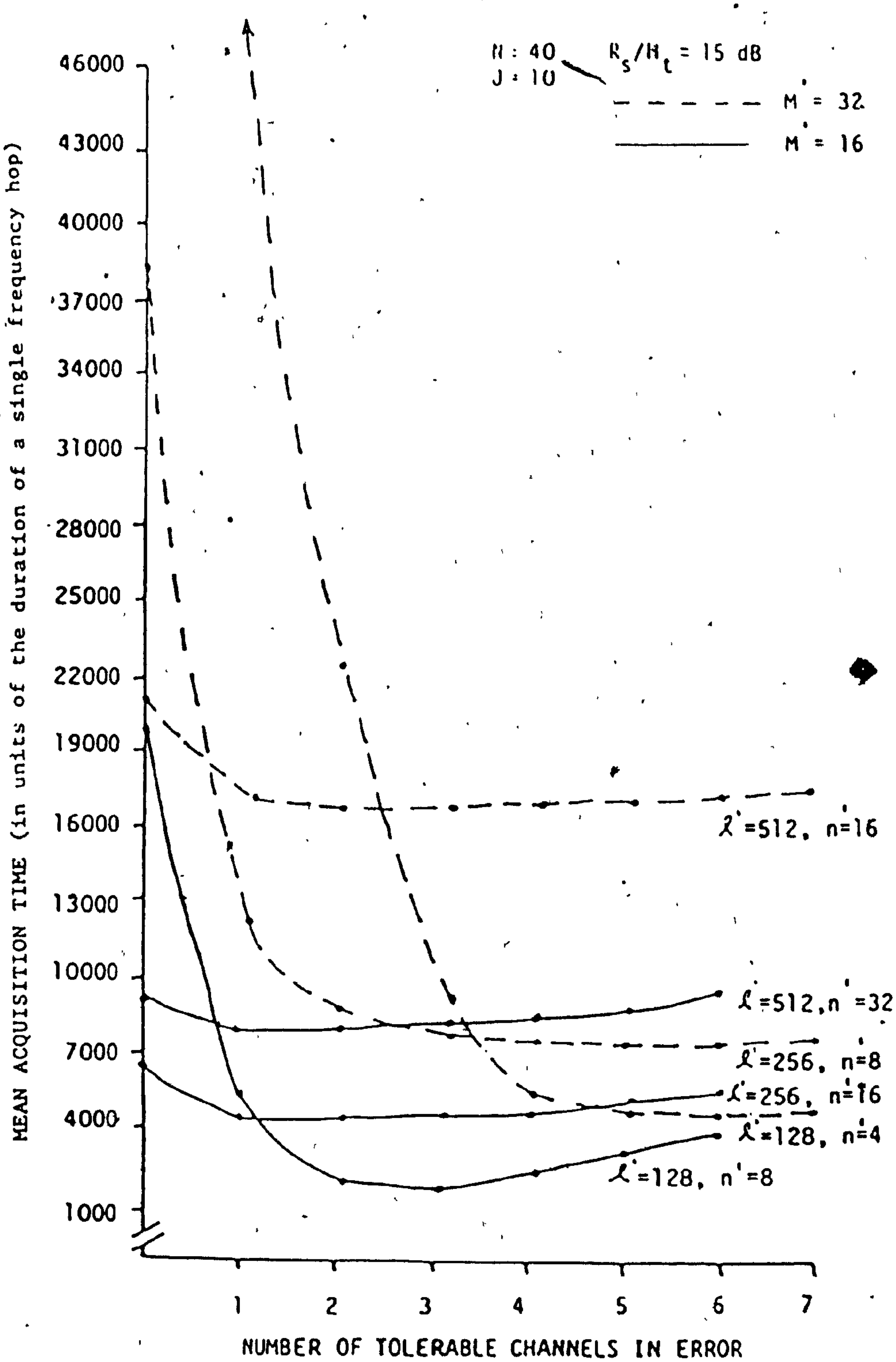


FIG. 3.15:

EFFECT OF INCREASING THE NUMBER OF SEQUENCE DETECTION CIRCUITS ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') AND CHANNELS PER FILTER BANK (M') (UNDER LARGE INTERFERENCE)

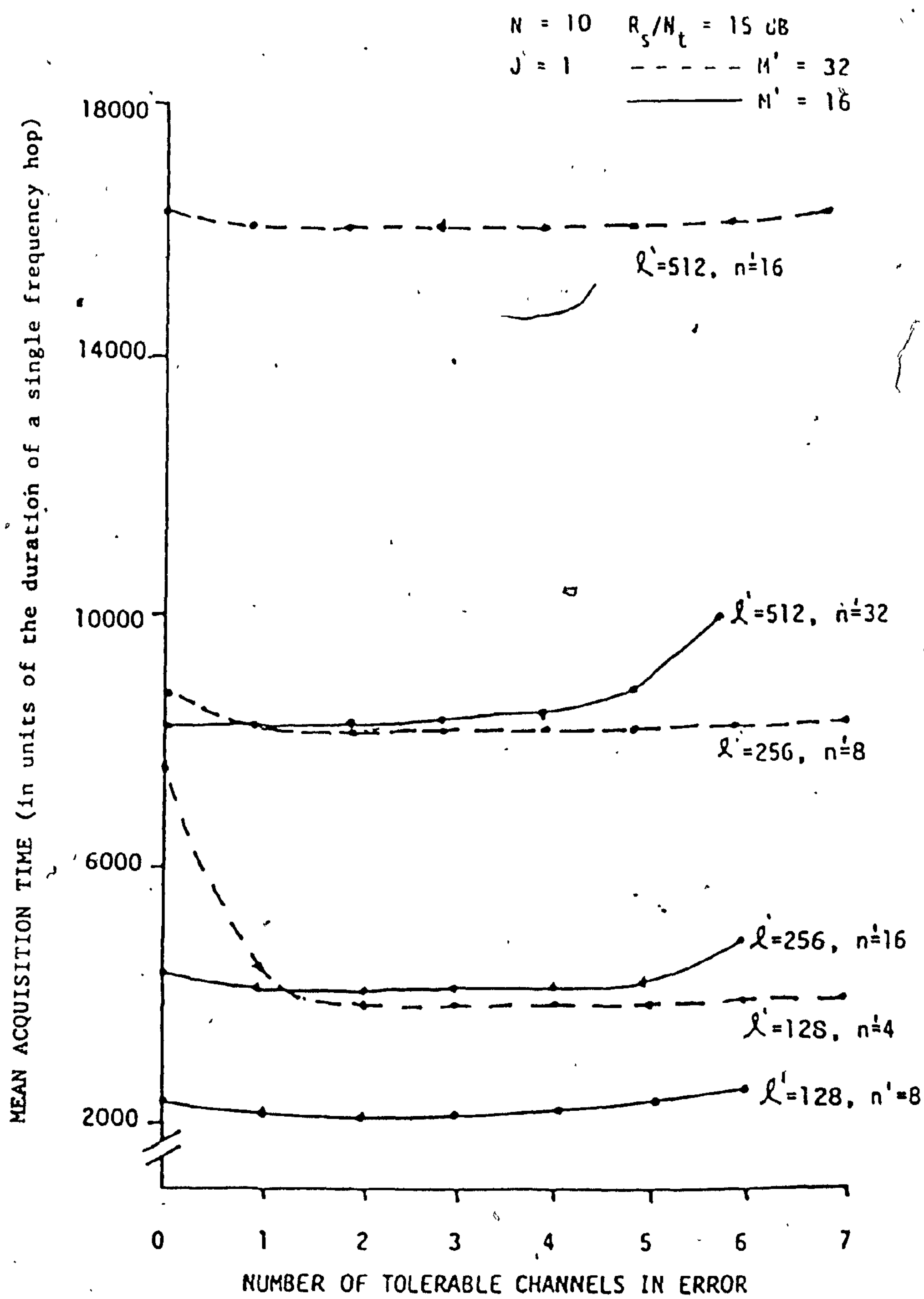


FIG. 3.16:
EFFECT OF INCREASING THE NUMBER OF SEQUENCE DETECTION CIRCUITS ON THE MEAN ACQUISITION TIME FOR A VARYING NUMBER OF TOTAL AVAILABLE CHANNELS (ℓ') AND CHANNELS PER FILTER BANK M' (UNDER LOW INTERFERENCE)

CHAPTER 4

SUMMARY, CONCLUSIONS AND RECOMMENDATIONS

In this thesis, a new technique for spread spectrum signal acquisition is proposed. The new acquisition technique is first applied to a direct sequence (DS) spread spectrum system, and its peformance, in terms of mean and standard deviation of the acquisition time, evaluated for various detection and false alarm probabilities. Secondly, with minimum modifications, the new acquisition scheme is applied to a frequency hopping (FH) spread spectrum system, and the mean acquisition time is determined for a hostile environment (spectral splatter under jamming). It is evident that the n-SDC acquisition scheme lends itself quite easily to both direct-sequence and frequency hopping applications, because of its "self-contained" nature, and also provides acquisition in a relatively short period of time.

Use of the n-SDC acquisition technique for direct sequence (DS) spread spectrum systems, has a significant effect in reducing the mean acquisition time. It was demonstrated that the reduction in the mean acquisition time was enhanced with the use of several SDCs, provided the detection probability was large. By increasing the number of sequence detection circuits in the system, the mean acquisition time can be lowered, however, there is an upper limit to the number of SDCs that can be used to reduce the acquisition time. We therefore must consider the tradeoff involved in selecting an appropriate number of sequence detection circuits for acquiring the signal. The codelength presented for the direct sequence case is acquired in the least amount of time, when 10 SDCs are use (α=0.9), however, as the number of sequence detection circuits is increased, the circuit complexity involved in implementing the system increases too. If we consider the use of 5 SDCs, to acquire the codelength, then the increase in mean acquisition time is not much more than that of the 10 SDC system, (α=0.9). Thus there is a tradeoff between the cost factor (i.e. the circuit complexity associated with using more SDCs) and the mean acquisition time. It should be noted that increasing the false alarm probability β, has little effect on the mean acquisition time, (for the results obtained in this analysis).

The extension of the n-SDC acquisition scheme to a frequency hopping (FH) system can also provide better acquisition performance. It has been shown that the mean acquisition time is reduced when more SDCs are employed to acquire the frequency hopped signal, independent of the total number of available channels. We have also observed the existence of a certain threshold level, which provides a minimum mean acquisition time for, one operating environment, but as the number of jamming tones or potential interferers increases, the same threshold value no longer leads to acquisition in a minimum amount of time. As expected, reducing the signal-to-noise ratio resulted in longer acquisition times, particularly under heavy jamming and a high threshold level, TH.

A few words are worth mentioning, at the end, concerning the complexity of the proposed multi-SDC system acquisition technique. Using off the shelf technology and CCD delay lines, a prototype model, employing 8 SDCs can be easily implemented. With VLSI technology and better resources, implementing a 32-SDC acquisition system will not pose any serious problems. The crucial rule of the acquisition problem in SS systems and the appreciable decrease in the acquisition times obtained from the new multi-SDC system technique, should justify the complexity (if any), involved in implementing a 32-SDC system.

We have seen how increasing the number of sequence detection circuits n, increases the probability of alignment and the probability of detection, yet at the same time, increases the false alarm probability. It is left for future research to implement and test performance of large (n ~ 32) and small (n ~ 8) SDC systems.

Finally, we should note that it is possible for one SDC to indicate false acquisition and move to tracking, while another SDC correctly detects acquisition, thereby causing the acquisition to be missed. For future research, a system can be designed and tested that is capable of immediately moving to tracking, once the previous false tracking has failed, and another SDC has indicated acquisition. Such a system will make full use of the multi-SDC acquisition technique benefits.

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APPENDIX A

DERIVATION OF THE VARIANCE OF THE MEAN ACQUISITION TIME

The generating function for the mean acquisition time is given by,

$$GF(z) = \frac{P_{align}\,\mu_d z}{1-[(P_{align}\,\lambda_m z)+(1-P_{align})(\mu_f z^{K+1}+\lambda_r z)} \qquad (A.1)$$

To determine the variance of the mean acquisition time the following expression

$$\sigma_T^2 = \left[\frac{\partial^2 GF(z)}{\partial z^2} + \frac{\partial GF(z)}{\partial z} - \left(\frac{\partial GF(z)}{\partial F}\right)^2\right]\Bigg|_{z=1} \qquad (A.2)$$

must be evaluated.

If we denote the numerator and denominator in (A.1) as N(z) and D(z) respectively, then the expression for the variance in (A.2) can be written as,

$$\sigma_T^2 = \frac{\partial^2 N(z)}{\partial z^2}\,\frac{1}{D(z)}\Bigg|_{z=1} - \frac{\partial N(z)}{\partial z}\,\frac{1}{D^2(z)}\,\frac{\partial D(z)}{\partial z}\Bigg|_{z=1} - \frac{\partial N(z)}{\partial z}\,\frac{1}{D^2(z)}\,\frac{\partial D(z)}{\partial z}\Bigg|_{z=1}$$

$$+\,2\,\frac{N(z)}{D^3(z)}\left[\frac{\partial D(z)}{\partial z}\right]^2\Bigg|_{z=1} - \frac{N(z)}{D^2(z)}\,\frac{\partial^2 D(z)}{\partial z^2}\Bigg|_{z=1} \qquad (A.3)$$

Substituting ($\lambda_m - 1 = \mu_d$) and ($\lambda_r = 1 - \mu_f$) into (A.1), and evaluating (A.3) yields

$$\begin{aligned}\sigma_T^2 = [&\mu_d^2 P_{align}^2 + K\mu_f\mu_d^2 P_{align}(1-P_{align}) - 2K\mu_f\mu_d P_{align}(1-P_{align}) \\ &- K^2\mu_f^2\mu_d P_{align}(1+P^2_{align}) + 2K^2\mu_f^2\mu_d P^2_{align} \\ &- P_{align}\mu_d - 2\mu_d^2 P^2_{align} - K\mu_f\mu_d^2 P_{align}^2(1-P_{align}) \\ &+ 2K^2\mu_f^2\mu_d P_{align}(1+P^3_{align}) - 4K^2\mu_f^2\mu_d P_{align}^2 \\ &+ 4K\mu_f\mu_d P_{align}(1-P_{align}) + 2\mu_d P_{align} \\ &+ K^2\mu_f\mu_d^2 P^2_{align}(1-P_{align})]/(P_{align}\mu_d)^3\end{aligned} \qquad (A.4)$$

Further simplification produces

$$\begin{aligned}\sigma_T^2 = [&-\mu_d^2 P_{align}^2 + \mu_d P_{align} - 2K^2\mu_f^2\mu_d P^2_{align} \\ &+ K^2\mu_f^2\mu_d P_{align}(1 + P^2_{align}) + 2K\mu_f\mu_d P_{align}(1 - P_{align}) \\ &+ K^2\mu_f\mu_d^2 P_{align}^2(1 - P_{align})]/(P_{align}\mu_d)^3\end{aligned} \qquad (A.5)$$

Finally the variance can be represented by

$$\sigma_T^2 = \frac{(1-P_{align}\mu_d)+[K\mu_f(1-P_{align})]^2+K\mu_f(1-P_{align})(2+K\mu_d P_{align})}{(P_{align}\,\mu_d)^2} \qquad (A.6)$$

APPENDIX B

CONDITIONAL BIT ERROR PROBABILITY FOR FSK

For completeness, the procedure for evaluating the conditional bit error probability for FSK, (Torrieri [2]) is summarized here.

If binary frequency-shift keying (FSK) is used to modulate the transmitted signal, then a demodulator, such as the one shown in Fig. B.1, is necessary for data recovery. The noise entering the bandpass filter is assumed to be additive white Gaussian (AWGN), with a flat spectrum across the filter passband. The band limited AWGN, at the output of the bandpass filter i, has the narrowband representation

$$n_i(t) = n_{ci}(t) \cos \omega_i t - n_{si}(t) \sin \omega_i t, \quad i = 1,2 \tag{B.1}$$

where ω_i is the center frequency and $n_{ci}(t)$ and $n_{si}(t)$ are independent Gaussian processes with noise powers equal to N_i.

The jamming signal at the output of the bandpass filter is given by

$$Ji(t) = B_i(t) \cos[\omega_i t + \phi_i(t)] \quad i = 1,2 \tag{B.2}$$

If we assume a binary symbol represented by

$$s_1(t) = A \cos \omega_1(t) \tag{B.3}$$

is received and is bandpassed through filter 1 with little distortion, then the total power at the outputs of bandpass filters 1 and 2 is respectively given by,

$$x_1(t) = A \cos \omega_1 t + B_1 \cos(\omega_1 t + \phi_1) + n_{c1} \cos \omega_1 t - n_{si} \sin \omega_1 t$$

and

$$x_2(t) = B_2 \cos(\omega_2 t + \phi_2) + n_{c2} \cos \omega_2 t - n_{s2} \sin \omega_2 t \tag{B.4}$$

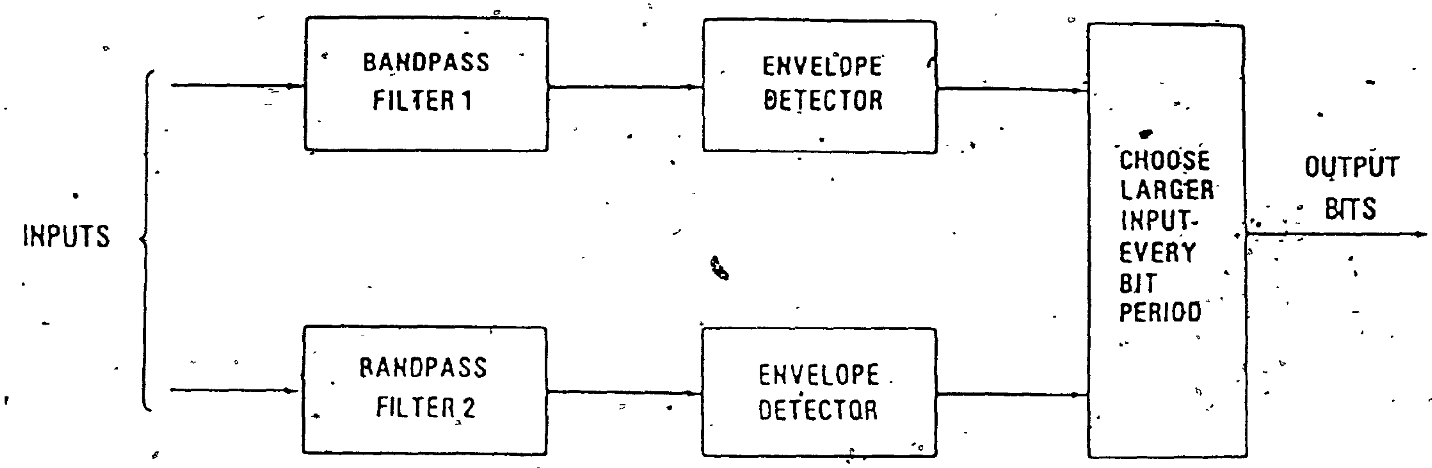


FIG. B.1: NONCOHERENT FREQUENCY-SHIFT KEYING DEMODULATOR

A typical bit interval is defined by $0 < t < T_b$, and the sampling time can occur anywhere during this interval at a time, $t = T_1$. The equations in (B.4) can be represented by $x_i(t) = R_i(t) \cos(\omega_i t + \phi_i)$, $i = 1,2$, and the outputs of the two envelope detectors at time $t = T_1$ are found to be

$$R_1 = (z_1^2 + z_2^2)^{1/2}$$

and (B.5)

$$R_2 = (z_3^2 + z_4^2)^{1/2}$$

respectively, where

$$\begin{aligned} z_1 &= A + B_1(T_1) \cos[\phi_1(T_1)] + n_{c1}(T_1) \\ z_2 &= B_1(T_1) \sin[\phi_1(T_1)] + n_{s1}(T_1) \\ z_3 &= B_2(T_1) \cos[\phi_2(T_1)] + n_{c2}(T_1) \\ z_4 &= B_2(T_1) \sin[\phi_2(T_1)] + n_{s2}(T_1) \end{aligned} \tag{B.6}$$

Denoting the mean of z_i by M_i, we have

$$\begin{aligned} M_1 &= A + B_1(T_1) \cos[\phi_1(T_1)] \\ M_2 &= B_1(T_1) \sin[\phi_2(T_1)] \\ M_3 &= B_2(T_1) \cos[\phi_2(T_1)] \\ M_4 &= B_2(T_1) \sin[\phi_2(T_1)] \end{aligned} \tag{B.7}$$

since the noise is assumed to be a zero-mean process. Assuming $B_1(T_1)$ and $\phi_1(T_1)$ are given, the joint probability density function of z_1 and z_2 is,

$$g_1(z_1,z_2) = \frac{1}{2\pi N_1} \exp\left[\frac{-(z_1-M_1)^2 - (z_2-M_2)^2}{2N_1}\right] \tag{B.8}$$

Defining $z_1 = R_1 \cos\theta$ and $z_2 = R_1 \sin\theta$, the Jacobian is equal to $|R_1|$, and hence the joint density of R_1 and θ is

$$g_2(r_1,\theta) = \frac{r_1}{2\pi N_1} \exp\left[\frac{-r_1^2+2r_1M_1\cos\theta+2r_1M_2\sin\theta-M_1^2-M_2^2}{2N_1}\right] \quad (B.9)$$

$$r_1 > 0, \ |\theta| < \pi$$

To determine the density of the envelope, R_1 we integrate (B.9) with respect to θ, yielding

$$f_1(r_1) = \frac{r_1}{N_1} \exp\left(\frac{-D_1^2-r_1^2}{2N_1}\right) I_o\left(\frac{D_1r_1}{N_1}\right), \ r_1 > 0 \quad (B.10)$$

where

$$D_1^2 = M_1^2 + M_2^2 = A^2 + B^2(T_1) + 2AB_1(T_1)\cos[\phi_1(T_1)] \quad (B.11)$$

and I_o is the modified Bessel function of the first kind satisfying

$$I_o(x) = \frac{1}{2\pi}\int_0^{2\pi} \exp[x\cos(u+v)]du \quad (B.12)$$

regardless of the value of v.

Similarly, we can obtain the density function of the output from the envelope detector in the alternate branch

$$f_2(r_2) = \frac{r_2}{N_2} \exp\left(\frac{-B_2^2-r_2^2}{2N_2}\right) I_o\left(\frac{B_2r_2}{N_2}\right), \ r > 0 \quad (B.13)$$

Since $s_1(t)$ has been transmitted, an error will result if $R_2 . R_1$. Therefore, the probability of error is

$$P(E|1) = \int_0^\infty f_1(r_1) \left[\int_{r_1}^\infty f_2(r_2)dr_2 \right] dr_1 \qquad (B.14)$$

Substituting equations (B.10) and (B.13) into (B.14) yields

$$P(E|1) = \int_0^\infty q\left(\frac{D_1}{\sqrt{N_1}}, x\right) Q\left(\frac{B_2}{\sqrt{N_2}}, \frac{\sqrt{N_1}\,x}{\sqrt{N_2}}\right) dx \qquad (B.15)$$

where the Rician function is defined as

$$q(\alpha,x) = x \exp\left(\frac{-x^2-\alpha^2}{2}\right) I_0(\alpha x) \qquad (B.16)$$

and the Q-function

$$Q(\alpha,\beta) = \int_\beta^\infty q(\alpha,x)dx \qquad (B.17)$$

By using the following identity, the integral in (B.15) can be evaluated

$$\int_0^\infty q(a,x)Q(b,rx)dx = Q(v_2,v_1) - \frac{r^2}{1+r^2} \exp\left[\frac{-a^2r^2-b^2}{2(1+r^2)}\right] \times I_0\left(\frac{abr}{1+r^2}\right) \qquad (B.18)$$

where

$$v_1 = \frac{ar}{\sqrt{1+r^2}}, \qquad v_2 = \frac{b}{\sqrt{1+r^2}}$$

Substituting the identity (B.18) into (B.15), we obtain

$$P(E|1) = Q\left(\frac{B_2}{\sqrt{N_1+N_2}}, \frac{D_1}{\sqrt{N_1+N_2}}\right) - \frac{N_1}{N_1+N_2} \exp\left[\frac{-B_2^{\,2}-D_1^{\,2}}{2(N_1+N_2)}\right] \times I_o\left(\frac{B_2D_1}{N_1+N_2}\right) \tag{B.19}$$

This expression yields P(E/1) for fixed values of $B(T_1)$ and $\phi_1(T_1)$. However, if these parameters are modelled as random variables, an overall $P(E|1)$ can be determined by integrating the product of equation (B.19) and the joint density functions of $B_i(T_1)$ and $\phi_1(T_1)$.

We assume narrowband angle-modulated jamming and thus $B_i(t) = B_i(T_1) = B_i$, a constant. The density function for $\phi_1(T_1)$ is assumed to be uniformly distributed over 0 to 2π radians, since $\phi_1(t)$ is nonsynchronous with the carrier frequency of $s_1(t)$. Thus the overall probability of error becomes

$$\bar{P}(E|1) = \frac{1}{2\pi} \int_0^{2\pi} P(E|1)\, d\phi_1 \tag{B.20}$$

The bit error probability can be determined, by an analogous manner, for a received symbol represented by

$$s_2(t) = A \cos \omega_2 t \tag{B.21}$$

Defining

$$D_2^{\,2} = A^2 + B_2^{\,2}(T_1) + 2AB_2(T_1) \cos[\phi_2(T_1)] \tag{B.22}$$

yields

$$P(E|2) = Q\left(\frac{B_1}{\sqrt{N_1+N_2}}, \frac{D_2}{\sqrt{N_1+N_2}}\right) - \frac{N_2}{N_1+N_2} \exp\left[\frac{-B_1^2-D_2^2}{2(N_1+N_2)}\right] \times I_o\left(\frac{B_1D_2}{N_1+N_2}\right) \tag{B.23}$$

If $B_i(t) = B_i(T_1) = B_i$, a constant, and $\phi_2(T_1)$ is uniformly distributed, then

$$\bar{P}(E|2) = \frac{1}{2\pi} \int_o^\infty P(E|2)\, d\phi_2 \tag{B.24}$$

Now, if the transmission of $s_1(t)$ or $s_2(t)$ is equally likely, the bit error probability becomes

$$\bar{P}(E) = \frac{1}{2}\bar{P}(E|1) + \frac{1}{2}\bar{P}(E|2) \tag{B.25}$$

Substituting (B.20) and (B.24) into (B.25), forms

$$\begin{aligned}\bar{P}(E) = \frac{1}{4\pi} \int_o^{2\pi} dx \, \{ & Q\left[\frac{B_2}{\sqrt{N_1+N_2}}, \frac{D_1(x)}{\sqrt{N_1+N_2}}\right] + Q\left[\frac{B_1}{\sqrt{N_1+N_2}}, \frac{D_2(x)}{\sqrt{N_1+N_2}}\right] \\ & - \frac{N_1}{N_1+N_2} \exp\left[\frac{-B_2^2-D_1^2(x)}{2(N_1+N_2)}\right] I_o\left[\frac{2D_1(x)}{N_1+N_2}\right] \\ & - \frac{N_2}{N_1+N_2} \exp\left[\frac{-B_1^2-D_2^2(x)}{2(N_1+N_2)}\right] I_o\left[\frac{B_1D_2(x)}{N_1+N_2}\right]\}\end{aligned} \tag{B.26}$$

where

$$D_i^{\,2}(x) = A^2 + B_i^{\,2} + 2AB_i \cos x\,, \qquad i = 1,2$$

The various conditional bit error probabilities can now be calculated form (B.26).

The probability of bit error givewn that an equal amount of jamming enters both channels, S_2, is found by setting $B_1 = B_2 = \sqrt{2R_j}$ in (B.26), where R_j is the jamming power. Setting $A = \sqrt{2R_s}$, where R_2 is the power of the received signal, yields

$$S_2 = \frac{1}{2\pi} \int_0^{2\pi} dx\{Q[\frac{2R_j}{N_1+N_2}, \frac{D(x)}{\sqrt{N_1+N_2}}] - \frac{1}{2}\exp[\frac{-2R_j-D^2(x)}{2(N_1+N_2)}]\, I_o[\frac{\sqrt{2R_j}\,D(x)}{N_1+N_2}]\} \qquad \text{(B.27)}$$

where

$$D^2(x) = 2R_s + 2R_j + 4\sqrt{R_sR_j}\cos x$$

By setting $B_i = \sqrt{2R_j}$ and $B_2 = 0$ in equation (B.26) the probability of bit error given that narrowband jamming enters a single channel, S_1, can be found. The result is simplified by using equation (B.12), $I_o(0) = 1$ and

$$Q(0,\beta) = \exp\left(-\frac{\beta^2}{2}\right) \qquad \text{(B.28)}$$

to obtain

$$S_1 = \frac{1}{2}\, Q\left(\overline{\frac{2R_j}{N_1+N_2}},\ \overline{\frac{2R_s}{N_1+N_2}}\right) \qquad \text{(B.29)}$$

where N_1 is the noise power in the jammed channel and N_2 is the noise power in the alternate (unjammed) channel.

For binary MSK with spectral splatter, we set $R_j = 0$, $R_s = \varepsilon R_s$ and $N_1 + N_2 = N_{th} + N_j + R_s x'$ in equation (B.29), to obtain

$$S_1(x') = \frac{1}{2} \exp \left(\frac{-\varepsilon R_s}{N_{th} + N_j + R_s x'}\right) \tag{B.30}$$

where ε (typically unity) is a parameter which depends on demodulation, N_{th} is thermal and backround noise power, and N, is the jamming power in the jammed channel.

The probability of error when neither channel is jammed, S_o, is given by

$$S_o = \frac{1}{2} \exp \left(\frac{-R_s}{N_1 + N_2}\right) \tag{B.31}$$

For binary MSK with spectral splatter, $R_s = \varepsilon R_s$ and $N_1 + N_2 = N_{th} + N_j + R_s x'$, where $N_j = 0$ (neither channel is jammed) in equation (B.31), thus

$$S_o(x') = \frac{1}{2} \exp \left(\frac{-\varepsilon R_s}{N_{th} + R_s x'}\right) \tag{B.32}$$

APPENDIX C

MEAN ACQUISITION TIME LISTING FOR CHAPTER 3

```
      PROGRAM  BITERR(INPUT,OUTPUT)
      REAL STDEVN, PL, PU, PU1, COMB, SMALLD, PBE, RS, JAM, PE
      REAL MEAN, VAR, STDEV, ALPHA, BETA, MUF, MUDEE, PALIGN
      REAL MEANOR, PENTIM
      INTEGER NHOP, LIM, TH, M, MPRIME, NPRIME, LPRIME, R, Z, LP
      DOUBLE PRECISION  PD, PF
C
C
C
C     ***********************************************************
C     ***********************************************************
C     ***                    VARIBLE LISTING                  ***
C     ***                                                     ***
C     ***   PL: LOWER LIMIT OF BIT ERROR PROBABILITY          ***
C     ***   PU: UPPER LIMIT OF BIT ERROR PROBABILITY          ***
C     ***   PU1: INTERMEDIATE VALUE OF PU                     ***
C     ***   COMB: FUNCTION USED TO DETERMINE THE COMBINATION  ***
C     ***         OF TWO NUMBERS                              ***
C     ***   SMALLD: DUTY FACTOR EQUIVALENT TO THE PROBABILITY ***
C     ***           THAT AN INTERFERER IS EMITTING POWER      ***
C     ***           DURING A BIT INTERVAL                     ***
C     ***   PBE: PROBABILITY OF BIT ERROR GIVEN THAT J OF THE ***
C     ***       INTERFERERS USE THE TRANSMISSION CHANNEL AND  ***
C     ***       K INTERFERERS USE AN ADJACENT CHANNEL         ***
C     ***   PE: THE PROBABILITY THAT J INTERFERERS USE THE    ***
C     ***       THE TRANSMISSION CHANNEL AND K INTERFERERS    ***
C     ***       USE AN ADJACENT CHANNEL                       ***
C     ***   RS: THE AVERAGE POWER IN THE INTENDED TRANSMISSION ***
C     ***       CHANNEL AT THE RECIEVER                       ***
C     ***   JAM: THE NUMBER OF CHANNELS THAT ARE JAMMED       ***
C     ***   NHOP: NUMBER OF POTENTIALLY INTERFERING LIKE      ***
C     ***         USERS IN THE NETWORK                        ***
C     ***   LIM: UPPER LIMIT FOR THE NUMBER OF ALLOWABLE      ***
C     ***        INTEGRATIONS                                 ***
C     ***   LPRIME: NUMBER OF AVAILABLE USER CHANNELS         ***
C     ***           EQUIVALENT TO THE CODE LENGTH             ***
C     ***    TH: THE THRESHOLD A FILTER BANK MUST EXCEED IN   ***
C     ***        ORDER TO DECLARE SYNCHRONIZATION             ***
C     ***   MPRIME: THE NUMBER OF CHANNELS (FREQUENCIES) IN   ***
C     ***           A MATCHED FILTER BANK                     ***
C     ***  NPRIME: THE TOTAL NUMBER OF MATCHED FILTER BANKS   ***
C     ***    PD: THE DETECTION PROBABILITY USED IN DETERMINING ***
C     ***        THE ACQUISITION TIME                         ***
C     ***    PF: THE FALSE ALARM PROBABILITY USED IN          ***
C     ***        DETERMINING THE ACQUISITION TIME             ***
C     ***  PENTIM: IS THE PENALTY TIME ASSOCIATED WITH        ***
C     ***          FALSE TRACKING                             ***
C     ***  MEANOR: IS THE NORMALIZED MEAN                     ***
C     ***  STDEVN: IS THE NORMALIZED STANDARD DEVIATION VALUE ***
C     ***                                                     ***
C     ***********************************************************
C     ***********************************************************
C
C
C
```

```
C
C
      PRINT 580
      DO 25     NHOP = 10,40,10
      DO 26     LP = 7,9
      LPRIME = 2**LP
      DO 27     Z = 0,2
      JAM = 10.0**Z
      SMALLD = 1.0
      MPRIME = 32
      DO 28     TH = 23,MPRIME
      PF = 0.0
      DO 29     R = 0,10
      IF (R .NE. 0  .AND.  R .NE. 1)THEN
         IF (R .NE. 10) THEN
            GOTO 29
         ELSE
            RS = 3.1623 * R
         END IF
      ELSE
         RS = 3.1623 * R
      END IF
      PL = PU1 = 0.0
      LIM = 2
      DO 10     J = 0,NHOP
      DO 20     K = 0,(NHOP-J)
C
      PE = 2**K * (1-3*SMALLD/LPRIME)**(NHOP-J-K) *
     *        (SMALLD/LPRIME)**(J+K)*COMB(NHOP,J)*COMB((NHOP-J),K)
C
      IF ((J+K) .LE. LIM) THEN
         PL = PL + PE * PBE(J,K,LPRIME,JAM,RS)
      ELSE
         PU1 = PU1 + PE * 0.5
      END IF
C
C
 20   CONTINUE
 10   CONTINUE
      PU = PL + PU1
C
      PD = 0.0
C
      IF (RS .EQ. 0.0) THEN
         DO 88  H = TH,MPRIME
         PF = PF + COMB(MPRIME,H) * PU**H * (1-PU)**(MPRIME-H)
 88      CONTINUE
         GOTO 212
      ELSE
         DO 99 H = TH,MPRIME
         PD = PD + COMB(MPRIME,H) * (1-PU)**H * PU**(MPRIME-H)
 99      CONTINUE
         GOTO 212
      END IF
C
C
 212  IF (RS .EQ. 0.0) THEN
         PRINT 590,NHOP,LPRIME,JAM
         PRINT 591,TH,MPRIME
C        PRINT 592,RS
```

```
C         PRINT 595,PL,PU
C         PRINT 600,PF
      ELSE
          PRINT 592,RS
C         PRINT 595,PL,PU
C         PRINT 601,PD
C
C
          NPRIME = LPRIME / MPRIME
          PALIGN = 1.0 / MPRIME
          PENTIM = 1.0
          MUDEE = MUF = 0.0
          CALL MUFCAL (PF,NPRIME,MUF)
          CALL MUDCAL (PD,NPRIME,MUDEE)
C
C
C         PRINT 709,MUDEE
C         PRINT 790,MUF
 709      FORMAT(' ','   THE MUDEE IS ',F15.10)
 790      FORMAT(' ','   THE MUF  IS   ',F40.35)
C
C
C
          MEAN = (1 + (1-PALIGN) * PENTIM * MUF) / (PALIGN * MUDEE)
          VAR  = (((PENTIM * MUF * (1-PALIGN))**2)+(1 - MUDEE * PALIGN)
     *               + (PENTIM * MUF * (1-PALIGN)
     *               * (2+PENTIM*MUDEE*PALIGN))) / ((PALIGN * MUDEE)**2)
C
          STDEV  =  SQRT(VAR)
          MEANOR = MEAN * LPRIME
          STDEVN = STDEV * LPRIME
C         PRINT 602,NPRIME
          PRINT 603,MEANOR
C         PRINT 604,STDEVN
      END IF
C
C
C
 29   CONTINUE
 28   CONTINUE
 27   CONTINUE
 26   CONTINUE
 25   CONTINUE
C
C
 580  FORMAT('1',10X,'  RESULTS FOR JAMMING TO SIGNAL RATIO OF ODB')
 590  FORMAT('-','  ALIKE USERS =',I4,' AVAILABLE FREQ OR CODELENGTH ='
     *          ,I5,' JAMMING FREQ = ',F7.1)
 591  FORMAT(' ','  THE THRESHOLD IS = ',I3,'    THE # OF TAPS IS ',I3)
 592  FORMAT(' ','  THE VALUE FOR SIGNAL RS IS ',F10.5)
 595  FORMAT(' ','  LOWER LIMIT IS ',F15.10,'  UPPER LIMIT IS ',F15.10)
 600  FORMAT(' ','  THE PROB OF FALSE ALARM, BETA IS ',E40.35)
 601  FORMAT(' ','  THE PROB OF DETECTION, ALPHA IS ',E40.35)
 602  FORMAT(' ','   NUMBER OF FILTER BANKS =  NUMBER OF SDC IS ',I5)
 603  FORMAT(' ','  THE MEAN ACQUISITION TIME IS ',F35.5)
 604  FORMAT(' ','  THE STANDARD DEVIATION IS ',F35.5)
C
C
 423  FORMAT(' ','   THE INTERMEDIATE VALUE PB IS',F30.25)
 424  FORMAT(' ','   THE VALUE OF PS IS ',F30.25)
```

```
C
C
C
      STOP
      END

C
C
C
C
C     ****************************************************************
C     ***     FUNCTION PBE CALCULATES THE PROBABILITY OF BIT       ***
C     ***     GIVEN THAT J INTERFERERS USE THE TRANSMISSION        ***
C     ***     CHANNEL AND K INTERFERERS USE AN ADJACENT            ***
C     ***     CHANNEL                                              ***
C     ****************************************************************
C
C
C
      FUNCTION PBE(SMALLJ,SMALLK,M,JAM,RS)
      INTEGER  SMALLJ, SMALLK, LIM, M, Y1, Y2, X1, X2
      REAL  PS1, P, F, PBE, KK, DIV, JAM, RS
      PBE = PS1 = 0.0
      KK = 0.05
      LIM = 25
      DIV = 10.0 / LIM
      IF (SMALLJ .EQ. 0  .AND.  SMALLK .EQ. 0) THEN
         PBE = P(0,M,JAM,RS)
         GOTO 444
      ELSE
         IF (SMALLJ .EQ. 1  .AND.  SMALLK .EQ. 0) THEN
            DO 30  X1 = 0,LIM
            IF (X1 .EQ. 0  .OR.  X1 .EQ. LIM) THEN
               PBE=PBE + DIV * (P(DIV*X1,M,JAM,RS) * F(DIV*X1) / 2)
            ELSE
               PBE=PBE + DIV * (P(DIV*X1,M,JAM,RS) * F(DIV*X1))
            END IF
   30       CONTINUE
            GOTO 444
         ELSE
            IF (SMALLJ .EQ. 0  .AND.  SMALLK .EQ. 1) THEN
               DO 41  Y1=0,LIM
               IF (Y1 .EQ. 0  .OR.  Y1 .EQ. LIM) THEN
                  PBE=PBE + DIV*(P(KK*DIV*Y1,M,JAM,RS) * F(DIV*Y1) / 2)
               ELSE
                  PBE=PBE + DIV * (P(KK*DIV*Y1,M,JAM,RS) * F(DIV*Y1))
               END IF
   41          CONTINUE
            ELSE
               IF (SMALLJ .EQ. 1  .AND.  SMALLK .EQ. 1) THEN
                  DO 50  X1=0,LIM
                  DO 60  Y1=0,LIM
                  IF (Y1 .EQ. 0  .OR.  Y1 .EQ.  LIM) THEN
                   PS1=PS1+DIV*(P(DIV*X1+KK*DIV*Y1,M,JAM,RS)*F(DIV*X1)
     *                   *F(DIV*Y1)/2)
                  ELSE
                    PS1=PS1+DIV*(P(DIV*X1+KK*DIV*Y1,M,JAM,RS)*F(DIV*X1)
     *                   *F(DIV*Y1))
                  END IF
   60             CONTINUE
                  IF (X1 .EQ. 0  .OR.  X1 .EQ. LIM) THEN
```

```
                  PBE = PBE + DIV * PS1/2
               ELSE
                  PBE = PBE + DIV * PS1
               END IF
   50          CONTINUE
               GOTO 444
            ELSE
               IF (SMALLJ .EQ. 2  .AND.  SMALLK .EQ. 0) THEN
                  DO 70  X1 = 0,LIM
                  DO 80  X2 = 0,LIM
                  IF (X2 .EQ. 0  .OR.  X2 .EQ. LIM) THEN
                     PS1=PS1+DIV*(P(DIV*X1+DIV*X2,M,JAM,RS)*F(DIV*X1)
                              *F(DIV*X2)/2)
                  ELSE
                     PS1=PS1+DIV*(P(DIV*X1+DIV*X2,M,JAM,RS)*F(DIV*X1)
                              *F(DIV*X2))
                  END IF
   80             CONTINUE
                  IF (X1 .EQ. 0  .OR.  X1 .EQ. LIM) THEN
                     PBE = PBE + DIV * PS1/2
                  ELSE
                     PBE = PBE + DIV * PS1
                  END IF
   70             CONTINUE
                  GOTO 444
               ELSE
                  IF (SMALLJ .EQ. 0  .AND.  SMALLK .EQ. 2) THEN
                     DO 90  Y1 = 0,LIM
                     DO 100 Y2 = 0,LIM
                     IF (Y2 .EQ. 0  .OR.  Y2 .EQ. LIM) THEN
                        PS1=PS1+DIV*(P(KK*DIV*Y1+KK*DIV*Y2,M,JAM,RS)
                              * F(DIV*Y1) * F(DIV*Y2)/2)
                     ELSE
                        PS1=PS1+DIV*(P(KK*DIV*Y1+KK*DIV*Y2,M,JAM,RS)
                              * F(DIV*Y1) * F(DIV*Y2))
                     END IF
  100                CONTINUE
                     IF (Y1 .EQ. 0  .OR.  Y1 .EQ. LIM) THEN
                        PBE = PBE + DIV * PS1/2
                     ELSE
                        PBE = PBE + DIV * PS1
                     END IF
   90                CONTINUE
                     GOTO 444
                  END IF
                END IF
              END IF
            END IF
         END IF
      END IF
  444 RETURN
      END
```

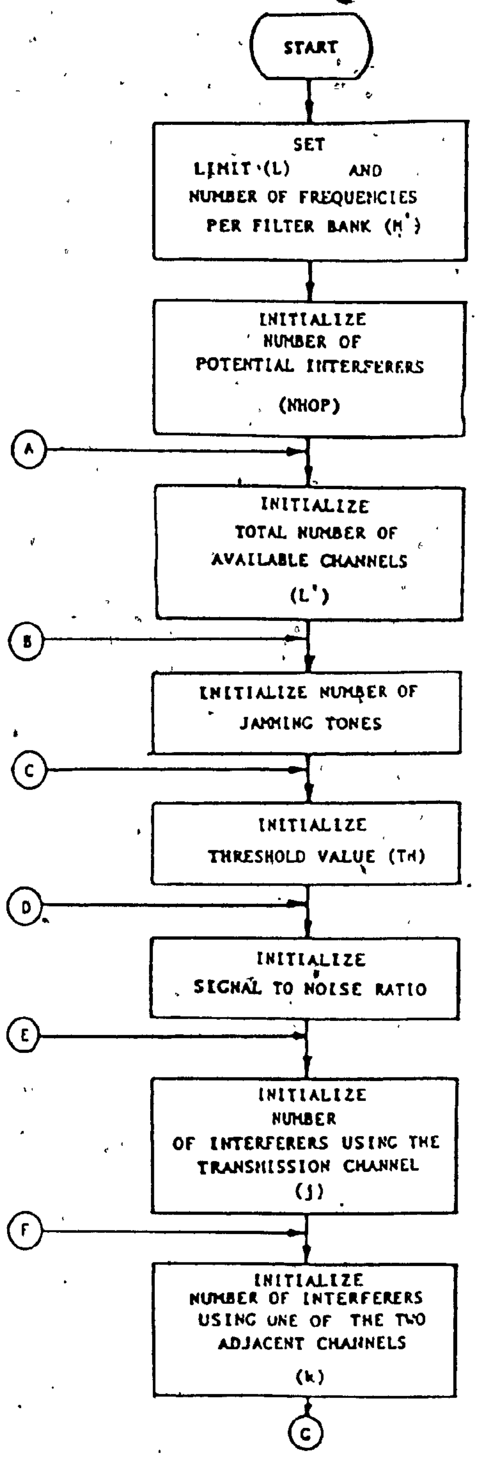
```
      C
      C
      C
      C     ***********************************************************
      C     ***       FUNCTION F CALCULATES THE DENSITY FUNCTION      ***
      C     ***       FOR AN INTERFERENCE-TO-SIGNAL RATIO GIVEN       ***
      C     ***       THAT THE INTERFERER ENTERS THE CHANNEL          ***
      C     ***********************************************************
      C
      C
            FUNCTION F(VARI)
            REAL F,VARI,D,RIN,ROUT,SMLN,LOWLIM,UPLIM
            D = 5.0
            RIN = 1.0
            ROUT = 10.0
            SMLN = 4.0
            LOWLIM = (D/ROUT)**SMLN
            UPLIM = (D/RIN)**SMLN
            IF (VARI .GT. LOWLIM  .AND. VARI .LT. UPLIM) THEN
              F = ((2*(D**2))/(SMLN*((ROUT**2) - (RIN**2))))
           *                 * VARI**(-1.0*((SMLN+2)/SMLN))
            ELSE
              F = 0.0
            END IF
            RETURN
            END
      C
      C
      C
      C
      C     ******************************************************
      C     ***     FUNCTION P CALCULATES THE PROBABILITY OF     ***
      C     ***     A BIT ERROR GIVEN VARI, A VARIABLE THAT      ***
      C     ***     REPRESENTS THE INTERFERENCE-TO-SIGNAL        ***
      C     ***     RATIO CAUSED BY J INTERFERERS USING THE      ***
      C     ***     TRANSMISSION CHANNEL AND K INTERFERERS       ***
      C     ***     USING AN ADJACENT CHANNEL                    ***
      C     ******************************************************
      C
      C
      C
            FUNCTION P(VALU,M,JAM,RS)
            REAL P,VALU,ECTA,NT,NJ,RS,JAM
            INTEGER M
            ECTA = 1.0
            NT = 1.0
            NJ = RS
            IF (RS .EQ. 0.0) THEN
               P = 0.5
            ELSE
            P = (1 - JAM/M) * 0.5 * EXP((-ECTA*RS)/(NT+RS*VALU)) +
           *    (JAM/(2*M)) * EXP((-ECTA*RS)/(NT+NJ+RS*VALU))
            END IF
            RETURN
            END
```

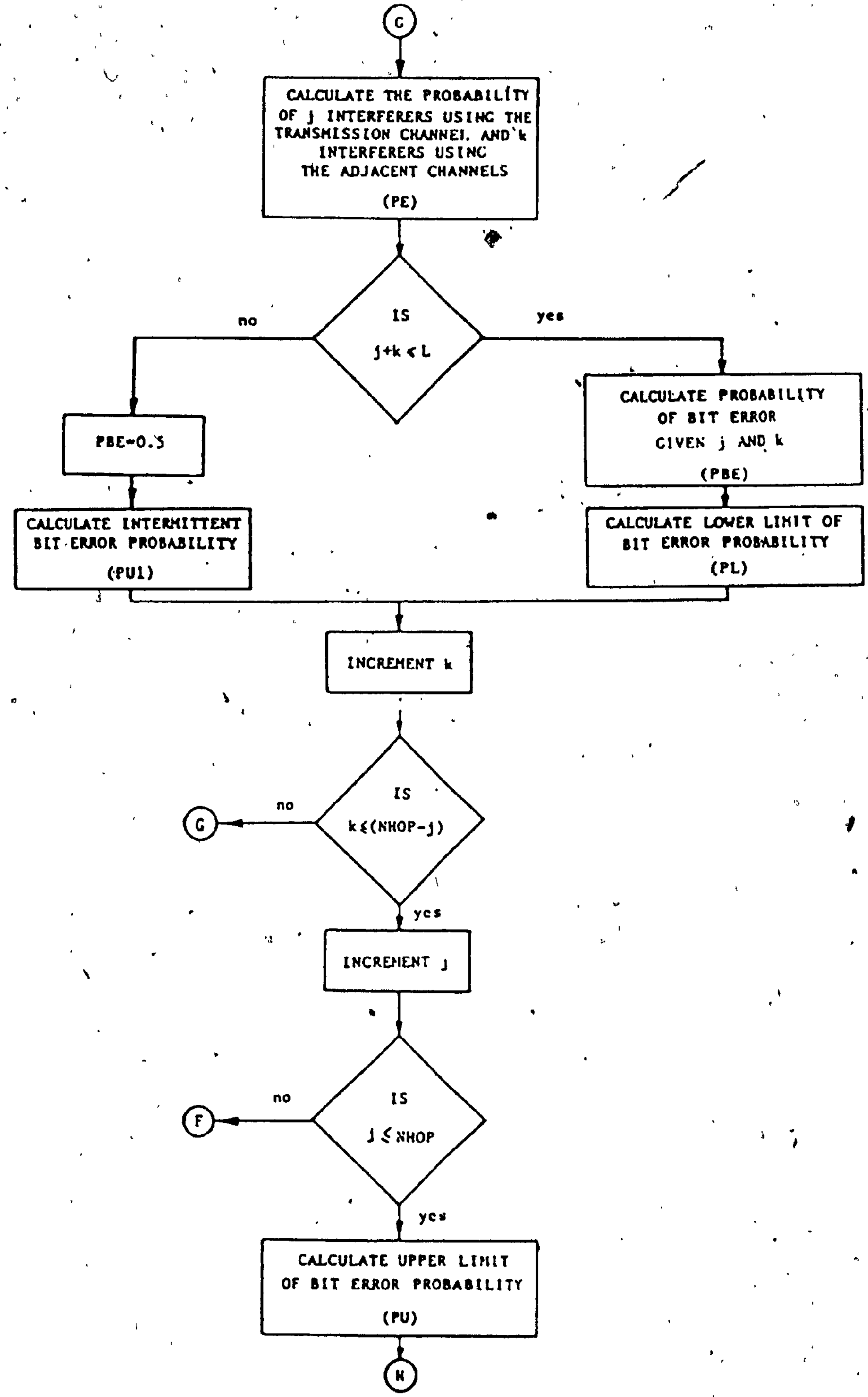
```
C
C
C
      ***************************************************************
      ***      FUNCTION FACT CALCULATES THE FACTORIAL NEEDED      ***
      ***      FOR THE EVALUATION OF THE COMBINATION OF TWO       ***
      ***                        NUMBERS                          ***
      ***************************************************************
C
C
      FUNCTION FACT(BEGIN,FIN)
      REAL FACT
      INTEGER BEGIN,FIN
      FACT = 1.0
         DO 45  N = BEGIN,FIN
         FACT = FACT * N
   45 CONTINUE
      RETURN
      END
C
C
C
      ***************************************************************
      ***       FUNCTION COMB GENERATES THE VALUE FOR THE         ***
      ***            COMBINATION OF TWO NUMBERS                   ***
      ***************************************************************
C
C
      FUNCTION COMB(NUM,DENOM)
      REAL COMB,FACT
      INTEGER NUM,DENOM,DIFF,START
      DIFF = NUM - DENOM
      IF (DENOM .EQ. 0) THEN
         COMB = 1.0
      ELSE
         IF (DENOM .GE. DIFF) THEN
            START = DENOM + 1
            COMB = FACT(START,NUM)/FACT(1,DIFF)
         ELSE
            START = DIFF + 1
            COMB = FACT(START,NUM)/FACT(1,DENOM)
         END IF
      END IF
      RETURN
      END
```

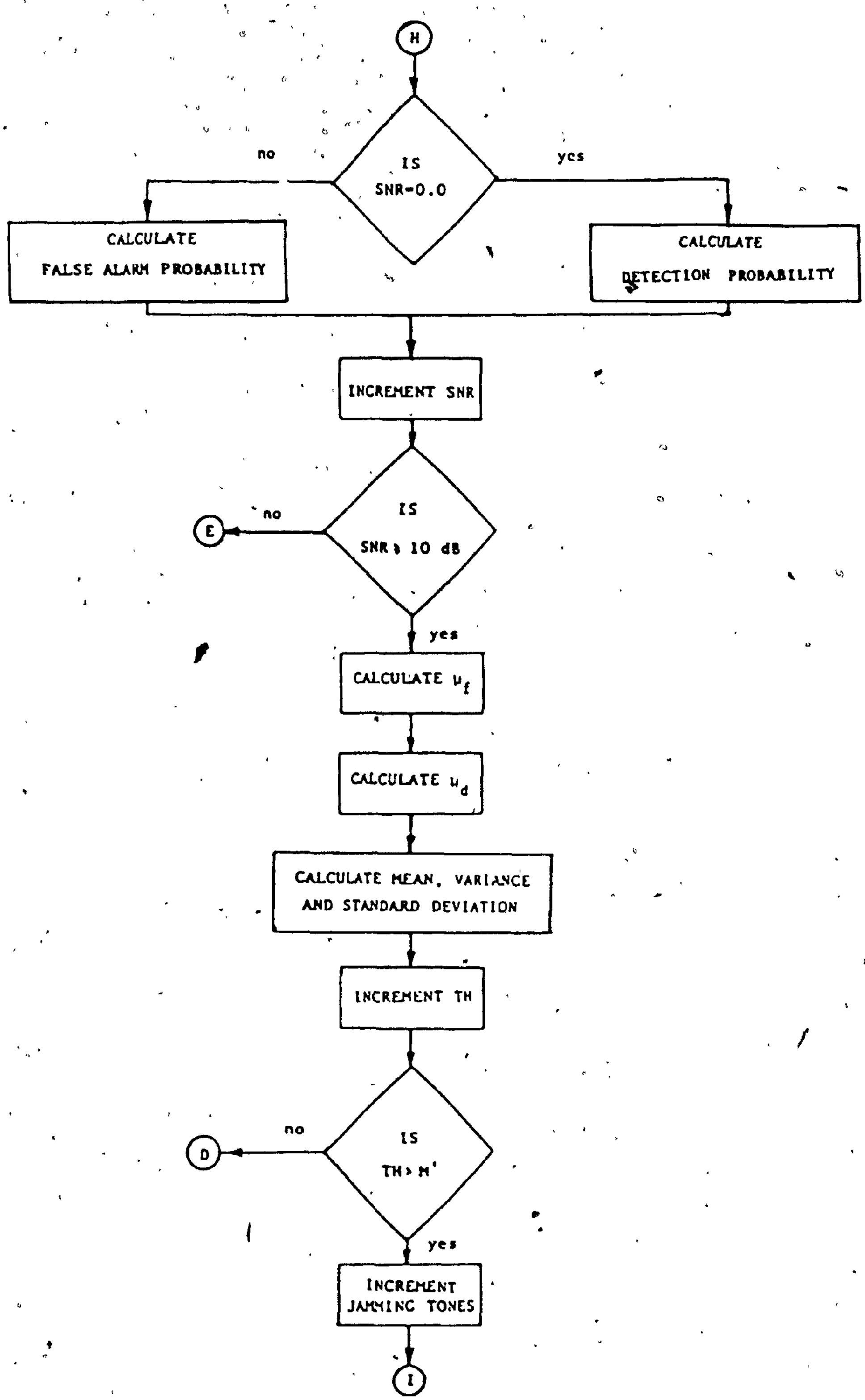
```
C
C
C
C       ****************************************************************
C       ***        SUBROUTINE MUDCAL CALCULATES THE DETECTION         ***
C       ***        PROBABILITY (MUDEE) FOR THE ENTIRE SYSTEM          ***
C       ***        GIVEN THE PROBABILITY OF DETECTION (PD OR          ***
C       ***        ALPHA) FOR EACH FILTER BANK                        ***
C       ****************************************************************
C
C
        SUBROUTINE MUDCAL(ALPHA,N,MUDEE)
           REAL ALPHA,MUDEE,PROD1
           INTEGER N
           DO 101  C=1,N
           PROD1 = 1.0
           IF (C .NE. 1) THEN
              DO 102  A=1,(C-1)
              PROD1 = PROD1 * (1-ALPHA**(N-A+1))
 102          CONTINUE
           END IF
           MUDEE = MUDEE + ALPHA**(N-C+1) * PROD1
 101       CONTINUE
           RETURN
        END
C
C
C
C       *************************************************************
C       ***     SUBROUTINE MUFCAL CALCULATES THE PROBABILITY       ***
C       ***     OF FALSE ALARM (MUF) FOR THE SYSTEM GIVEN THE      ***
C       ***     PROBABILITY OF FALSE ALARM (PF OR BETA) FOR        ***
C       ***     EACH FILTER BANK                                   ***
C       *************************************************************
C
C
        SUBROUTINE MUFCAL(BETA,N,MUF)
           REAL BETA,MUF,PROD2
           INTEGER N
           DO 103    D=1,N
           PROD2 = 1.0
           IF (D .NE. 1) THEN
              DO 104   B=1,(D-1)
              PROD2 = PROD2 * (1-BETA**(N-B+1))
 104          CONTINUE
           END IF
           MUF = MUF + BETA**(N-D+1) * PROD2
 103       CONTINUE
           RETURN
        END
```

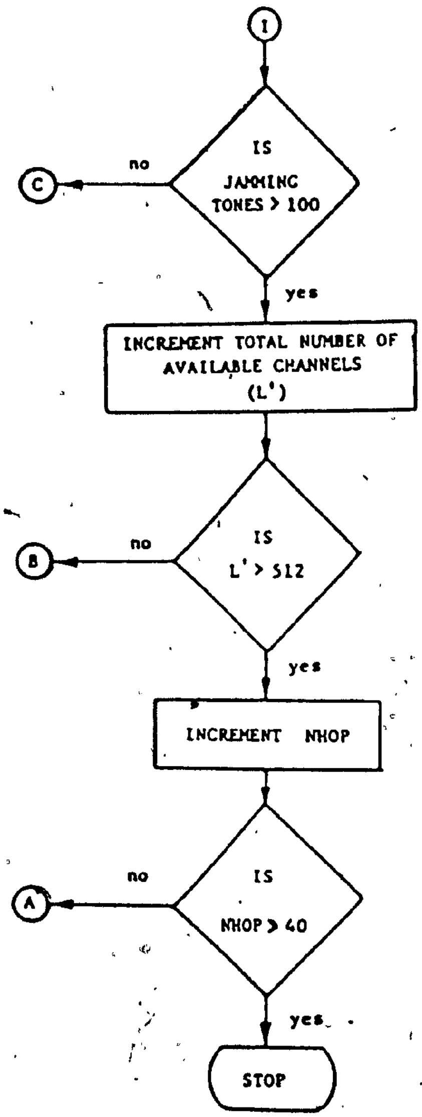
APPENDIX D

FLOW CHART OF THE MEAN AND STANDARD DEVIATION EVALUATION ROUTINE




G
CALCULATE THE PROBABILITY OF j INTERFERERS USING THE TRANSMISSION CHANNEL AND k INTERFERERS USING THE ADJACENT CHANNELS
(PE)
IS
j+k ≤ L
no
yes
PBE=0.5
CALCULATE INTERMITTENT BIT ERROR PROBABILITY
(PU1)
CALCULATE PROBABILITY OF BIT ERROR GIVEN j AND k
(PBE)
CALCULATE LOWER LIMIT OF BIT ERROR PROBABILITY
(PL)
INCREMENT k
IS
k≤(NHOP-j)
no
G
yes
INCREMENT j
IS
j ≤ NHOP
no
F
yes
CALCULATE UPPER LIMIT OF BIT ERROR PROBABILITY
(PU)
H


H
IS
SNR=0.0
no
yes
CALCULATE
FALSE ALARM PROBABILITY
CALCULATE
DETECTION PROBABILITY
INCREMENT SNR
IS
SNR ≥ 10 dB
no
E
yes
CALCULATE μf
CALCULATE μd
CALCULATE MEAN, VARIANCE
AND STANDARD DEVIATION
INCREMENT TH
IS
TH > M'
no
D
yes
INCREMENT
JAMMING TONES
I


I
IS
JAMMING
TONES > 100
no
C
yes
INCREMENT TOTAL NUMBER OF
AVAILABLE CHANNELS
(L')
IS
L' > 512
no
B
yes
INCREMENT NHOP
IS
NHOP > 40
no
A
yes
STOP