SPREAD SPECTRUM DIGITAL COMMUNICATIONS SYSTEM

by

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SPREAD-SPECTRUM COMMUNICATION SYSTEM

ABSTRACT

Spread-spectrum communications, with its inherent interference attenuation capability, has over the years become an increasingly popular technique for use in many different systems. Applications range from antijam systems, to code-division multiple access systems, to system designed to combat multipath. It is the intention of this thesis to provide a tutorial treatment of the theory of spread-spectrum communications including a computer aided analysis on Direct-Sequence and Frequency-Hopping systems. Results of the jamming rejection algorithms for estimating and suppressing narrow-band interference in DS and FH systems are presented. Techniques for determining the coefficients of a linear interference suppression filter (which are based on linear prediction and conventional spectral analysis methods), are described.

YAYGIN SPEKTRUMLU HABERLEŞME SİSTEMİ

ÖZET

Yaygın spektrumlu haberleşme, karışımı önleme, gizlilik, gürültü giderililiciliği ve aynı bandı çoklu kullanım gibi özellikleri sayesinde, son yıllarda birbirinden farklı birçok sistemlerde kullanılan gayet popüler bir teknik haline gelmiştir. Özellikle askeri haberleşmede birçok uygulaması bulunan bu sistemin kullanım alanı, karıştırma önleyici sistemlerden, kod-bölmeli çoklu kullanımlara ve çok yollu yayılmadan kaynaklanan girişimleri yoketmek için oluşturulan sistemlere kadar uzanır. Bu tezin amacı, doğrudan kodlamalı ve frekans atlamalı teknikleri kullanarak yapılan yaygın spektrumlu sistemlerin bilgisayar benzetimiyle birlikte yaygın spektrumlu haberleşme sisteminin teorisini bir bütün halinde sunmaktır. Karışımı önleyici özelliğini geliştirmek maksadıyla, doğrudan-kodlamalı ve frekans-atlamalı tekniklerde, dar-bandlı karıştırmayı ve geniş-bandlı gürültüyü tahmin edici ve bastırıcı algoritmaların sonuçları verilmiştir. Bu arada, doğrusal öngörü ve klasik frekans bandı analiz metotlarına dayanan karıştırmayı bastırıcı süzgeç hesapları yapılmış, sistemin verimliliği hesaplanmıştır.

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LIST OF SYMBOLS

AI	:	Anti-interference
AJ	:	Anti-jamming
A_m	:	Amplitude of the interference
AM	:	Amplitude Modulation
AWGN	:	Additive white gaussian noise
B	:	Bandwidth of the jammed part
B_e	:	Bandwidth expansion factor
CDMA	:	Code division multiple access
DFT	:	Discrete-fourier-transform
DS	:	Direct-sequence
D(t)	:	Data sequence
FH	:	Frequency-hopping
F_m	:	Frequency of the interference
FM	:	Frequency Modulation
FFT	:	Fast fourier transform
G_p	:	Processing Gain
JSR	:	Jamming-to-signal ratio
L	:	Losses implementation
L_c	:	Number of chips per information bit.
LFSR	:	Linear feedback shift register
M_J	:	Jamming margin
P_J	:	Jammer power
P_N	:	Average power of error
PN	:	Pseudo-noise

P_S	:	Signal power
PSK	:	Phase shift keying
P(t)	:	Code sequence
R_{data}	:	Information rate
R_{code}	:	Code rate
R_c	:	Repetition rate
SNR	:	Signal-to-noise ratio
SS	:	Spread-spectrum
SSRG	:	Simple Shift Register Generator
T_b	:	Bit interval of an information signal
T_c	:	Chip interval of a PN signal
TH	:	Time-hopping
W	:	Bandwidth of transmitted signal

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I. INTRODUCTION

Spread-spectrum systems have been developed since about the mid-1950's. This group of modulation techniques is characterized by having modulated signal spectra that do not resemble anything used before, in that they deliberately employ large bandwiths to send small amounts of information. The initial applications have been to military antijamming tactical communications, to guidance systems, to experimental antimultipath systems and to other applications. A definition of spread-spectrum that adequately reflects the characteristics of this technique is as follows :

Spread-spectrum is a means of transmission in which the signal occupies a bandwidth in excess of the minimum necessary bandwidth to send the information ; the band spread is accomplished by means of a code which is independent of the data, and a synchronized reception with the code at the receiver is used for despreading and subsequent data recovery.

Through the properties of their coded modulation. spread-spectrum systems can provide multiple access, low interference to other systems, message privacy, interference rejection and more. Spreading is accomplished by using

Direct-Sequence, Frequency-Hopping, Time-Hopping methods and hybrid combinations of these techniques.

In the second section, it is desired to provide a tutorial treatment of the theory of spread-spectrum communications. A combination of advantages not available in any other way of communicating are also given in this section.

In section III, the theory of direct-sequence spread spectrum systems is explained using the ideal and jammed communications. Digital whitening techniques for improving DS system performance in the presence of narrow-band interference are presented using the linear prediction algorithm. Power spectrum relations and performance criteria of DS system with broad-band noise and narrow-band interference are also evaluated in this section.

It is the intention of section IV. to provide the theoritical explanations of FH systems including the discussion on the digital whitening techniques and performance criteria. Effects of broad-band and narrow-band jamming is examined using the same rules as in DS systems.

In section V, it is desired to provide a computer aided analysis of spread-spectrum system using DS and FH techniques. Power spectrums of the communication signals with and without interference, are examined, plotting the input and output values of message signal. Performance criteria of the system is obtained using the SNR values of ideal and jammed signals.

II. THEORY OF SPREAD SPECTRUM COMMUNICATIONS

2.1. INTRODUCTION TO SPREAD SPECTRUM

A spectrum is a frequency domain representation of a signal. Any signal can be described in the time domain or frequency domain, and transforms are available for conversion of descriptive functions from one domain to the other and back again. Fourier transform, in which the relationship between the time and frequency domains is defined by the integral pair

$$F(f) = \int_{-\infty}^{\infty} f(t)\, e^{-j2\Pi ft}\, dt \qquad (2.1)$$

which transforms a known function of time to a function of frequency, and

$$f(t) = \int_{-\infty}^{\infty} F(f)\, e^{j2\Pi ft}\, df \qquad (2.2)$$

which performs the inverse operation.

Fourier transforms do not exist for some function because they require the existence of the integral

$$\int_{-\infty}^{\infty} f(t)\, dt \qquad (2.3)$$

Therefore discontinuous signals must be transformed by use of the laplace integral

$$L(s) = \int_{0}^{\infty} f(t)\, e^{-st} dt \qquad (2.4)$$

Even as an oscilloscope is a window in the time domain for observing signal waveforms, so is a spectrum analyzer a window in the frequency domain. Many power spectrums in this thesis are obtained by adding the squared values of imaginary and real parts of the output of the Fast Fourier Transforms. This power level is then plotted using some logarithmic plot programs. All spectral referred to are power spectra.

Literally, a spread spectrum system is one in which the transmitted signal is spread over a wide frequency band, much wider, in fact, then the minimum bandwidth required to transmit the information being sent. A voice signal, for example, can be sent, with amplitude modulation in a bandwidth only twice that of the information itself. Other forms of modulation, such as Low deviation FM or single sideband AM, also permit information to be transmitted in a bandwidth comparable to the bandwidth of the information itself. A spread spectrum system, on the other hand, often takes a baseband signal with a bandwidth of only a few kiloHertz, and distributes it over a band that may be many megaHertz wide. This is occomplished by modulating the information tobe sent with a wideband encoding signal.

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"Spread spectrum is a means of transmission in which the signal occupies a bandwidth in excess of the minimum necessary to send the information ; the band spread is accomplished by means of a code which is independent of the data, and a synchronized reception with the code at the receiver is used for despreading and subsequent data recovery."

Spreading the spectrum is the crucial point in this system. Several of the techniques are "direct-sequence" modulation in which a fast pseudorandomly generated sequence causes phase transitions in the carrier containing data, "frequency hopping", in which the carrier is caused to shift frequency in a pseudorandom way, and "time hopping", where in bursts of signal are initiated at pseudorandom times. Hybrid combinations of these techniques are frequently used.

Although the current applications for spread spectrum continue to be primarily for military communications, there is a growing interest in the use of this technique for mobile radio networks, timing and positioning systems, some specialized applications in satellites, etc. While the use of spread spectrum naturally means that each transmission utilizes a large amount of spectrum, this may be compensated for by the interference reduction capability inherent in the use of spread-spectrum techniques, so that a considerable number of users might share the same spectral band. There are no easy answers to the question of whether spread spectrum is better or worse than conventional methods for such multiuser channels. However, the one issue that is clear is that spread spectrum affords anopportunity to give a desired signal a power advantage over many types of interference, including most intentional interference.

2.2. SPREAD SPECTRUM SIGNALS FOR DIGITAL COMMUNICATIONS

Spread spectrum signals used for the transmission of the digital information are distinguished by the characteristic that their bandwidth W is much greater than the information rate R in bits per second. That is, the bandwidth expansion factor

$$B_e = \frac{W}{R} \tag{2.5}$$

for a spread spectrum signal is much greater than unity. The large redundancy inherent in spread spectrum signals is required to overcome the several levels of interference that are encountered in the transmission of digital information over some radio and satellite channels. Since coded waveforms are also characterized by a bandwidth expansion factor that is greater than unity and since coding is an efficient method for introducing redundancy it follows that coding is an important element in the design of spread spectrum signals. (1). A second important element employed in the design of spread spectrum signals is pseudo-randomness, which makes the signals appear similar to random noise and difficult to demodulate by receivers other than the intended ones. This element is intimately related with the application or purpose of such signals.

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A spread spectrum system, on the other hand, often takes a baseband signal with a bandwidth of only a few kilohertz, and distributes it over a band that may be many megahertz. This is accomplished by modulating with the information to be sent and with a wideband encoding signal.

It is common in spread spectrum systems to find transmitted RF signal bandwidths that are as much as 10^4 times the bandwidth of the information being sent. Some spread spectrum systems have employed RF bandwidths 10^5 or 10^6 times their information bandwidth.

2.2.1. MODEL OF SPREAD SPECTRUM COMMUNICATION SYSTEM

The block diagram shown in fig 2.1 illustrates the basic elements of a spread spectrum digital communications system with a binary information sequence at its input at the transmitting end and at its output at the receiving end (2). The channel encoder and decoder and the modulator and demodulator are basic elements of the system.

In addition to these elements, we have two identical pseudorandom pattern generators, one which interfaces with the modulator at the transmitting end and the second which interfaces with the demodulator at the receiving end. The generators generate a pseudo-random or pseudo-noise (PN) binary valued sequence which is impressed on the transmitted signal at the modulator and removed from the received signal at the demodulator.

Synchronization of the PN sequence generated at the receiver with the PN sequence contained in the incoming received signal is required in order to demodulate the received signal. Initially, prior to the transmission of information, synchronization may be achieved by transmitting a fixed pseudo-random bit pattern which the receiver will recognize in the presence of interference with a high probability. After time synchronization of the generators is established, the transmission of information may commence.

Interference is introduced in the transmission of the information-bearing signal through the channel. The characteristics of the interference depend to a large extent on its origin. It may be categorized as being either broadband or narrowband relative to the bandwidth of the information bearing signal, and either continuous in time or pulsed in time. For example, a jamming signal may consist of one or more sinusoids in the bandwidth used to transmit the information. The frequencies of the sinusoids may remain fixed or they may change with time according to some rule. As a second example, the interference generated in CDMA by other users of the channel may be either broadband or narrowband depending on the type of spread spectrum signal that is employed to achieve multiple access.

Our treatment of spread spectrum signals will focus on the performance of the digital communications system in the presence of narrow-band interference. (3). Two types of modulation are considered, Phase shift keying (PSK) and Frequency shift keying (FSK). PSK is appropriate in applications where phase coherence between the transmitted signal and the received signal can be maintained over a time interval that is relatively long compared to the reciprocal of the transmitted signal bandwidth. On the other had, FSK modulation is appropriate in applications where such phase coherence can not be maintained due to timevariant effect on the communications links. This may be the case in a communications link between two high-speed aircraft or between a high-speed aircraft and a ground terminal.

Spreading the spectrum is the crucial point in this system. The types of modulation in the spread spectrum are :

a) Modulation of a carrier by a digital code sequence whose bit rate is much higher than the information signal bandwidth. Such systems are called "DIRECT SEQUENCE" (DS) modulated systems.

b) Carrier frequency shifting in discrete increments in a pattern dictated by a code sequence. These are called "FREQUENCY-HOPPERS". (FH). The transmitter jumps from frequency to frequency with in some predetermined set ; the order of frequency usage is determined by a code sequence.

c) Pulsed-FM or "CHIRP" modulation in which a carrier swept over a wideband during a given pulse interval.

Closely akin to the frequency hoppers are "time hopping" and "time-frequency hopping" systems whose chief distinguishing feature is that their time of transmission is governed by a code sequence. Hybrid combinations of these techniques are frequently used.

2.2.2. PROCESSING GAIN AND JAMMING MARGIN

Interference rejection, selective addressing and codedivision multiplexing occur as a result of the spectrum spreading and consequent despreading necessary to the operation of a spread spectrum receiver. In a particular system, the ratio of the spread or transmitted bandwidth to the rate of the information sent is called the "Process Gain" of that system. A spread spectrum system develops its process gain in a sequential signal bandwidth spreading and despreading operation. The transmit part of the process may be accomplished with any one of the band-spreading modulation methods. Despreading is accomplished by correlating the received spread spectrum signal with a similar local reference signal. When the two signals are matched, the desired signal collapses to its original bandwidth before spreading, whereas any unmatched input is spread by the local reference to its bandwidth or more. A filter then rejects all but the desired narrowband signal, that is, given a desired signal and its interference, a spread spectrum receiver enhances the signal while suppressing the effects of all other inputs.

In spread spectrum processors the process gain available may be estimated by the rule of thumb :

$$\text{Process Gain} = G_p = \frac{W}{R} \tag{2.6}$$

where the RF bandwidth (W) is the bandwidth of the transmitted spread spectrum signal and the information rate (R) is the data rate in the information baseband channel. For a system in which the transmitted signal bandwidth is 20 MHz. and the information rate is 10 k bits/s, process gain would be approximately 33 dB.

$$G_p = \frac{W}{R} = \frac{2x10^7}{1x10^4} = 2000 \qquad (2.7)$$

$$G_p = 10 \log 2000 = 33 \text{ dB} \qquad (2.8)$$

This system would offer a 33 dB improvement in the signal to noise (SNR) ratio between its receivers RF inputs and its narrowband baseband output. Table 2.1 compares the process Gain that can be expected from various types of spread-spectrum systems.

This explanation doesnot mean, however, that a processor can perform when faced with on interfering signal having a power level larger than the desired signal by the amount of the available process gain. Another term, "Jamming Margin", which expresses the capability of a system to perform in such hostile environments, must be introduced.

Jamming margin is determined by a system's process gain, acceptable output signal-to-noise ratio, and implementation losses. This margin, sometimes called Antijamming margin (AJ), is the amount of interference that a receiver can withstand while operating and producing an acceptable output signal-no-noise ratio. For the above system, which has a 33-dB process gain, if the minimum acceptable output signal-to-noise ratio is 10 dB and implementation losses are 2 dB, then the jamming margin is 21 dB.

$$M_J = \text{Jamming Margin} = G_p - (\text{Lsys} + (S/N)\text{out.}) \quad (2.9)$$

$$M_J = G_p - (L+SNR) = 33-(10+2) = 33-12 = 21 \text{ dB} \quad (2.10)$$

A 21 dB jamming Margin could permit a receiver to operate in an environment in which its desired signal is 121 times samaller than the interference at its output. Expressed another way on interfering transmitter can have 121 times more power output than the desired signal's transmitter before it affects the receiver's operation.

In general, spread spectrum processing offers the most flexible means of providing unwanted signal rejection because it is not necessary to design for rejection of any particular kind of interference and geometric considerations are not normally important. Spread spectrum processing does not, however, offer the highest process gain or undesired signal rejection for every situation. The alternatives mentioned before may be combined with spread spectrum techniques to produce a compatible system with the advantages of spread spectrum and other signal-to-noise improving methods.

The "Jamming Threshold" of a particular system is of interest in determining how well that system will operate in the presence of interference. Consider the following processor model with signal and noise inputs (The noise inputs include interference)

$$\left.\begin{array}{l}\text{signal input}\\ \text{noise} \quad\quad \text{"}\end{array}\right] \quad (SNR)_{in} \longrightarrow \boxed{G_p} \longrightarrow (SNR)_{out} = (SNR)_{in} \times G_p$$

For the region of interest $S \ll N$. Converting to decibels, we have ;

$$(SNR)_{out}(dB) = (SNR)_{in}(dB) + G_p\ (dB) \qquad (2.11)$$

but,

$$(SNR)_{in}\ (dB) = -(JSR)\ (dB) \qquad (2.12)$$

Therefore

$$(SNR)_{out}(dB) = G_p(dB) - (JSR)\ (dB) \qquad (2.13)$$

For the region above the Jamming threshold, where the threshold for a particular system is that level for which (see fig 2.2)

$$\left[G_p(dB) - (JSR)dB\right] - \left[G_p(dB) - (JSR)dB\right]_{\text{measured}} = 1\ dB \qquad (2.14)$$

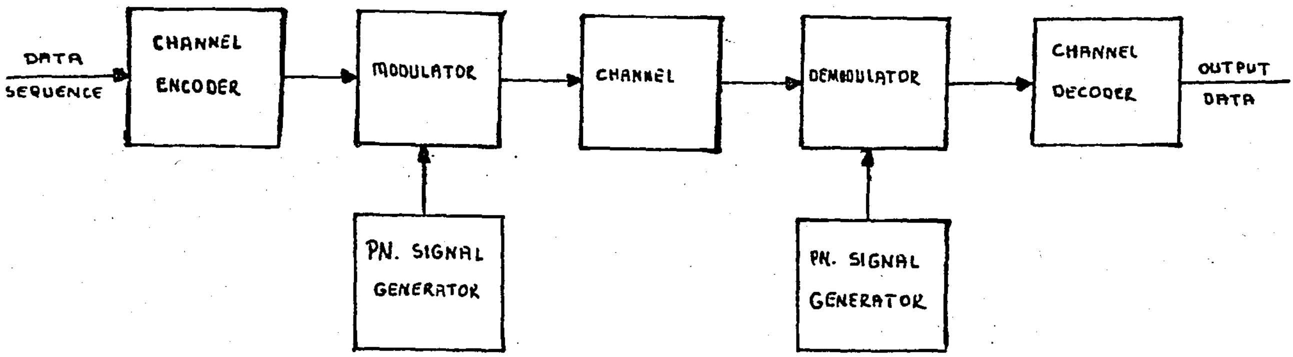


Fig. 2.1 Model of spread spectrum communucation system

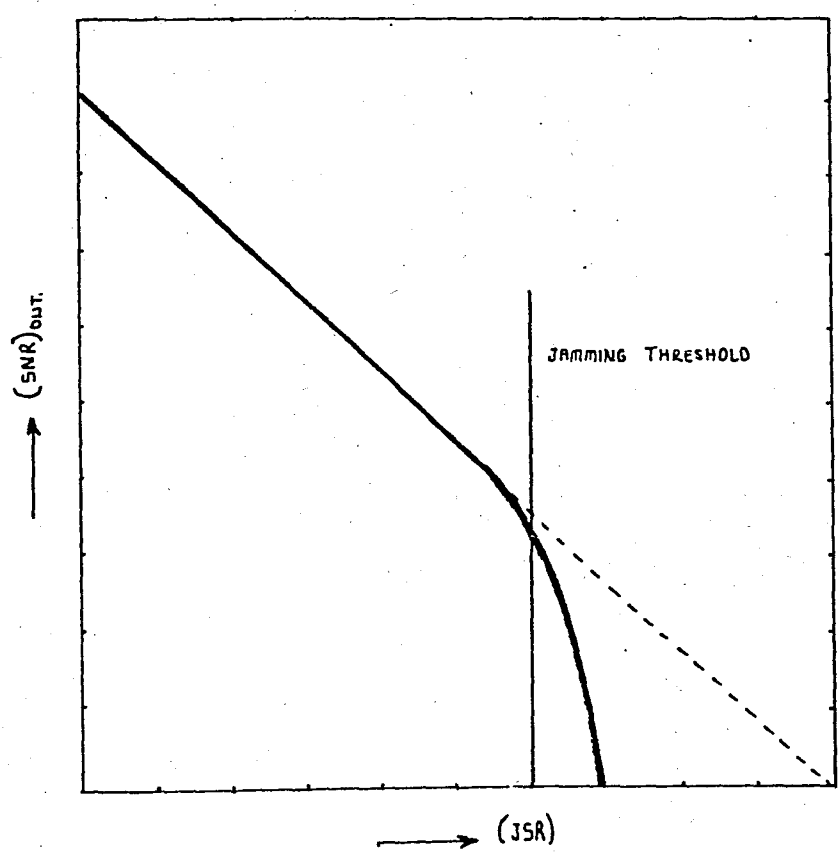


Fig.2.2 Typical process gain curve

$$G_p = (\text{SNR})\text{output} - (\text{SNR})\text{input} \tag{2.15}$$

The cause of thresholding lies in such things as tracking loss, nonlinearities, and thresholding of the postcorrelation detector. No system can be designed without a jamming threshold, but with care the threshold point can be placed beyond the normal operating region. All possible thresholding sources would tend to degrade at the same point, for when one part of the system operates at interference levels beyond the others the overall system performance does not improve.

2.2.3. WHY SPREAD SPECTRUM ?

In recent years, a new class of communications systems has grown up around a modulation technique that comes under the general classification of "spread spectrum". This group of modulation techniques is characterized by having modulated signal spectra that donot resemble anything used before, in that they deliberately employ large bandwidths to send small amounts of information in general, to be classified as spread-spectrum variety, the system must meet two criteria.

a) The transmitted bandwidth is much greater than the bandwidth of the information being sent.

b) Some function other than the information being transmitted is employed to determine the resultant transmitted signal bandwidth.

It is common in spread spectrum systems to find transmitted RF signal bandwidths that are as much as 10^4 times the bandwidth of the information being sent. Some spread-spectrum systems have employed RF bandwidths 10^5 or 10^6 times their information bandwidth.

Why would a communications engineer in his right mind even consider employing such signals in a practical system? Because the process of spreading the signal bandwith, and then collapsing it through coherent correlation with a local reference contained in the receiver offers a combination of advantages not available in any other way of communicating.

The first reaction common among those encountered spread spectrum techniques is "Why bother". The answers are varied and seldom the same. In a world beset by too little RF spectrum to satisfy the over-growing demands of military, commercial, and private users the question "Why spread spectrum" must certainly be considered valid, for spread spectrum systems have almost as many reasons for being as they have users (4). Some of the properties that may be cited are the following :

a) Interference rejection capability

Spread spectrum systems provide an interference rejection capability that can not be matched in any other way. Both deliberate and unintentional interference are rejected by a spread-spectrum receiver, up to some maximum which is known as the "Jamming Margin" for that receiver. This jamming margin is also a function of the code sequence rate or the number of frequency channels available.

b. Message Privacy

Message privacy is inherent in spread-spectrum signals because of their coded transmission format. Ofcourse, the degree of privacy, or security, is a function of the codes used. Spread-spectrum have been constructed to employ every kind of code from the relatively simple linear maximals to the truly secure nonlinear encryption types. Proper design of the system can provide for substitution as required when higher or lower level message security is desired.

c. Selective addressing

Selective addressing is possible through use of the modulating code sequences to recognize a particular signal. Assignment of a particular code to a given receiver would allow it to be contacted only by a transmitter which is using that code to modulate its signal. With different codes assigned to all of the receivers in a network, a transmitter can select any one receiver for communication by simply transmitting that receiver's code only that receiver will receive the message.

d. Code-division multiplexing

Code-division multiplexing is similar, in that a number of transmitters and receivers can operate on the same frequency at the same time by employing different codes. Either continuous transmission or time division is facilitated, since synchronization inherent to transmission and reception of spread-spectrum signals provides an excellent time base for on and off timing.

e. Low density

Low-density transmitted signals are advantageous for prevention of interference to other systems as well as for providing low probability of intercept. The low-density of spread-spectrum signals is an inherent property which exists because of the bandwidth expansion. In a Direct sequence system, for instance where the spectrum spreading code is at a 20 M bits/s. rate, the transmitted output is at least 24 MHz. wide (at the 3 dB points) and the transmitter's power is spread over this bandwidth. In that 24 MHz. band, a 10 W transmitter would average a power density of approximately 4.16 mW/Hz. To a narrowband receiver with a 50 kHz bandwidth, this 10 W signal would have less effect than a 200 mW transmitter of anything less than 50 kHz bandwidth. In addition, a spread spectrum output signal appears to be incoherent and is therefore often less objectionable than a narrow-band signal.

f) High-resolution range measurement

Spread spectrum signals of the direct sequence type excel in their capability to provide high resolution range measurement. Again, this property is due to the high-speed codes used for modulation. Since synchronization of a spread spectrum receiver depends on the receiver matching its code reference to the signal it receives to within one bit (typically, a spread spectrum receiver's code will be matched to the incoming signal's code to within one tenth to one hundredth of a bit), then the inherent resolution capability of the signal is better than the range which corresponds to a bit period. Given that same system with a 20 Mbits/s. code, the range between transmitter and receiver can easily be measured to within 50 ns, or 50 ft, and little difficulty is found in narrowing the resolution to 5 ft or less. A spread spectrum ranging system that provides 50 ft basic resolution capability at 10 miles will also provide that same resolution capability at 100 miles or 500 miles. DS ranging techniques have been more than proven on deep space probes, where they provide accurate tracking for space probes millions of milles away, In addition, spread-spectrum ranging has been employed in high performance aircraft where accurate tracking has been demonstrated at 30-mil ranges with 2-w transmitter power.

To be sure, there are disadvantages, but these are often outweighed by the advantages (5). Two prime disadvantages may be listed for spread-spectrum systems :

a) They employ more bandwidth than a more "conventionally modulated" system using AM or FM,

b) They are more complex in that thay must include code sequence generators, correlators, code tracking loops, chirp or phase-coded matched filters, or other subsystems not necessarily needed in the more conventional system.

The major systems questions associated with the design of a spread-spectrum system are : How is performance measured ? What kind of coded sequences are used ? How much jamming/interference protection is achievable ? What is the performance of any userpair in an environment where there are many spread spectrum users (code division multiple access)? To what extent does spread spectrum reduce the effects of multipath ?

It is the aim of this thesis to answers some of these questions using some computer programs. Especially on the subject of interference/jamming and performance, some algorithms are developped and results of these programs are applied to the Direct-Sequence and Frequency-Hopping systems.

Technique	Process Gain
DS	BW/R_{data}
FH	Hopped BW/R_{data}
TH	(1/duty factor)
DS/FH	$G_{pDS}+G_{pFH}$
DS/TH	G_{pDS}/duty factor
FH/TH	G_{pFH}/duty factor

Table 2.1. Spread Spectrum System Process Gain

III. DIRECT-SEQUENCE (DS) SPREAD SPECTRUM SYSTEMS

3.1. AN INTRODUCTION TO DS SIGNALS

In the model shown in Fig. 3.1, we assume that the information rate at the input to the encoder is R bits/s. and the available channel bandwidth is W Hz. The modulation is assumed to be binary PSK. In order to utilize the entire available channel bandwidth, the phase of the carrier is shifted pseudo-randomly according to the pattern from the PN generator at a rate W times/s. The reciprocal of W, denoted as T_c, defines the duration of a rectangular pulse which is shown in Fig. 3.1. This rectangular pulse is called a chip and its time duration T_c is called the chip interval. This pulse is the basic element in a PN spread spectrum signal.

If we define T_b : 1/R to be the duration of a rectangular pulse corresponding to the transmission time of an information bit, the bandwidth expansion factor W/R may be expressed as

$$B_e = \frac{W}{R} = \frac{T_b}{T_c} \tag{3.1}$$

In practical systems the ratio T_b/T_c is an integer,

$$L_c = T_b/T_c \tag{3.2}$$

which is the number of chips per information bit. That is, L_c is the number of phase shifts that occur in the transmitted signal during the bit duration $T_b = 1/R$.

Suppose that the encoder takes k information bits at a time and generates a binary linear (n,k) block code. The time duration available for transmitting the n code elements is (kT_b) seconds. The number of chips that occur in this time interval is (kL_c). Hence we may select the block length of the code as $n = kL_c$. If the encoder generates a binary convolutional code of rate k/n, the number of chips in the time interval kT_b is also $n = kL_c$.

A simple illustration of these ideas using random binary sequences will be used to bring out some of these points. Consider the transmission of a single bit d(t) with energy E_b of duration T seconds. This signal is one-dimensional as shown in Fig 3.2. and Fig 3.3., the transmitter multiplies the data bit d(t) by a binary ±1 "chipping sequence" p(t) chosen randomly at rate n chips/s. for a total of nT_c chips/bit. The dimensionality of the signal d(t)p(t) is then (nT_c). The received signal is

$$r(t) = d(t)\, p(t) + J(t) \qquad 0 < t < T \tag{3.3}$$

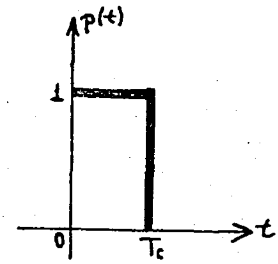


Fig.3.1 A PN chip

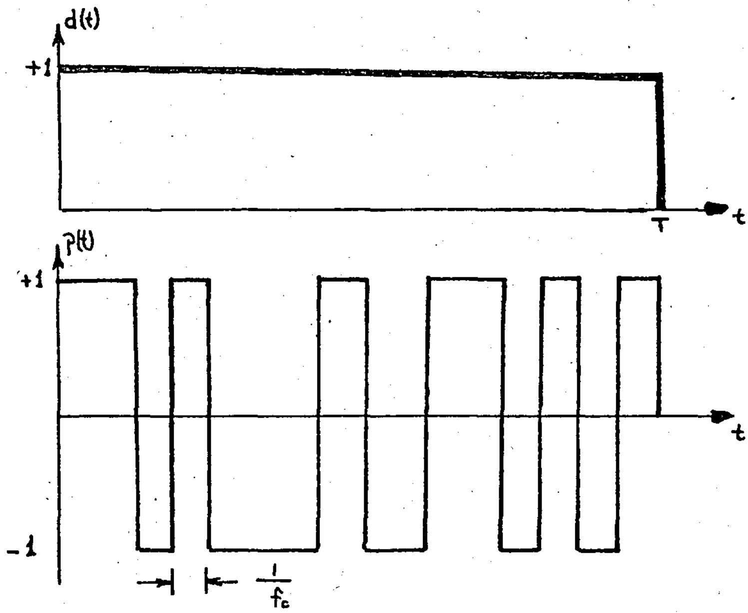


Fig. 3.2 Data bit and chipping sequence

ignoring, for the time being, thermal noise.

The receiver, as shown in Fig 3.2., performs the despreading operation

$$U \overset{\Delta}{=} \int_0^T r(t)\, p(t)\, dt \tag{3.4}$$

The integrand can be expanded as

$$\begin{aligned} r(t)p(t) &= d(t)\ p^2(t) + J(t)p(t) \\ &\simeq d(t) + J(t)p(t) \end{aligned} \tag{3.5}$$

and hence the data bit appears in the presence of a code modulated jammer. J(t) may be an additive white gaussian noise or narrow-band interference.

3.1.1. CHARACTERISTICS OF DS SYSTEMS

a) Principles and advantages of DS System

DS Modulation is just exactly the modulation of a carrier by a code sequence. In the general case the format may be AM, FM, or any other amplitude or angle modulation form. The basic form of direct sequence form is that produced by a simple, biphase-modulated (PSK) carrier. The main lobe bandwidth of the signal is twice the clock rate of the code sequence used as a modulating signal. That is, if the code sequence being used as a modulating waveform has a

5-Mbps operating rate, the main lobe bandwidth will be 10 MHz and each sidelobe will be 5 MHz wide.

In our computer programs, the code sequence being used has a 1 Mbps operating rate so 2 MHz bandwidth is obtained.

Typically, the DS biphase modulator has the form shown in Fig 3.3. A balanced mixer whose inputs are a code sequence and an RF carrier operates as the biphase modulator (6). The carrier is transmitted with one phase when the code sequence is a "one" and a 180^0 phase shift when the code sequence is a "zero" or "minus one" as in our programs.

It is worthy of note that, although other modulation forms such as PAM (pulse amplitude modulation) and FSK (frequency shift keying) could be used, biphase balanced modulation is the most common in DS systems (7). This is true for several reasons :

i. The suppressed carrier produced is difficult to detect without resorting to somewhat sophisticated methods.

ii. More power is available for sending useful information because the transmitter power is used to send only the code-produced signal.

iii. The signal has a constant envelope level so that transmitted power efficiency is maximized for the badwidth used.

iv. The biphase modulator, an extremely simple device, consist of only one pair of transformers and a few diodes.

The simplified block diagram in Fig. 3.3. illustrates a typical DS communications system. It shows that the DS system is similar to a conventional AM or FM communications system with code modulation overlaid on the carrier. In actual practice the carrier is not usually modulated by baseband information. The baseband information is digitized and added to the code sequence.

After being amplified a received signal is multiplied by a reference with the same code and, assuming that the transmitter's code and receiver's code are synchronous, the carrier inversions transmitted are removed and the original carrier restored. This narrow-band restored carrier can then flow through a band-pass filter designed to pass only the carrier.

Undesired signals are also treated in the same process of multiplication by the receiver's reference that maps the received DS signal into the original carrier bandwidth. Any incoming signal not synchronous with the receiver's coded reference is spread to a bandwidth equal to its own bandwidth plus the bandwidth of the reference.

Because an unsynchronized input signal is mapped into a bandwidth at least as wide as the receiver's reference, the bandpass filter can reject almost all the power of an undesired signal. This is the mechanism by which process gain is realized in a DS system ; that is, the receiver transforms synchronous input signals from the code-modulated bandwidth to the baseband modulated bandwidth. At the same time nonsynchronous input signals are spread at least over the code-modulated bandwidth.

b) DS Process Gain

Process Gain in a direct sequence system is a function of the bandwidth of the signal transmitted compared with the bit rate of the information. The gain in question is a signal-to-noise improvement resulting from the RF-to-information band width tradeoff.

The usual assumption taken is that the bandwidth is that of the main lobe of the DS spectrum, which is twice the bandwidth-spreading code clock rate. Therefore for our system having a 1 Mbps code clock rate, and a 1 kbps information rate the process gain would be

$$PG = \frac{2x10^6}{1x10^7} = 2x10^3 = 33 \text{ dB}$$

Let consider for a moment the limitations that exist with respect to expanding the bandwidth ratio arbitrarily so that process gain may be increased indefinitely. (Unfortunately physical limitations prevent this increase.) Only two parameters are available to adjust process gain (8).

i. Bandwidth depends on the code rate used. If we wish to have an RF bandwidth 100 MHz wide, the code clock rate would be 50 Mbps. On this basis, how wide is it practical to make the system RF bandwidth? At present, integrated circuits are available which allow limited code generation at rates up to 200 or even 300 Mbps. Is there a profit in going higher, or for that matter, using this bit rate? Consider higher bit rates : doubling the present state of the art code rates would increase process gain by only 3 dB, which is at best a modest gain when compared to the effort required to double the operating speed of present circuits.

ii. Another consideration is that as code rates go higher, operating errors must go lower in inverse proportion. To be operationally useful, a code generator should be able to operate for hours or evendays without error. On the other hand, high speed logic circuits tend toward noise sensitivity and are more susceptible to error. Finally, we can say that high-speed digital circuits consume large amounts of current and their power dissipation is high.

Equipment implementation, and propagation constraints, tend to limit the code rates used for band spreading. For the near future it appears that code rates of 50 to 100 Mbps are the highest that are practical for general use.

Data rate reduction to improve process gain is limited in its extent by the willingness of a user to slow his information transfer and by the overall stability of the transmission link. Once data rate is slowed to the tens of bits-per-second area or lower such things as local oscillator phase noise or instability in the propagation medium become significant and can cause errors.

Now, if we consider systems with code rate limited at 100 Mbps and data rate limited at 10 bps, we see that for practical systems process gain is limited at about

$$PG = \frac{2x10^8}{10} = 2x10^7 = 73 \text{ dB}$$

It is realized that future systems will transmit data at gigabit rates.

RF bandwidth in DS systems directly affects system capabilities. Several approaches are useful in choosing the proper bandwidth in a signal-hiding application ; the interest is in reducing the power transmitted percycle of bandwidth, and wide bandwidths are used. When maximum processgain for interference rejection is desired, BW again should be large.

c) Bit rate and code length.

Code bit rates in spread spectrum systems affect their systems in many ways. The most obvious is in a direct-sequence system, in which the transmission bandwidth is a direct function of the code bit rate. (i.e. Main lobe bandwidth is twice the code bit rate.) Code repetition rate is also a function of bit rate ; that is, code repetition rate is simply

$$R_c = \frac{\text{Clock Rate in bits per second}}{\text{Code length in bits}} \tag{3.6}$$

This repetition rate determines the line spacing in the RF output spectrum and is an important consideration in a system design.

One criterion for selecting code repetition rate is that the period of the code must exceed the length of any mission in which it is to be used. In most aircraft, for instance, an eight-hour code period would exceed the flight capability. Table 3.1 list the various code lengths for a 1 Mbps bit rate. Other considerations that bear on the choice of code rate and length are the relationship of the repetition rate to the information baseband and use of the system for ranging.

It is advisable that a DS systems code repetition rate be adjusted by choosing a satisfactory code length so that it will not be in the information band. Otherwise, unnecessary noise will be passed into the information demodulators, especially under Jamme conditions (9).

In our computer programs Code clock Rate is 1 Mbps and information rate is 1 kbps. After selecting a given code length, 8191 bits, we select a 13-stage shift register generator from the table. Using these values we can calculate the Repetition rate of our system.

$$R_c = \frac{\text{Code rate}}{\text{Code length}} = \frac{1\text{x}10^6}{8191} = 122.085 \text{ bps.} \qquad (3.7)$$

Repetition rate can be adjusted by choosing code length, for code rate must be chosen to give a particular RF band-width (2 MHz) in this system.

3.1.2. PSEUDO-RANDOM SEQUENCE ASPECTS

a) Introduction to PN sequences

The importance of the code sequence to a spread spectrum communications is difficult to overemphasize, for the type of code used, its length, and its bit rate set bounds on the capability of the system that can be changed only by changing the code.

To be sure that the point is understood, a spread spectrum system is not a secure system unless the codes used are cryptographically secure, and although they are useful in spread spectrum systems linear codes are not secure (9).

The PN sequences that are used for the spreading in any system must meet the two critical criteria of :

i. Denying any information about future sequence k-tuples to the unintended party.

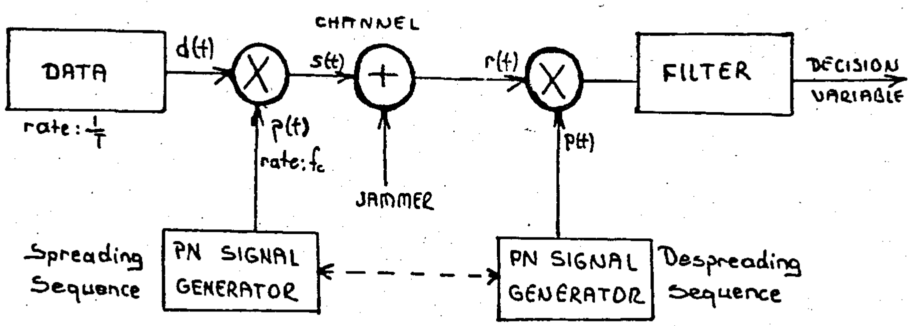


Fig.3.3 Direct Sequence spread spectrum system

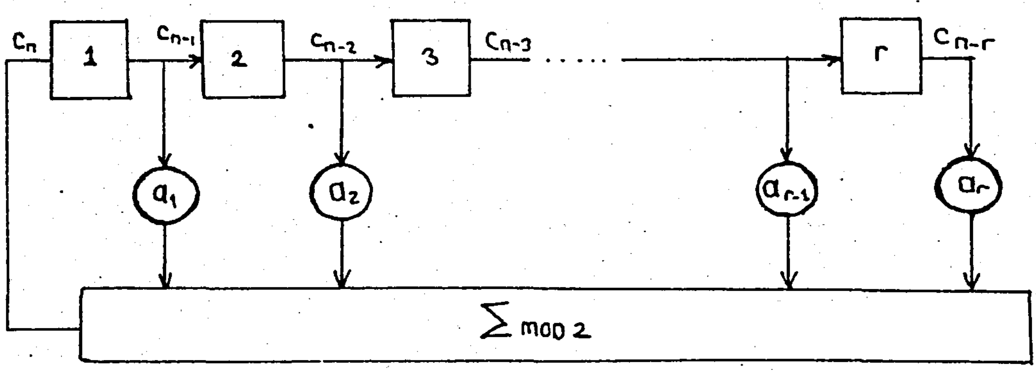


Fig.3.4 Simple shift register (SSRG)

ii. Permitting practical implementation, including convenint code changes.

Denying future information is probably the foremost criteria, if future sequence values are not totaly uncertain, the unintended party should not be able to reduce the intended system processing gain, at least in antijam systems. This means that the PN code period must exceed the time between code changes, and the code must not be "crackable" in the encryption sense.

A fundamental issue in spread spectrum is how the spreading technique affords protection against interfering signals with finite power. The underlying principle is that of distributing a relatively low dimensional data signal in a high dimensional environment so that a jammer with a fixed amount of total power is obliged to either spread that fixed power over all the coördinates, thereby inducing just a little interference in each coördinate, or else place all of the power into a small subspace, leaving the remainder of the space interference free.

b) PN Sequence generators

A purely random sequence can be used to spread the signal spectrum. Unfortunately, in order to despread the signal, the receiver needs a replica of the transmitted sequence. In practice, therefore, we generate PN sequences so that the following properties are satisfied.

i. They are easy to generate

ii. They have long periods.

iii. They have randomness property

iv. They are difficult to reconstruct from a short segment.

Linear feedback shift registers (LFSR) sequences posses properties i and ii most of property iii, but not property iv. One canonical form of a binary LFSR known as a simple shift register generator (SSRG) is shown in Fig. 3.4 The shift register consists of binary storage elements (boxes) which transfer their contents to the right after each clock pulse. The contents of the register are linearly combined with the binary coefficients a_k and are feedback to the first stage. The binary code sequence C_n then clearly satisfies the recursion

$$C_n = \sum_{k=1}^{r} a_k C_{n-k} \qquad a_r = 1 \tag{3.8}$$

The periodic sycle of the states depends on the initial state and on the coefficients a_k. For example, the four-stage LFSR generator shown in Fig.3.5. has four possible cycles as shown. The all zeros is always a cycle for any LFSR.

For spread spectrum we are looking for maximal length cycles, that is, cycles of period (2^r-1) (all binary r-tubles

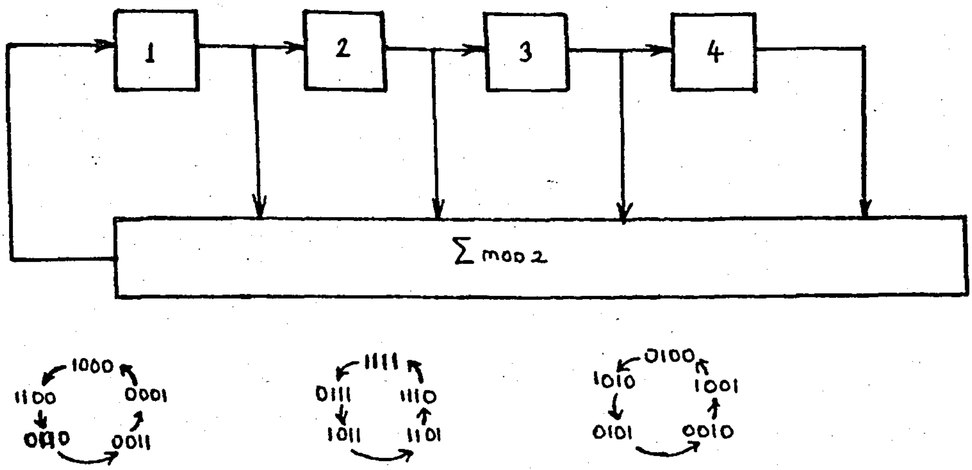


Fig.3.5 Four stage linear feedback shift register (LFSR) and its state cycles

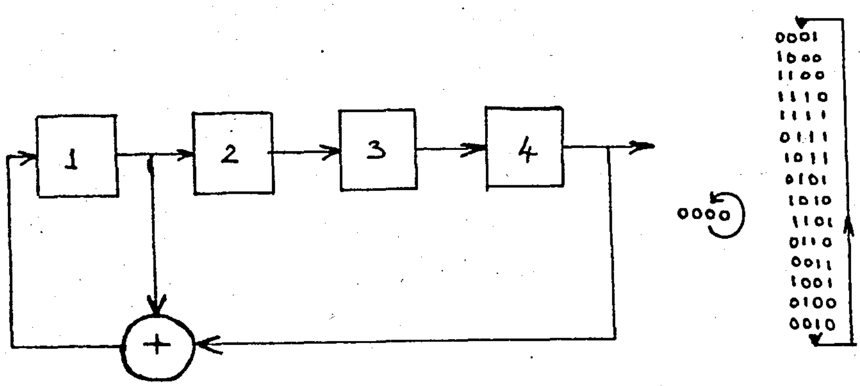


Fig.3.6 Four stage maximal length LFSR and its state cycles

except all-zeros). An example is shown for a four stage register in Fig 3.6. The sequence output is 100011110101100 (period $2^4-1) = 15$) if the initial contents of the register are 1000. It is always possible to choose the feedback coefficients so as to achieve maximal length.

Maximal codes are, by definition, the longest codes that can be generated by a given shift register or a delay element of a given length. Properties held by all maximal code sequences are briefly these :

i. There is a balance of zeros and ones. The number of ones in a sequence equals the number of zeros with in one priod. For a 1023 bit rate there are 512 ones and 512 zeros.

ii. The statistical distribution of ones and zeors is well defined and always the same. Relative positions of their runs vary from code sequence to code sequence but the number of eachrun length doesnot.

iii. Autocorrelation of a maximal linear code is such that for all values of phase shift the correlation value is (-1), in which correlation varies linearly from the (-1) value to (2^r-1).

iv. A Modulo-2 addition of a maximal linear code with a phase shifted replica of itself results in another replica with a phase shift different from either of the originals.

v. Every possible state, or r-tuple, of a given r-stage generator exists at some time during the generation of a complete code cycle. Each state exists for one and only one clock bit. The exception is that all-zeros state does not normally occur and cannot be allowed to occur.

3.2. DS WITH INTERFERENCE

Interference is introduced in the transmission of the information-bearing signal through the channel. The characteristics of the interference depend to a large extent on its origin. It may be categorized as being either broadband or narrowband relative to the bandwidth of the information-bearing signal, and either continuous in time or pulsed in time. For example a Jamming signal may consist of one or more sinusoids in the bandwidth used to transmit the information. The frequencies of the sinusoids may remain fixed or they may change with time according to some rule. As a second example, the interference generated in CDMA by other users of the channel may be either boradband or narrowband depending on the type of spread spectrum signal that is employed to achieve multiple access. If it is broadband, it may be characterized as an equivalent additive white gaussion noise.

Probably the single most important application of spread spectrum techniques is that of resistance to intentional interference or jamming. The two most common types of jamming signals analyzed are single frequency sine waves (tones) and broad-band noise. The simplest case to analyze is that of broad-band noise jamming. If the jamming signal is modelled as a zero-mean wide sense stationary gaussian noise process with a flat power spectral density over the bandwidth of interest, then for a given fixed power (P_J) available to the Jamming signal, the power spectral density of the jamming signal must be reduced as the bandwidth that the jammer occupies is increased.

For a DS system, if we assume that the jamming signal occupies the total RF bandwidth, typically taken to be twice the chip rate, then the despread jammer will occupy an even greater bandwidth and will appear to the final integrate-and-dump detection filter as approximately a white noise process.

A fundamental issue in spread spectrum is how this technique affords protection against interference signals with finite power. The underlying principle is that of distributing a relatively low dimensional data signal in a high dimensional environment so that a jammer with a fixed amount of total power is obliged to either spread that fixed power over all the coodinates, (place all of the power into a small subspace), leaving the remainder of the space interference free.

A brief discussion of a classical problem of signal detection in noise should clarify the emphasis on finite interference power. The standart problem of digital transmission in the presence of thermal noise is one where both transmitter and receiver know the set of M signalling waveforms $\{S_i(t),\ 0 < t < T,\ 1 < i < M\}$. The transmitter selects one of the waveforms every T seconds to provide a data rate of $\log_2 M/T$ bits/s. If for example $s_j(t)$ is sent, the receiver observes

$$r(t) = S_j(t) + n_w(t) \tag{3.9}$$

where $n_w(t)$ is additive, white gaussion noise.

Since the desired signal can be "collapsed" by correlating the signal at the receiver with the known code, the desired signal is protected against a jammer in the sense that it has an effective power advantage relative to the jammer. This power advantage is often proportional to the ratio of the dimensionality of the space of code sequences to that of the data signal. It is necessary, of course, to hide the pattern by which the data are spread. This is usually done with a PN sequence which has desired randomness properties and which is available to the cooperating transmitter and receiver, but denied to other undesirable users of the common spectrum.

3.2.1. INTERFERENCE REJECTION

Spread spectrum, Direct sequence, or PN modulation is employed in digital communication system to reduce the effects of interference due to other users and intentional jamming. When the interference is narrow-band the crosscorrelation of the received signal with the replica of the PN code sequence reduces the level of the interference by spreading it across the frequency band occupied by the PN signal. Thus the interference is rendered equivalent to a lower level noise with a relatively flat spectrum. Simultaneously, the cross-correlation operation collapses the desired signal to the bandwidth occupied by the information signal prior to spreading.

The interference immunity of a DS system corrupted by narrow-band interference can be further improved by filtering the signal prior to cross correlation, where the objective is to reduce the level of the interference at the expense of introducing some distortion on the desired signal. This filtering can be accomplished by exploiting the wideband spectral characteristics of the desired PN signal and the narrow-band characteristic of the interference is easily recognized and estimated (11).

3.2.2. ALGORITHMS FOR ESTIMATING AND SUPPRESSION OF NARROWBAND INTERFERENCE

The algorithms may be classified into two general categories. The algorithms in the first category employ the Fast Fourier Transform (FFT) algorithm for performing a spectral analysis from which an appropriate transversal filter is specified. These algorithms are termed nonparametric, since no prior knowledge of the characteristics of the interference is assumed in forming the estimate. The algorithms in the second category are based on linear prediction and may be termed parametric. That's, the interference is modelled as having been generated by passing white noise through an all pole filter.

a) Nonparametric spectral estimation

The basis for this method is that the power density spectrum of the PN sequence is relatively flat while the spectrum of the narrowband interference is highly peaked. The first step in this method is to estimate the power spectral density of the received signal. The spectral estimate can be obtained by any one of the well-known spectral analysis techniques, (i.e. Welch method) (10).

Once the power spectral density of the received signal is estimated, the interference suppression filter can be designed. A transversal filter is an appropriate filter structure for this application, since we desire to use a filter that contains notches in the frequency range occupied by the interference.

In effect, the filter characteristic attemptes to approximate an inverse filter to the power spectral density. That is, the interference suppression filter attempts to suppress the spectrum of the incoming signal. Thus, the filter will have a large attenuation in the frequency range occupied by the interference and a relatively small attenuation elsewhere. A relatively simple method for designing the transversal filter is to select its DFT to be the reciprocal of the square root of the power spectral density at equally spaced frequencies.

b) Parametric spectral estimation or linear prediction.

In constrast to the nonparametric spectral analysis method, the following method for estimating the narrow-band interference is based on modeling the interference as white noise passed through an all-pole filter that is, instead of using the received signal to estimate the spectrum directly, the signal is used to estimate the pole positions. This estimation is accomplished by means of linear prediction. An estimate of the power spectral density is easily obtained from the all-pole model. The interference suppression filter is simply a transversal filter having zero positions that coincide with the estimated pole positions. Thus, the spectrum of the signal at the output of the transversal filter is rendered white.

In order to develop the mathematical formulation for the all-pole model, we begin with the received signal,

$$r(t) = s(t) + i(t) + n(t) \tag{3.10}$$

where $s(t)$ is the information bearing signal, $i(t)$ denotes the narrow-band interference, and $n(t)$ is assumed to be a sample function of a white gaussian noise process. For convenience, we assume that $r(t)$ is sampled at the chip rate of the PN sequence. Thus r(k) can be expressed as

$$r(k) = s(k) + i(k) + n(k) \qquad k : 1,2, \ldots \tag{3.11}$$

We assume that s(k), i(k), and n(k) are mutually uncorrelated. An estimate of the interference i(t) is formed from r(k). Assume for the moment that the statistics of i(t) are known and are stationary.

Then we can predict i(k) from r(k-1), r(k-2), ..., r(k-m). That is

$$\hat{i}(k) = \sum_{\ell=1}^{m} a_\ell r(k-\ell) \tag{3.12}$$

where $\{a_\ell\}$ are the coefficients of the linear predictor. The coefficients are determined by minimizing the mean square error between r(k) and i(k), which is defined

$$\begin{aligned} \varepsilon(m) &= E\left(r(k) - \hat{i}(k)\right)^2 \\ &= E\left(r(k) - \sum_{\ell=1}^{m} a_\ell r(k-\ell)\right)^2 \end{aligned} \tag{3.13}$$

minimization of $\varepsilon(m)$ with respect to the predictor coefficients $\{a_\ell\}$ can be easily accomplished by involving the orthogonality principle in mean square estimation. This leads to the set of linear equations (10).

$$\sum_{\ell=1}^{m} a_{\ell}\rho(k-\ell) = \rho(k) \qquad k = 1,2, \ldots m \tag{3.14}$$

where

$$(k) = E\left(r(m)r(k\ m)\right) \tag{3.15}$$

is the autocorrelation function of the received signal $r(k)$. The above equations $\rho(k)$ are usually called the Yule-walker equations. They can be written in matrix form as

$$R_m a_m = b_m$$

where R_m is the mxm autocorrelation matrix, a_m is the vector of filter coefficients and b_m is a vector of outocorrelation coefficients $\rho(k)$. The matrix R_m is a Toeplitz matrix, which is efficiently inverted by use of the levinson Durbin Algorithms

The solution of the Yule-Walker equations for the coefficients a_{mk} of the prediction filter requires knowledge of the outocorrelation function $\rho(k)$. In practice, the outo-correlation function of $i(k)$ and, hence, $r(k)$ is unknown and it may also be slowly varying intime. Consequently, one must consider methods for obtaining the predictor coefficients directly from the received signal $r(k)$. This may be occomplished in a number of ways. In this investigation, three different methods can be mainly considered

i. Direct application of the levinson Algorithm.

The first method is simply based on the direct estimation of $\rho(k)$ from the block of N samples. The estimate of $\hat{\rho}(k)$ is

$$\hat{\rho}(k) = \sum_{n=0}^{N-k} r(n)r(n+k) \qquad k = 0,1,\ldots,m \quad (3.16)$$

the estimate $\hat{\rho}(k)$ may then be substituted in the Yule-Walker equations in place of $\rho(k)$ and the levinson-Durbin algorithm can be used to solve the equations efficiently. The levinson Durbin algorithm, described in Appendix A, is used in our computer programs for obtaining the filter coefficients.

ii. Burg Algorithm

iii. Least Squares Algorithm

The good estimation of the predictor coefficients can be obtained with as few samples as twice the number of predictor coefficients. The levinson algorithm appears to require a few more samples. We expect the least squares algorithm to be comparable to the Burg algorithm in its performance and its sample size requirements. We did not perform a threugh study of the sample size requirements for obtaining good estimates, however.

Once the prediction coefficients are determined, the estimate $\hat{i}(k)$ of the inteference, is substructed from r(k) and the difference signal is processed further in order to extract the digital information. Thus the equivalent transversal filter for suppressing the interference is described by the transfer function

$$A_m(Z) = 1 - \sum_{k=1}^{m} a_m Z^{-k} \tag{3.17}$$

where Z^{-1} denotes a unit of delay. The corresponding all-pole model for the interference signal is

$$\frac{1}{A_m(Z)}$$

3.2.3. PERFORMANCE CRITERIA ON INTERFERENCE SUPPRESSION

A model of the spread spectrum communication system is shown in Fig. 3.7. The information signal U(t) is assumed to be binary PSK with symbol values U = ±1. The SS modulation is accomplished by PN signaling with signaling element or "chips" given by P = ±1. The interference is modelled as a sum of fixed amplitude and fixed frequency tones with random phases.

The received wideband signal is then sampled once per chip and can be represented as (10).

$$r(k) = s(k) + i(k) + n(k)$$

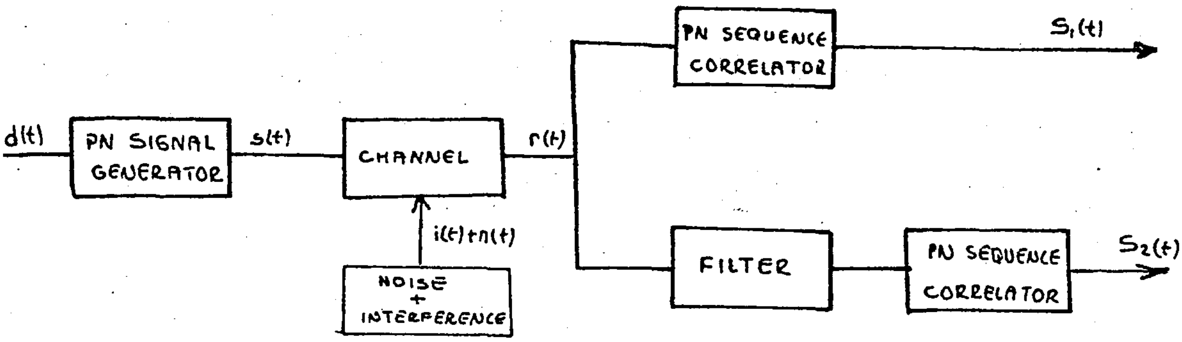


Fig. 3.7 Spread spectrum communication system model

Register	Sequence Length
7	127
8	255
9	511
10	1023
11	2047
12	4095
13	8191
14	131071
15	524287
16	8388607

Table.3.1.Code sequence lengths for various registers, 1 Mbps Rate

or

$$r(k) = UP_k + \sum_{m=1}^{m} C_m \cos(2\Pi f_m k\Delta t + \Phi_m) + n_k \tag{3.18}$$

where U is the information signal to be transmitted, P_k is the modulated PN sequence, C_m, f_m and Φ_m are the amplitude, frequency and phase of the narrow-band interference, Δt is the sampling interval and n_k is the zero mean white Gaussian noise with a variance G^2. In narrow-band gaussian interference,

$$\begin{aligned} C_m &= (2P(f_m)\Delta f)^{1/2} \\ f_m &= m\Delta f \end{aligned} \tag{3.19}$$

where $P(f_m)$ is the power spectrum at frequency f_m and Δf is the frequency speration between two neighboring frequencies. In Fig. 3.7. the received signal is processed in a conventional SS receiver to yield the output $S_1(t)$ and in an SS receiver using digital filter prior to correlation to produce the output $S_2(t)$.

A measure of the performance improvement due to filtering can then be obtained by comparing the Signal-to-Noise ratios of the outputs $S_2(t)$ and $S_1(t)$. The signal-to-noise ratio (SNR_i) of the output S_i (i:1,2) shown in Fig. 3.7. can be defined by

$$SNR_i = \frac{E^2\{S_i\}}{V_{ar}\{S_i\}} \qquad i = 1,2 \tag{3.20}$$

It can be shown that (W), the ratio of SNR_2 to SNR_1 gives us the performance criteria of the system.

$$W = \frac{SNR_2}{SNR_1} \tag{3.21}$$

a) THEORETICAL SNR PERFORMANCE IN A CONIENTIONAL SS SYSTEM (SNR_1 WITHOUT FILTER)

The sampled wideband received signal can be represented by

$$R_k = UP_k + q_k$$

where

$$q_k = i_k + n_k \tag{3.22}$$

and

$$i_k = \sum_{m=1}^{m} C_m \operatorname{Cos}(2\Pi f_m k\Delta t + \Phi_m)$$

If the signal is correlated with the known PN code, the correlator output, S_1, can be expressed

$$S_1 = \sum_{k=1}^{L} R_k P_k = UL + \sum_{k=1}^{L} i_k P_k + n_k P_k \qquad (3.23)$$

where L is the number of chips per information symbol and the PN code chips $P_k = \pm 1$. In order to simplify the analysis, an ensemble of independent PN systems is assumed where

$$E(P_k P_L) = \delta_{k-\ell} = \begin{cases} 1 \ \ldots \ k=\ell \\ 0 \ \ldots \ \text{otherwise} \end{cases} \qquad (3.24)$$

The SNR of the correlator output can now be computed from the mean nad variance of S_1. The mean of S_1 is given by,

$$E(S_1) = UL \qquad (3.25)$$

Since $E(q_k) = 0$ as a result of the uniformly distributed random phase Φ_m and the zero mean noise n_k. The variance of S_1 is given by

$$V(S_1) = E(S_1^2) - L^2$$

where $U = \pm 1$.

$$V(S_1) = \sum_{k=1}^{L} E(q_k^2) \qquad (3.26)$$

since i_k and n_k are independent and zero mean

$$V(S_1) = L(G_i^2 + G^2) \tag{3.27}$$

where G^2 is the noise power and G_i^2 is the interference power.

$$G_i^2 = E(i_k^2)$$

$$= \frac{1}{2} \sum_{m=1}^{m} C_m^2 \tag{3.28}$$

then

$$V(S_1) = L \sum_{m=1}^{m} \frac{1}{2} C_m^2 + LG^2 \tag{3.29}$$

and

$$SNR_1 = \frac{E^2(S_1)}{V(S_1)} = \frac{L}{\sum_{m=1}^{m} \frac{1}{2} C_m^2 + G^2} \tag{3.30}$$

$$SNR_1 = \frac{L}{\sum_{m=1}^{m} \frac{1}{2} C_m^2 + G^2}$$

b) THEORETICAL SNR PERFORMANCE IN SS SYSTEM USING WHITENING FILTER (SNR_2 WITH FILTER)

An estimate of $\{i(t)\}$ interference is formed from $r(k)$

$$\hat{\imath}(k) = \sum_{\ell=1}^{m} a_\ell r(k-\ell) \tag{3.31}$$

where (a_ℓ) are the coefficients of the linear predictor. After the prediction coefficients (a_ℓ) are estimated from the whitening algorithms, an error signal is formed as

$$E_k = r(k) - \hat{\imath}(k)$$

$$= r(k) - \sum_{\ell=1}^{m} a_\ell r(k-\ell) \tag{3.32}$$

$$= \sum_{\ell=0}^{m} a_\ell r(k-\ell)$$

The error signal is correlated with the PN code to produce the output signal S_2.

$$S_2 = \sum_{k=1}^{\ell} \varepsilon_k p_k$$

$$= \sum_{k=1}^{L} \sum_{=0}^{M} a_\ell p_k (UP_{k-\ell} + q_{k-\ell}) \tag{3.33}$$

where

$$r(k-\ell) = UP(k-\ell) + q(k-\ell)$$

the mean of S_2 is obtained as in SNR_1 calculations

$$E(S_2) = UL \tag{3.34}$$

Since $a_o = 1$ and $E(q-\ell) = 0$ the variance of S_2 is

$$V(S_2) = E(S_2^2) - L^2$$

$$= I_2 + I_1 - L^2 \tag{3.35}$$

where

$$I_1 = \sum_{k=1}^{L} \sum_{\ell=0}^{M} \sum_{i=1}^{L} \sum_{j=0}^{M} a_\ell a_j E(P_k P_i) E(q_{k-} \; q_{i-j})$$

and (3.36)

$$I_2 = \sum_{k=1}^{L} \sum_{\ell=0}^{M} \sum_{i=1}^{L} \sum_{j=0}^{M} a_\ell a_j E(P_k P_i P_{k-1} P_{i-j})$$

Evaluation of I_1 requires the identity

$$E\, q_{k-1} q_{k-i} = \frac{1}{2} \sum_{m=1}^{m} C_m^2 \operatorname{Cos} 2\Pi f_m (j-\ell)\Delta t + G^2 \delta_{j-\ell}$$

and

$$E(P_k P_\ell) = \delta_{k-\ell} = \begin{cases} 1, & K=1 \\ 0, & \text{otherwise} \end{cases} \tag{3.37}$$

using these two identities in equation I_1,

$$I_1 = \frac{L}{2} \sum_{\ell=0}^{m} \sum_{j=0}^{m} a_\ell a_j \sum_{m=1}^{m} C_m^2 \operatorname{Cos} 2\Pi f_m (i-\ell)\Delta t + LG^2 \sum_{\ell=1}^{m} a_\ell^2$$

(3.38)

The power spectrum of the optimum filter is given by

$$P(f_m) = \frac{P_N \Delta_t}{\left|\sum_{\ell=0}^{m} a_\ell e^{-j2\Pi f_m \ell \Delta t}\right|^2}$$

$$= \frac{P_N \Delta t}{\sum_{\ell=0}^{m} \sum_{\ell=0}^{m} a_\ell a_i \operatorname{Cos}(2\Pi f_m(\ell-i)\Delta t)} \tag{3.39}$$

where $p(f_m) = \frac{1}{2} \frac{C_m^2}{\Delta f}$

Substituting these two equations of $P(f_m)$ in equation I_1

$$I_1 = LP_N \Delta t \Delta f M + LG^2 \sum_{\ell=1}^{m} a_\ell^2 \tag{3.40}$$

Evaluating I_2 requires the identity

$$E\, P_k P{:}P_{k-}\, P_{i-j} = \begin{cases} 1, & \ell = j = 0 \\ \delta_{\ell-j}, & \ell >, j > 0, i = k \\ 0, & \text{otherwise} \end{cases} \tag{3.41}$$

using this identity in equation I_2, given

$$I_2 = L^2 + L \sum_{\ell=1}^{m} a_\ell^2 \tag{3.42}$$

combining I_2, and I_1 in equation $V(S_2) = I_2 + I_1 - L^2$ then

$$V(S_2) = L(1+G^2) \sum_{n=1}^{m} a_n^2 + LP_N \Delta t \Delta f \, M \tag{3.43}$$

$$SNR_2 = \frac{E^2(S_2)}{V(S_2)}$$

$$= \frac{U^2 L^2}{L(1+G^2) \sum_{\ell=1}^{m} a_\ell^2 + LPN\Delta t\Delta fM}$$

$$SNR_2 = \frac{L}{(1+G^2) \sum_{\ell=1}^{m} a_\ell^2 + MP_N \Delta t \Delta f} \tag{3.44}$$

The ratio of SNR_2 to SNR_1 (Performance) is finally given by

$$W = \frac{\frac{1}{2} \sum_{m=1}^{M} C_m^2 + G^2}{(1+G^2) \sum_{\ell=1}^{m} a_\ell^2 + MP_N \Delta f \Delta t} \tag{3.45}$$

where P_N is the average Power of the error signal and this can be computed from these equations.

$$P_N = P_{N-1}(1-a_{N,N}^2) \tag{3.46}$$

$$P_o = \frac{1}{M} \sum_{k=0}^{m-1} r_k^2 \tag{3.46}$$

IV. FREQUENCY HOPPING SPREAD SPECTRUM SYSTEMS

4.1. AN INTRODUCTION TO FH SIGNALS

The simplest form of FH results from frequency shift keying (FSK) a carrier with the binary message and then hopping this carrier over a wide range of frequencies in accordance with some pseudo-random frequency pattern. The transmitter jumps from frequency to frequency within some predetermined set. The order of frequency usage is determined by a code sequence. This same pseudo-random hopping pattern must be generated at the receiver in synchronism with the received signal in order to despread it to a normal binary FSK signal.

Frequency Hopping modulation is more accurately termed "Multiple frequency, code selected, frequency shift keying" (2) simple FSK most often uses only two frequencies ; for example f_1 is sent to signify a "MARK", f_2 is to signify a "SPACE". Frequency hoppers, on the other hand, often have thousands of frequencies available. The number of frequencies and the rate of hopping from frequency to frequency in any frequency hopper is governed by the requirements placed on it for a particular use.

The wide-band frequency spectrum desired is generated in a different manner in a FH system. It does just what its name implies. That is, it "hops" from frequency to frequency over a wide-band. The specific order in which frequencies are occupied is a function of a code sequence, and the rate of hopping from one frequency to another is a function of the rate at which information is to be sent.

4.1.1. CHARACTERISTICS OF FH SYSTEM

A FH system or "Frequency Hopper" consists basically of a code generator and a frequency synthesizer capable of responding to the coded outputs from the code generator. A great deal of effort has been expanded in developing rapid-response frequency synthesizers for spread spectrum systems.

Ideally, the instantaneous frequency hopper output is a single frequency. Practically, however, the system user must be satisfied with an output spectrum which is a composite of the desired frequencies, sidebands generated by hopping and spurious frequencies generated as by-products.

Over a period of time the ideal frequency hopping spectrum would be perfectly rectangular, with transmissions distributed evenly in every available frequency channel. The transmitter should also be designed to transmit to a degree as close as practical, the same amount of power in every channel.

As in a DS system, any signal that is not a replica of the local reference is spread by multiplication with the local reference. Bandwidth of an undesired signal after multiplication with the local reference is again equal to the covariance of the two signals, for example, a continuous wave (CW) signal appearing at the FH receiver's input would be identical to the local reference when translated to the IF frequency. A signal with the same bandwidth as the local reference would have twice the reference bandwidth at the IF. The IF following the correlator, then can reject all of the undesired signal power that lies outside its bandwidth. Because this IF bandwidth is only a fraction of the bandwidth of the local reference, we can see that almost all the undesired signals power is rejected, where as a desired signal in enhanced by being correlated with the local reference.

In the preceding section on DS systems we saw that a DS system's operation is identical from the stand point of undesired signal rejection and remapping the desired signal. From this general viewpoint DS and FH systems are identical ; they are different, however, in the details of their operation.

a. Model of FH Spread Spectrum System

In a FH spread spectrum communications system the available channel bandwidth is subdivided into a large number of contiguous frequency slots. In any signaling interval, the transmitted signal occupies one or more of the available frequency slots. The selection of the frequency slots in each signaling interval is made pseudorandomly according to the output from a PN generator. Figure 4.1. illustrates a particular frequency-hopped pattern in the time-frequency plane

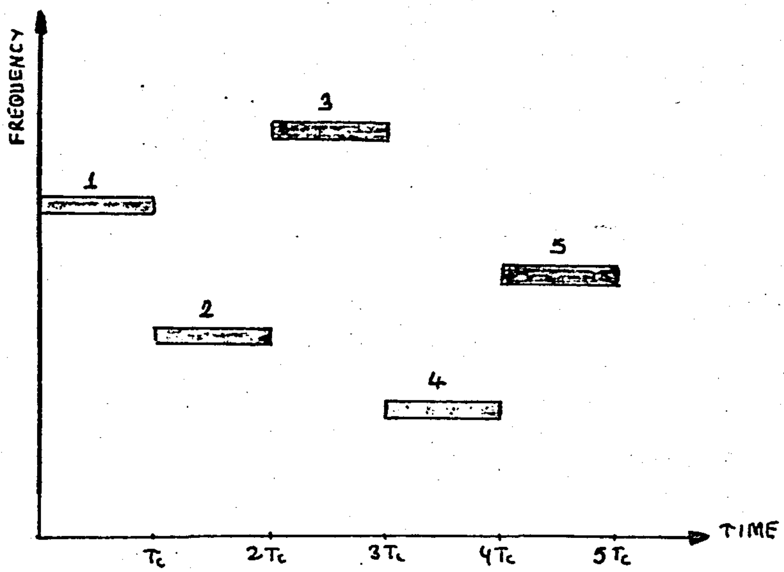


Fig .4.1 An example of a FH pattern

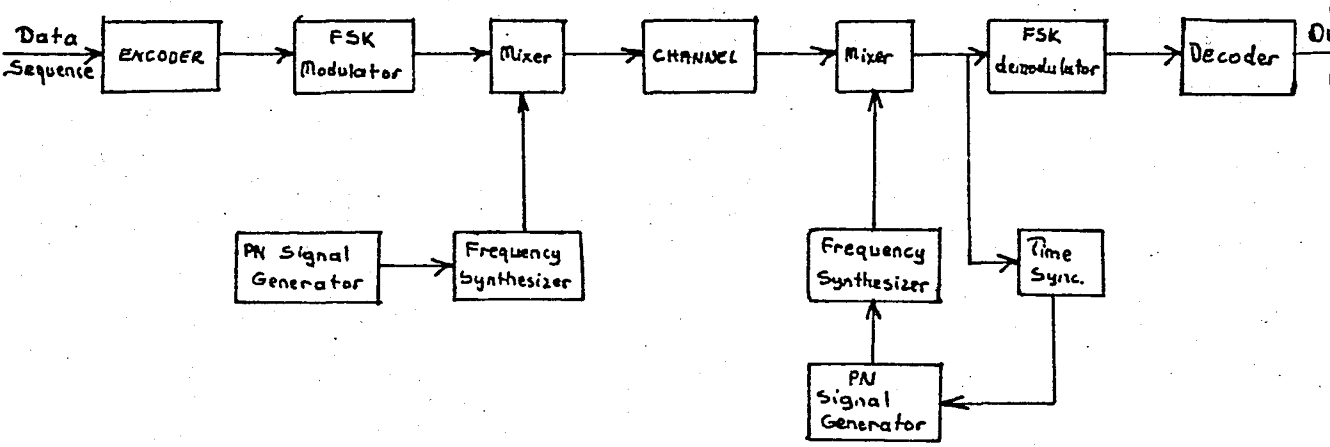


Fig .4.2 Model of an FH spread spectrum system

A block diagram of the transmitter and receiver for a FH spread spectrum system is shown in Figure 4.2. The modulation is usually binary FSK, for example, if binary FSK is employed, the modulator selects one of two frequencies corresponding to the transmission of either a 1 or a -1 The resulting FSK signal is translated in frequency by an amount that is determined by the output sequence from the PN generator which, in turn, is used to select a frequency that is synthesized by the frequency synthesizer. This frequency is mixed with the output of the modulator and the resultant frequency translated signal is transmitted over the channel.

At the receiver, we have an identical PN generator, synchronized with the received signal, which is used to control the output of the frequency synthesizer. Thus the PN frequency translation introduced at the transmitter is removed at the receiver by mixing the synthesizer output with the received signal (12).

The resultant signal is demodulated by means of an FSK demodulator. A signal for maintaining synchronism of the PN generator with the frequency-translated received signal is usually extracted from the received signal.

Although PSK modulation gives better performance than FSK in an Additive White Gaussion Noise (AWGN) channel, it is difficult to maintain phase coherence in the synthesis of the frequencies used in the hopping pattern and, also, in the propagation of the signal over the channel as the signal is hopped from one frequency to another over a wide bandwidth. Consequently FSK modulation with noncoherent detection is usually employed with FH spread spectrum signals.

FH spread spectrum signals are used primarily in digital communications that require antijamming protection and in Code-division multiple access (CDMA) where many users share a common bandwidth. In most cases, a FH signal is preferred over a PN spread spectrum signal because of the stringent synchronization requirements inherent in PN SS signals. In a PN system, timing and synchronization must be established to within a fraction of the chip interval $T_c = \frac{1}{W}$. On the other hand, in an FH system, the chip interval is the time spent in transmitting a signal in a particular frequency slot of bandwidth $B \ll W$. But this interval is approximately $1/B$, which is much larger than $1/W$. Hence the timing requirements in a FH system are not as stringent as in a PN system.

b. Process Gain of FH System.

The kind of spread spectrum technique, illustrated by the use of a PN sequence, produces an instantaneous spreading of the transmitted bandwidth. It may be that the resultant processing gain is still not enough to overcome the effects of some kinds of jammemrs. It would be possible to use even narrower chip widths (and thus wider bandwidth and more chips per bit), but there is a practical limit to this imposed by the capabilities of the physical devices used to generate spread spectrum signals.

Process gain for the FH system is the same as that for the DS system, that is (13)

$$\text{Process Gain} = G_p = \frac{BW}{R_{data}}$$

An alternative method of forcing the jammer signal to cover a wider spectrum is to randomly hop the transmitted frequency of the data (information) symbol on a symbol-by-symbol basis. This spreads the spectrum sequentially rather than instantaneously. Assuming that the jammer will decide to spread its energy over the entire frequency-hopped spectrum, the potentially available processing gain is then gives by (1)

Processing Gain = 10(log Hopped BW/Data Rate)
(Logarithmic)

However the jammer may decide to concentrate on just a few of the hopped frequencies, assuming that it is more effective to attempt to cause errors in only some of the information bits. In this case, the effective processing gain realized by the receiver would be less than that given by this equation. If the information bit or symbol is also spread by means of PN coding, then the use of both FH and PN codes provides a processing gain given by

Processing Gain = 10 log(chip Rate/data Rate)
(Logarithmic)

$$+ 10 \log(BW/R_{data})$$

A system that uses both techniques is referred to as a hybrid. This spread spectrum systems are useful when a single technique such as PN coding doesnot provide an adaquate SNR margin.

In our FH spread spectrum system which is used in the computer programs, there are 32 code bits per one information symbol and

R_{data} = information Rate = 10 kbits/s

BW = Hopped Bandwidth = 2x320.000 = 0.64 MHz

then

$$\text{Processing Gain} = \frac{640.000}{10.000} = 64$$

G_p = 10 log 64 = 18 dB.

Jamming margin is determined by a system's process gain, acceptable output SNR and implementation losses. This margin, sometimes called anti-jamming (AJ) margin, is the amount of interference that a receiver can withstand while operating and producing an acceptable output SNR. For our system which has 18 dB Process gain, 10 dB output SNR and 2 dB losses, then the jamming margin is

$$\text{Jamming Margin} = G_p - (\text{Limp.} + SNR_{out})$$

$$\text{Jamming Margin} = 18 - (2 + 10) = 6 \text{ dB.}$$

4.1.2 FH RATE AND NUMBER OF FREQUENCIES

The minumum frequency switching rate usable in a frequency hopping system is determined by a number of parameters :

i) The type of information being sent and its rate.

ii) The amount of redundancy used, if any.

iii) Distance to the nearest potential interferer.

Information in a FH system may be transmitted in a way available to other systems. Usually, however, some form of digital signal is used, whether the information is a digitized analog signal or data. Assume for the present that some digital rate is prescribed and that frequency hopping has been chosen as the transmission medium. How, then, is the FH or chip rate chosen?

A FH system must have a large number of frequencies usable on demand. The number required is dependent on system error rate ; for instance, a library of 1000 frequencies could provide good operation when interference or other noise is evenly distributed at every available frequency. For equal distribution of noise in every channel the noise power required to block communications would approach 1000 times desired signal power (in other words, jamming margin would be 30 dB). Unless some form of redundancy that allows for bit decisions based on more than one frequency is used, however, a single narrowband interferer would cause an error rate of 1×10^{-3}, which is generally unacceptable for digital data. For a simple FH system without any form of transmitted data redundancy the expected error rate is just J/N, where J equals the number of interferers with power greater than or equal to signal power and N equals the number of frequencies available to the system.

The practicality of increasing redundancy to improve bit error rate depends on system parameters. It is obvious that the more chips sent for a bit, the lower the bit error rate. The hopping rate required and the Bandwidth increase in direct proportion. If either the bandwidth allocated or the frequency capability of the frequency synthesizer is limited, then some trade off must be made between sending a larger number chips per bit and reducing the number of frequencies available.

As an example, assume that a data source at 1 kbps is to be transmitted and that a 10 MHz band-width is allowable. 1 kb data requires that frequency hopping be at least 1 khps, so that the main lobe of the dehopped carrier spectrum is 2 kHZ wide. Therefore, if no overlap is allowed, 10 MHz/2 kHZ = 5000 frequency hopping channels are available.

It has been assumed for the FH systems that contiguous frequency spacing is used ; that is, the dehopped signal appearing at the receiver is not allowed to overlap from one channel into another. This is not strictly a true picture in many systems, depending on the receiver used, transmit spacing can be such that significant overlap occurs. This overlap greatly reduces the bandwidth required for the transmitted spread spectrum signal. Fig 4.3. illustrates overlapping channels and the bandwidth savings. Between figures 4.3.a and 4.3.b there is a doubling of channels in the same bandwidth. Overlapping is such that the center of one channel falls at a null for the adjacent channels.

One other significant consideration regarding chip rate is the effect of signals which are received at the same frequency as the desired signal but are different in phase. These signals are due to multipath or, evenworse, deliberate interference. In most cases the multipath signal arriving at a receiver is much smaller than the desired and is therefore not of great consequence. The deliberate interferer who receives a signal from the FH transmitter, amplifies it, modulates with noise can be extremely effective with transmitter power

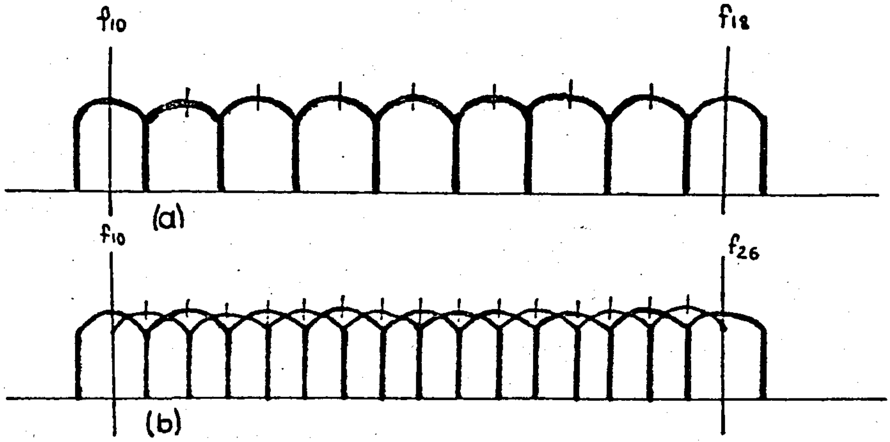


Fig.4.3 Illustration of contiguous versus overlapping channels showing gain in channels per unit bandwith (a)contiguous channel spacing (b) overlapping channel spacing

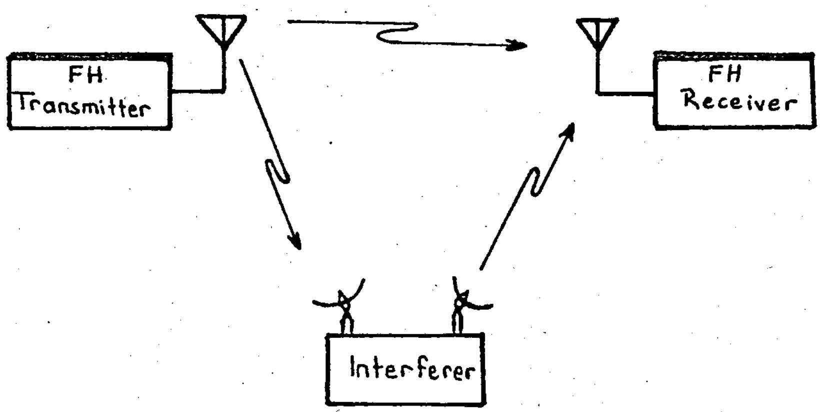


Fig.4.4 Repeater type interference diagram

similar to the friedly one. Against this kind of threat the FH might have no advantage. To combat the threat the frequency hopper must hop at a rate that allows it to skip to another frequency before the interferer can respond to the last one. The desired hopping rate is then greater than $(T_r - T_d)^{-1}$, where T_r is the total propagation time from the FH transmitter to the interferer and from there to the intended receiver. T_d is direct path delay, illustrated in Figure 4.4.

As we absorve Fig. 4.4. we see that the minimum required hopping rate is a function of distance to the interfering station and the angle of offset from the direct path. For fixed stations we can drive a minimum hop rate, but for mobile applications the conclusion must be "make the system hopping rate as fast as possible".

Because of its importance in frequency hopping, let us look at what has been said here ;

1) FH rate is a function of information rate

2) Information error rate can not be reduced below J/N unless some form of redundant transmission is used.

3) The number of frequency channels required is determined by the desired interference rejection capability and chip error tolerance.

4) The dehopped signal seen in a noncoherent frequency hop receiver is $\left[(\mathrm{Sin}x)/x\right]^2$ in shape and has main lobe bandwidth

equal to twice the chip rate.

5) Chip rate is bounded on the low side by multipath and repeater interference considerations. The best rule for variable geometry situations is to hop as fast as possible.

The fact that frequency hopping does not provide instantaneous coverage of the broad signal band leads to consideration of the rate at which the hops occur. Clearly the faster the hopping, the more nearly the frequency hop approximates true spectrum spreading. Two basic characterizations of frequency hopping are fast frequency hop and slow frequency hop (7). These are distinguished from one another by the amount of time spent at each discreate frequency before hopping to the next.

a. Slow Frequency Hop.

When slow frequency hop is employed, the carrier frequency remains constant for time preiods far in excess of the propagation time. Several miliseconds is a reasonable dwell time in this case. This usually allows many data bits to be transmitted at each frequency, and the resulting transmitter and receiver equipments simpler and less expensive than that for a faster frequency hop.

The disadvantage of slow frequency hop is that an enemy can implement smart jammers that could defeat the antijam protection in many instances. This can be accomplished by providing the jammer with a search receiver that scans the signal frequency band and locates the transmission ; then the jammer's power can be concentrated at the frequency where the signal is beign transmitted. If the jammer can adapt quickly enough, it may be able to follow the slow frequency hop.

On the otherhand, slow frequency hop can be used to interleave many frequency multiplexed channels within the same frequency hop band. In this application, each channel could be assigned a unique carrier frequency within the overall band. The frequency assignments would be changed from time to time so that each channel would hop among the frequencies in the band in coordination with the other channels, but in a manner that would appear random to the jammer.

As a consequence, slow frequency hop can be useful either against simple jammers or in conjuction with frequency division multiplexing of many signals in the wide bandwidth range of the hop. In many cases, these features, along with the lower relative cost compared with fast FH, may make this technique attractive.

b. Fast Frequency Hop.

As the name implies, fast FH involves very rapid retuning of the signal and very short dwell times at each frequency. Generally, a fast hop is applied to defeat the smart jammer's attempt to measure signal frequency and tune the interference to that portion of the band. To defeat this tactic, the signal must be hopped to a new frequency before the jammer can complete its measurement, retuning, and interference functions.

The required hopping rate is determined by considering time delays introduced by signal propagation to the receiver and jammer, and time delays involved in processing and tuning at the jammer. The FH rate is usually selected to be either equal to the symbol rate or faster than that rate. If there are multiple hops per symbol, we have a fast-hopped signal.

Fast frequency hopping is employed in antijam applications when it is necessary to prevent a type of jammer, called a follower jammer , from having sufficient time to intercept the frequency and retransmit it along with adjacent frequencies so as to create interfering signal components. However, there is a penalty incurred in subdividing a signal into several frequency-hopped elements because the energy from these separate elements is combined noncoherently.

4.2. FH SYSTEM WITH INTERFERENCE

Frequency-Hopping communication system is a class of spread spectrum systems which can employ extremely wide bandwidth much greater than that actually required for communication. In FH systems the carrier is switched to a new frequency occupying a new frequency cell of bandwidth, say BHz, which is a small fraction of the total spread spectrum bandwidth, say WHz, where $B \ll W$.

We consider a particular form of FH system where the carrier is phase modulated by a differantially encoded binary bit stream and each hopped carrier conveys 1 bit of information. This type of FH system is referred to as a pure FH or slow hopping system.

In the receiver the spread spectrum modulation is removed by correlating the received signal with the spread spectrum reference signal. The primary advantage provided by spread spectrum systems is the ability to reject narrowband interference due to jamming signals and other users of the band during the recovery of the desired signal. To enhance the performance of conventional spread spectrum signals, digital whitening techniques can be employed prior to correlation (14).

The received signal consists of the sum of the spread spectrum transmitted signals, interference and receiver thermal noise. This interference can be predicted and suppressed by use of digital whitening techniques as in DS systems.

4.2.1 RESISTANCE TO JAMMING

One of the important reasons for the application of spread spectrum techniques to military communication systems is its ability to resist the effects of intentional jamming (8).

When narrowband jamming signals exist in one or more of the frequencies used by the transmitted signal, all message data carried by those particular hops will in all likelihood be destroyed. The type of processing needed to restore this data depends upon how the message data rate compares with the hop rate. If there is exactly one message bit per hop, a common situation, then error correction coding may be employed at the transmitter. The receiver is then able to recover the bits that are destroyed by the jamming signal. In a fast hop system, for which there are several hops during each message bit, there may be sufficient information from the hops that are not interfered with to recover the data (message) without any forward error correction.

Alternatively, in a slow hop system, in which there are several message bits per hop, it may be possible to include burst error correction in the transmitted signal and, thus, recover the lost information.

When partialband jamming is present, the situation becomes much more difficult. If a substantial fraction of the hop frequencies are jammed, forward error correction may be inadequate and it may be necessary to resort to some form of retransmission. The processing that is required in this case calls for the receiver to detect the existence of errors and inform the transmitter. Ideally the transmitter should utilize some knowledge concerning the jamming frequencies to select a retransmission time that minimizes the number of jammed hops in that particular message segment.

4.2.2. DIGITAL WHITENING TECHNIQUES FOR FH COMMUNICATIONS IN THE PRESENCE OF INTERFERENCE

a) Narrow-band interference and wide-band noise

The system to be analyzed is shown in Fig. 4.5. The information signal U(t) is assumed to be binary FSK with symbol values U = ±1. The spread spectrum modulation is accomplished by PN signaling with signaling element given by P = ±1. In the FH system shown in Fig. 4.5. the carrier frequency is pseudorandomly hopped in every signaling interval. The interference is modelled as a sum of fixed amplitude and fixed frequency tones with random phases. The received wide-band signal is then sampled once per chip and can be represented as ;

$$R(t) = S(t) + I(t) + n(t)$$

where S(t) is the transmitted signal, I(t) is the narrow-band interference and n(t) is the gaussian noise. Interference can be represented as in the DS systems.

$$I(k) = \sum_{m=1}^{m} A_m \mathrm{Cos}(2\Pi f_m k\Delta t + \Phi_m)$$

where M and k are the sample number of narrow-band interference and PN signal. A_m, f_m and Φ_m are the amplitude, frequency and phase of the interference Δt is the sampling interval.

In Fig. 4.6(a). it is shown that the power spectrum of the FH signals without interference, and in Fig. 3.6(b). We can see tha effect of narrow-band interference on the FH signals spectrum. Adjusting the frequency (f_m) of this narrow-band interference we can change the place of jamming zone.

We can estimate the narrow-band interference with linear prediction algorithm and suppress the jamming with levinson-durbin algorithm as in the DS system. Related power spectrums and whitening techniques of FH system will be examined step by step in chapter V. using some computer programs.

Let us consider the FH spread spectrum signal in the presence of broad-band interference characterized statistically as Additive White Gaussian Noise (AWGN). The jammer in the ith slot will be taken to be

$$J(t) = j_{ci}(t)\text{Cos}W_{it} - j_{si}(t)\text{Sin}w_{it}$$

where w_i is the center frequency of the ith slot and $j_{ci}(t)$ and $j_{si}(t)$ are independent zero-mean Gaussian random processes with power equal to J and flat power spectrum. When we use this gaussian noise jamming in our computer programs, results become different from the narrow-band interference. Since the zero mean gaussian noise has the flat power spectral density like the transmitted FH signals. We can not see any jammed region (peaked band) on the power spectrum of the received FH signals. But examining the values of received signal we

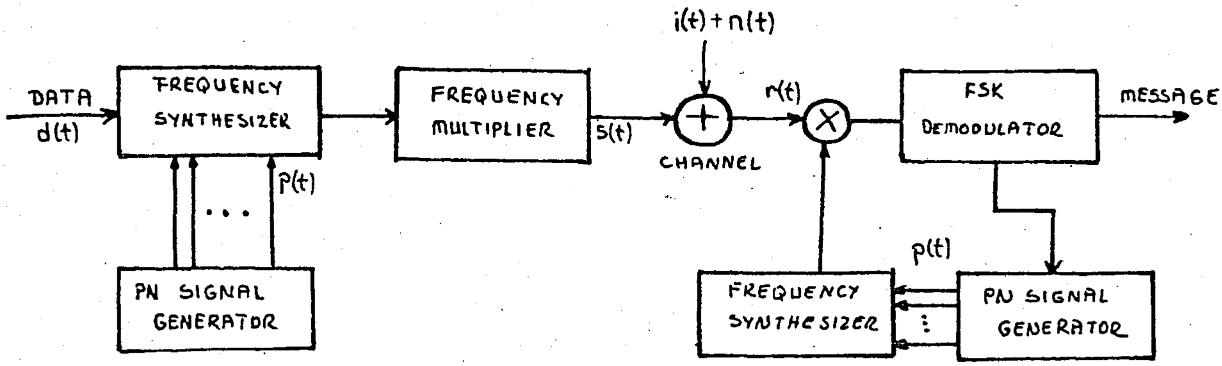


Fig.4.5 Frequency-Hopping communications system

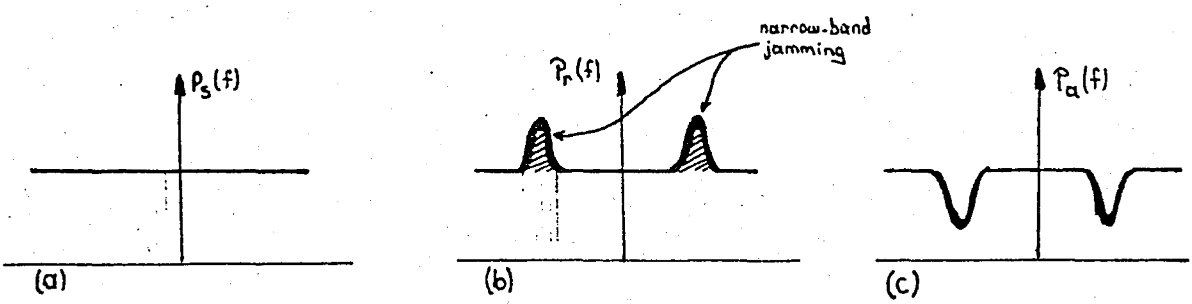


Fig.4.6 Power spectrum of FH signals
(a)transmitted FH signal
(b)received FH signal
(c) charecteristics of filter

we can easily see the effect of jamming noise.

b. Performance of FH signals in the presence of interference.

When there is a narrow-band interference in FH system, the correlator outputs with and without whitening are denoted by $S_2(t)$ and $S_1(t)$, respectively, as in DS systems. A measure of the performance improvement due to whitening can then be obtained by comparing the signal-to-noise ratios of the outputs $S_2(t)$ and $S_1(t)$. The signal-to-noise ratio can be defined by

$$SNR_i = \frac{E^2\{S_i\}}{V_{ar}\{S_i\}}, \qquad i = 1,2$$

we can calculate the values of SNR_2 and SNR_1 using the same way as in DS system and the ratio of SNR_2 to SNR_1 (W) is approximately given by

$$W = \frac{SNR_2}{SNR_1}$$

$$W = \frac{\frac{1}{2}\sum_{m=1}^{m} A_m{}^2}{\sum_{n=0}^{n} a_n^2 + MP_N \Delta f \Delta t}$$

where a_n is the prediction error filter coefficient, P_N is the average power of the error series. The performance factor, W, indicates the improvement resulting from whitening. In our computer results about FH system these values are obtained ; Average power

$$P_N = 16.9 \text{ dB}$$

$$SNR_2 = -10.983 \text{ dB}$$

$$SNR_1 = -18.403 \text{ dB}$$

Performance $W = SNR_2/SNR_1 = 11.959$ dB.

Let us consider the performance of a FH spread spectrum signal in the presence of broad-band noise characterized statistically as additive white gaussian noise with power spectral density J. For binary FSK with noncoherent detection and slow frequency hopping (1 hop/1 bit), the probabality of error (2),

$$P = \frac{1}{2} \; e^{-\gamma_b/2}$$

where $\gamma_b = \varepsilon b/J$, ε_b is the energy per bit. Finally, we observe that, ε_b, the energy per bit, can be expressed as,

$$b = \frac{E_{av}}{R}$$

Where E is the transmitted signal energy and R is the information rate in bits per second and

$$J = Jav/W$$

therefore γ_b may be expressed as

$$\gamma_b = \frac{b}{J} = \frac{E_{av}/R}{J_{av}/W}$$

$$\gamma_b = \frac{W/R}{J/E}$$

In this expression we recognize W/R as the processing gain and J/E as the jamming margin for the FH system.

V. COMPUTER AIDED ANALYSIS OF SPREAD SPECTRUM COMMUNICATIONS

5.1. DIRECT-SEQUENCE SYSTEMS

In a DS system the spreading is accomplished by biphase modulating a carrier with a high rate pseudo-random binary sequence. Binary message modulation is accomplished by multiplying the pseudo-random sequence by the message. In the receiver it is necessary to despread the incoming signal by multiplying it by a replica of the original spreading sequence.

5.1.1. TRANSMITTING AND RECEIVING OF DS SIGNALS

Spreading which is the crucual point in the field of spread spectrum communications, is accomplished by a code sequence. The high speed code sequences spoken of here are just long binary sequences (of ones and minus ones) at bit rates usually in the range from one to a few hundred megahertz.

1 -1 -1 1 1 1 1 -1 1 1

-1 1 1 -1 -1 -1 1 -1 1 -1

In our simulation, code clock rate is chosen 1 Mbps and information rate is chosen 1 kbps. It means that the main lobe of this spectrum has a bandwidth of 2 MHz. (twice the code clock rate)

$$BW = 2 \times R_{code} = 2 \times 1 \text{ MHz} = 2 \text{ MHz}$$

$$R_{data} = 1 \text{ kbps}$$

Using these values we can easily calculate the process gain of this DS system.

$$G_p = \frac{BW}{R_{data}} = \frac{2 \text{ MHz}}{1 \text{ kHz}} = 2 \times 10^3$$

$$G_p = 33 \text{ dB}$$

Code lenght is chosen 8191 bits for practicality. This is accomplished by using the theoritical explanations and tables in DS system (chapter 3.). This code sequence can be generated by using 13 stage shift registers. Now we can calculate the repetition rate of this system.

$$\text{Code repetition rate } (R_c) = \frac{\text{Code Rate}}{\text{Code length}} = \frac{\text{bps}}{\text{b}}$$

$$R_c = \frac{1 \times 10^6}{8191} = 122 \quad \text{times per second}$$

It means that our code sequence is being sent 122 times-in 1 second. In our application, there is 128 code bits per one data symbol and 1024 code bits are generated for 8 data symbols. Data signals also consist of ones and minus ones

1 -1 1 -1 1 -1 1 -1

Using these principles, code and data signals are generated and modulated. Modulation is accomplished by multiplying the code and data signals. Transmitted signal can be represented as

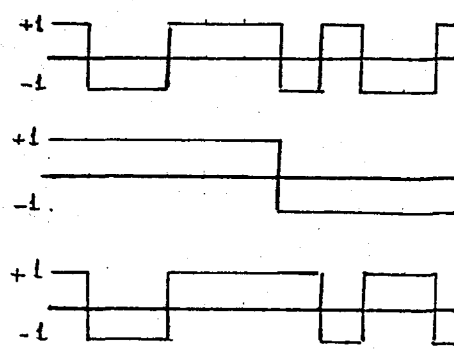
$$s(t) = p(t).d(t)$$

where p(t) is code sequence which has 1 Mbps and d(t) is data signal which has 1 kbps bit rate. S(t) is a transmitted DS signal which has a flat power spectrum shown in Fig 5.1. The values of the transmitted signal can be represented as,

p(t): 1 -1 -1 1 1 1-1 1-1 -1 1

d(t): 1 1 1 1 1 1-1 -1-1 -1 -1

s(t): 1 -1 -1 1 1 1 1 -1 1 1 -1



a. Ideal signal correlation and demodulation

Ideally we can suppose that there is no any interference and noise in channel so pure transmitted signal is received. Received signal is correlated by using the code sequence which is same as the sequence in transmitter then the demodulation techniques are applied in order to obtain the data values. It is known that there is 128 code bits per data symbol so correlated sequence consists of 128 ones and 128 minus ones sequences.

b. DS system with narrow-band interferences

The received wide-band signal can be represented as ;

$$r(t) = s(t) + i(t) + n(t)$$

where s(t) is the transmitted signal, i(t) is the narrowband interference and n(t) is the gaussian noise. The interference is modelled as a sum of fixed amplitude and fixed frequency tones. It can be represented as.

$$i(k) = \sum_{m=1}^{M} A_m \mathrm{Sin}(2\pi f_m k\Delta t)$$

where A_m is the amplitude of the interference, f_m is the frequency of the interference, M and k are the sample number of sequence and interference and Δt is the sample interval.

$A_m = 5$ $\quad M = k = 128$

$f_m = 1000$ Hz $\quad \Delta t = 0.000194$

we can change the value of Δt and sample number for obtaining the better state for interference. The power spectrum of the received signal with narrowband interference is shown in Fig 5.2. Using these values we can state the interference on the frequency band between 40^{th} and 50^{th} slots. How can we suppress this narrow-band interference? Rejection of this interference consists of two steps. First step is estimation (linear prediction explained in section 3.), second step is suppression.

A transversal filter is designed for estimating the interference by using levinson-Durbin algorithm. Better suppression can be obtained changing the order of filter. **Input and output data values for a Direct sequence system are shown in fig.5.3.** The best suppression is obtained using the filter with order 8. shown in Fig. 5.4. After filtering and correlation, the demodulation is employed and the new data values are obtained

0.598 -0.996 0.470 -0.384 0.441 -0.515

0.295 -0.387

It is easly seen that we can't obtain the ideal data values (ones and minus ones) because of the noisy channel. In other words we can say that Filtering is not enough to ideally suppress the interference. After filtering, the received signal's power spectrum is plotted in Fig. 5.5. It is convenient to compare Fig. 5.2. and 5.5. in order to see the effects of filter and interference. The error spectrum, which is obtained using the difference values between filtered and ideal DS signals, is plotted in Fig. 5.6. We can see the errors on every frequency components in this figure. The values of input and output data are plotted in Fig. 4.7. in order to show the performance of DS system.

c. Performance criteria for DS system.

In chapter 3 the correlator outputs for a spread spectrum receiver with and without whitening are denoted by $S_2(t)$ and $S_1(t)$. A measure of the performance improvement due to filtering can then be obtained by comparing the SNR values of the outputs $S_2(t)$ and $S_1(t)$. Using the theoritical explanations in chapter 3, we can calculate the SNR values,

$$SNR_2 = \frac{L}{\sum_{n=0}^{N} a_n^2 + MP_N \Delta f \Delta t}$$

where L is the number of chips per symbol, M is the sample number, Δf is the frequency separation between two neighboring frequencies, a_n is the filter coefficient and P_N is the average power.

$$P_N = P_{N-1}\,(1-a_{N,N}^2) = 11.07 \text{ dB}$$

$$P_o = \frac{1}{M}\sum_{k=0}^{M-1} s(k)^2$$

$$L = 128$$

$$SNR_2 = -9.477 \text{ dB}$$

$$SNR_1 = \frac{L}{\sum_{m=1}^{M} \frac{1}{2} A_m^2} = 15.229 \text{ dB}$$

now wecan find the performance value W

$$W = \frac{SNR_2}{SNR_1} = SNR_2(\text{dB}) - SNR_1(\text{dB}) = 7.118 \text{ dB}.$$

5.2. FH. SYSTEMS

FH results from frequency shift keying a carrier with the binary message and then hopping this carrier over a wide range of frequencies in accordance with some pseudo-random frequency pattern.

5.2.1. TRANSMITTER PART OF FH SYSTEM

In our computer aided FH system analysis, there are 32 code bits per one data symbol,

R_{data} = Data Rate = 10 kbps.

R_{code} = Code Rate = 0.32 Mbps.

ransmitted BW = 2 x 0.32 MHz = 0.64 MHz.

Code sequence consists of ones and minus ones as in DS system but it changes from frequency to frequency at every 32 bits blocks. Code sequence values are shown in Fig. 5.8. Data values are the same as in DS system.

1 -1 1 -1 1 -1 1 -1

Every data bit is modulated by the code sequence at the different frequencies.

5.2.2 RECEIVER PART OF FH SYSTEM

Power spectrum of the ideal transmitted signal is shown in Fig. 4.9. Ideally there is not any difference between FH and DS system receiver parts. Correlation with the same code sequence and demodulation are employed and above data values are obtained. When there is a zero mean gaussian noise in the received signal, we can't see the any jammed region on the spectrum because the transmitted signal and noise have the same flat spectrum, shown in Fig. 5.10. After correlating and demodulation, we can see the effect of noise on data values.

1.032 -1.020 1.099 -1.118

Correlated FH values are plotted in Fig. 5.11.

When there is a narrow-band interference, the message is obtained, using the same techniques as in DS system. Jammed FH signal's power spectrum is plotted in Fig. 5.12. Interference is estimated (linear prediction) and suppressed for obtaining the data as shown in Fig. 5.13 and Fig. 5.14. New data values are ;

0.474 -0.548 0.608 -0.579 0.681 -0.680

0.784 -0.736

We can see the correlated FH signals in Fig. 5.14.

Performance criteria of FH system is accomplished using the same rules as in DS system and

$SNR_2 = -10.983$ dB

$SNR_1 = -18.403$ dB

$W = 11.31$ dB

where Average power = 16.9 dB

$L = 32$ and $\Delta t = 0.119$

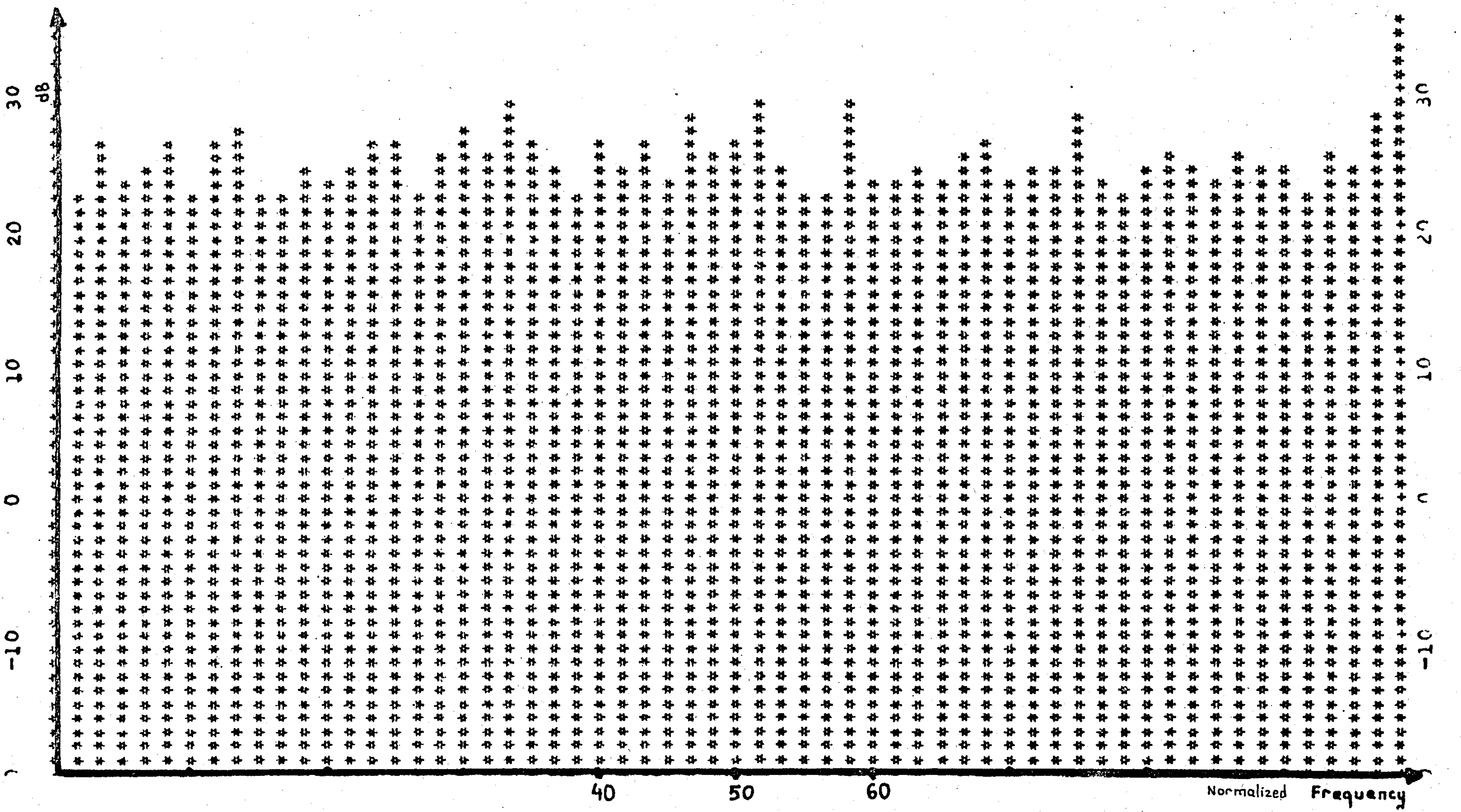
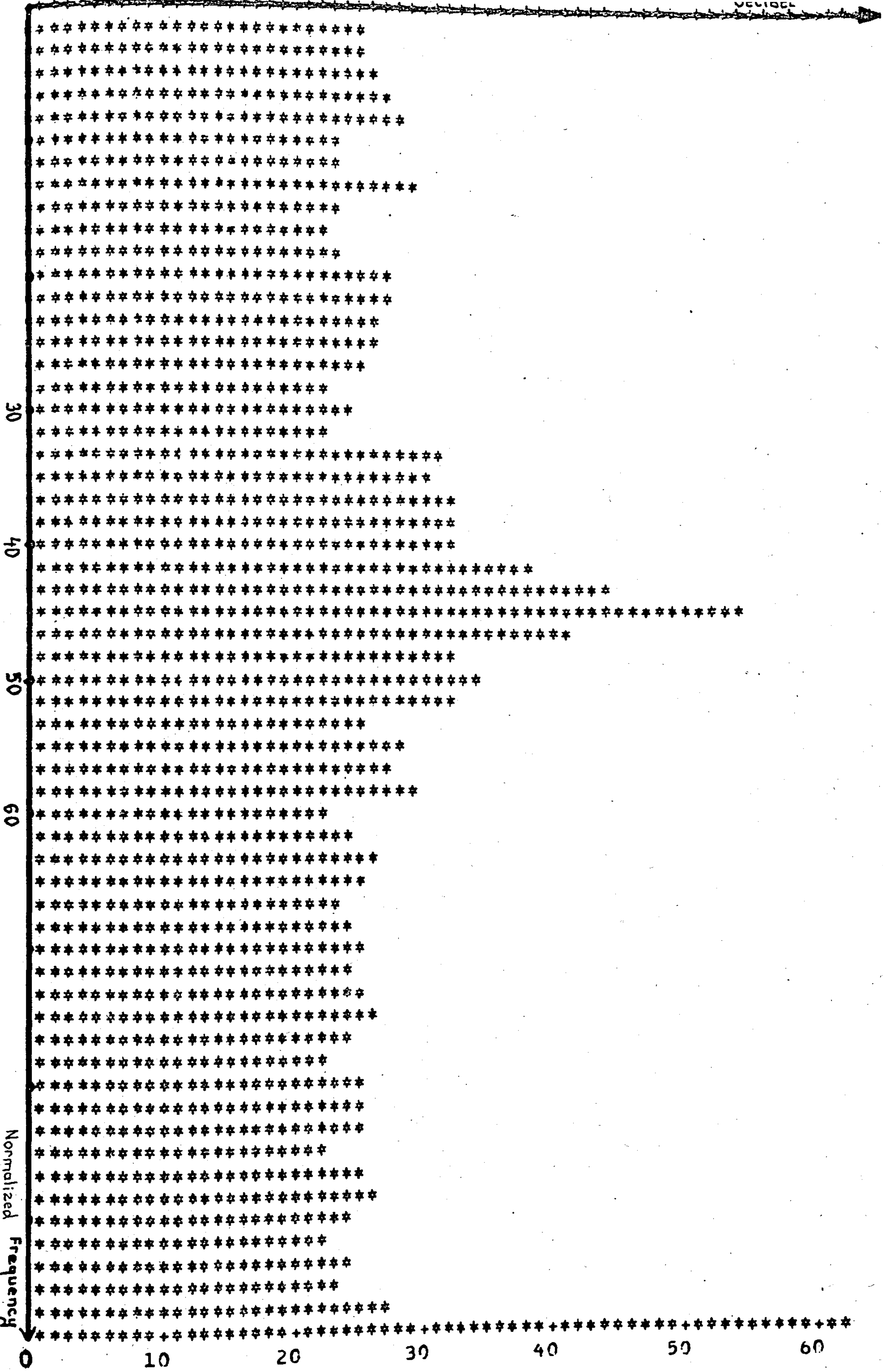
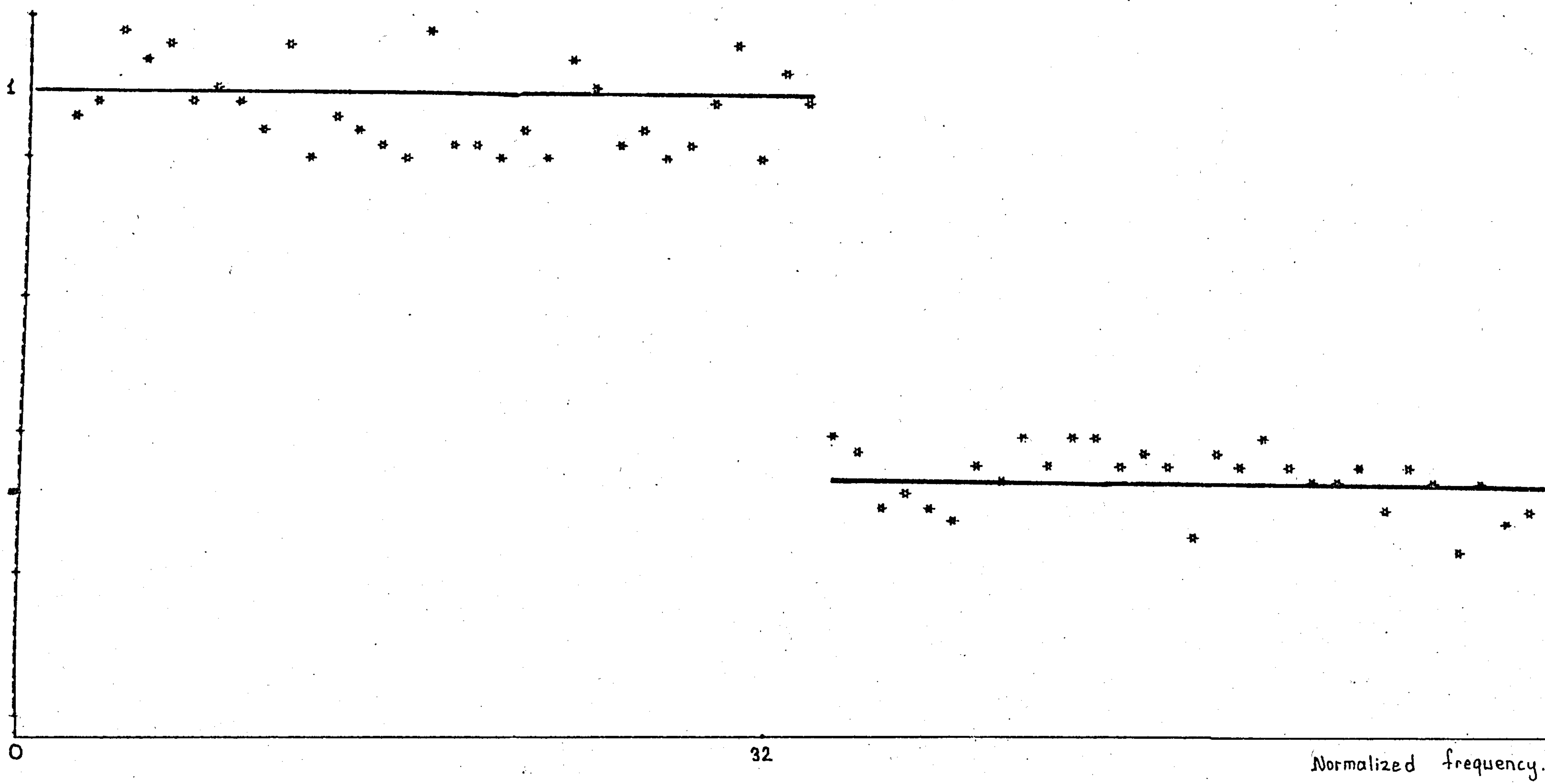


Fig.5.1. Power spectrum of transmitted DS signals


Decibel
30
40
50
60
Normalized Frequency
0
10
20
30
40
50
60


1
0
32
Normalized frequency.

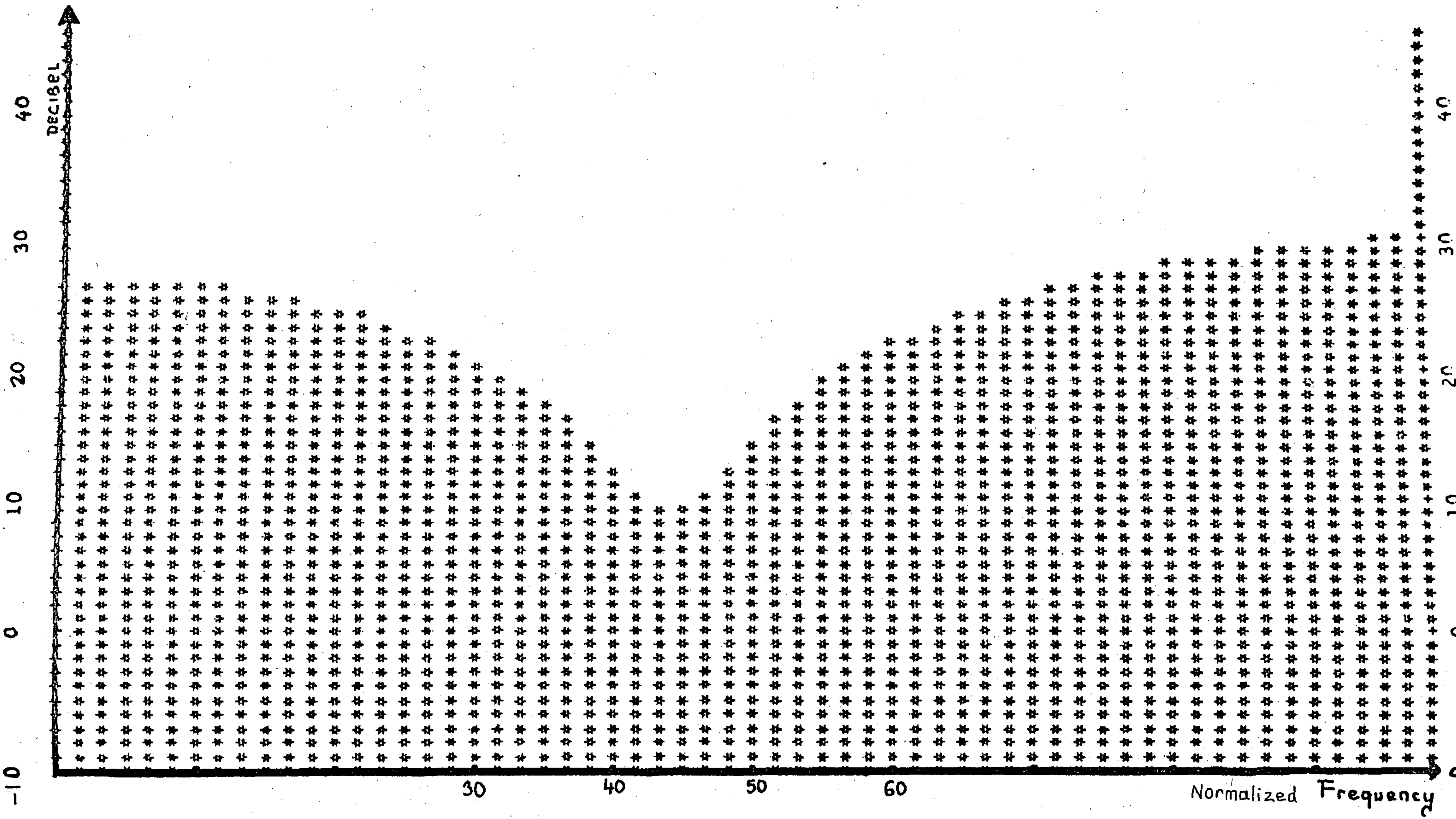
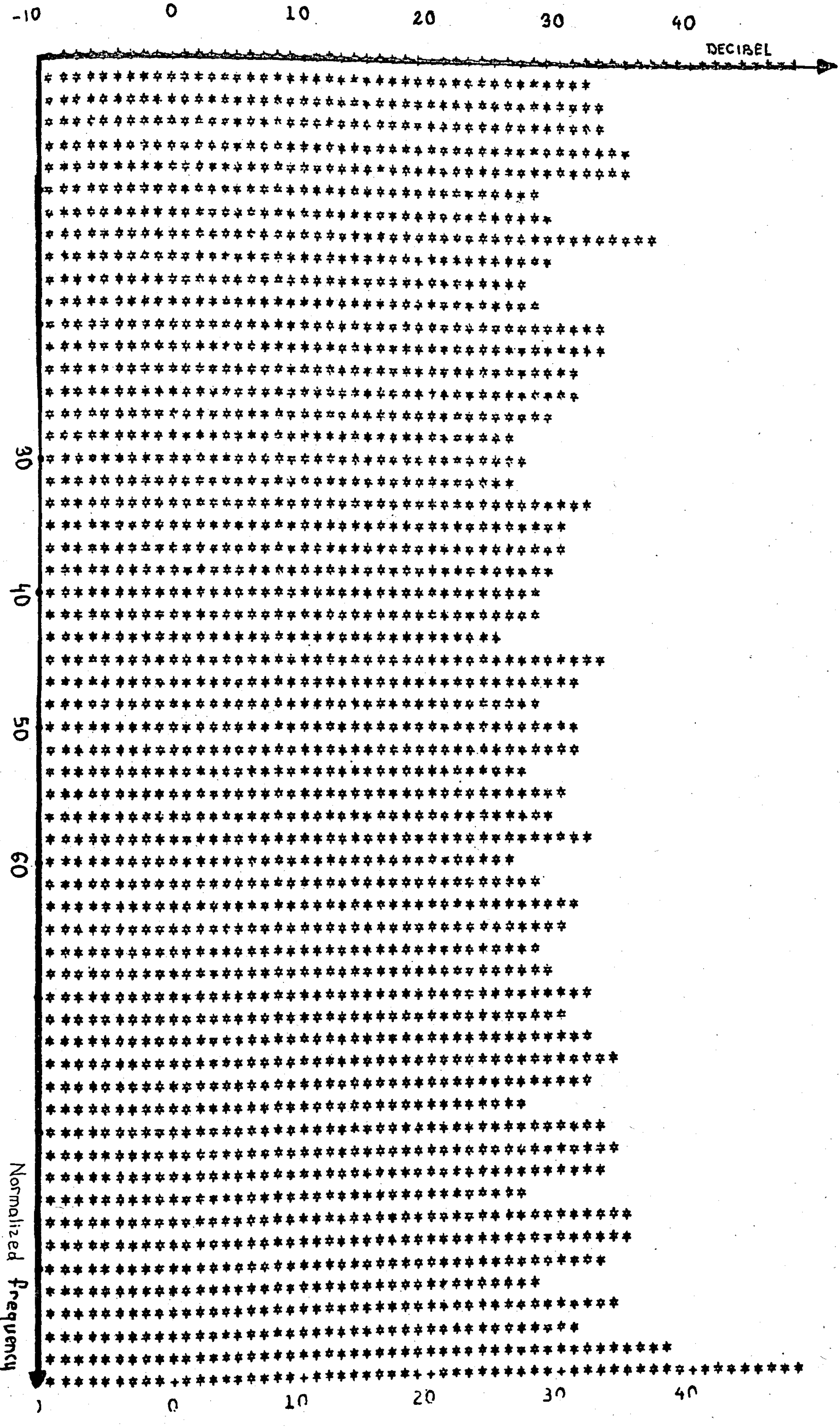
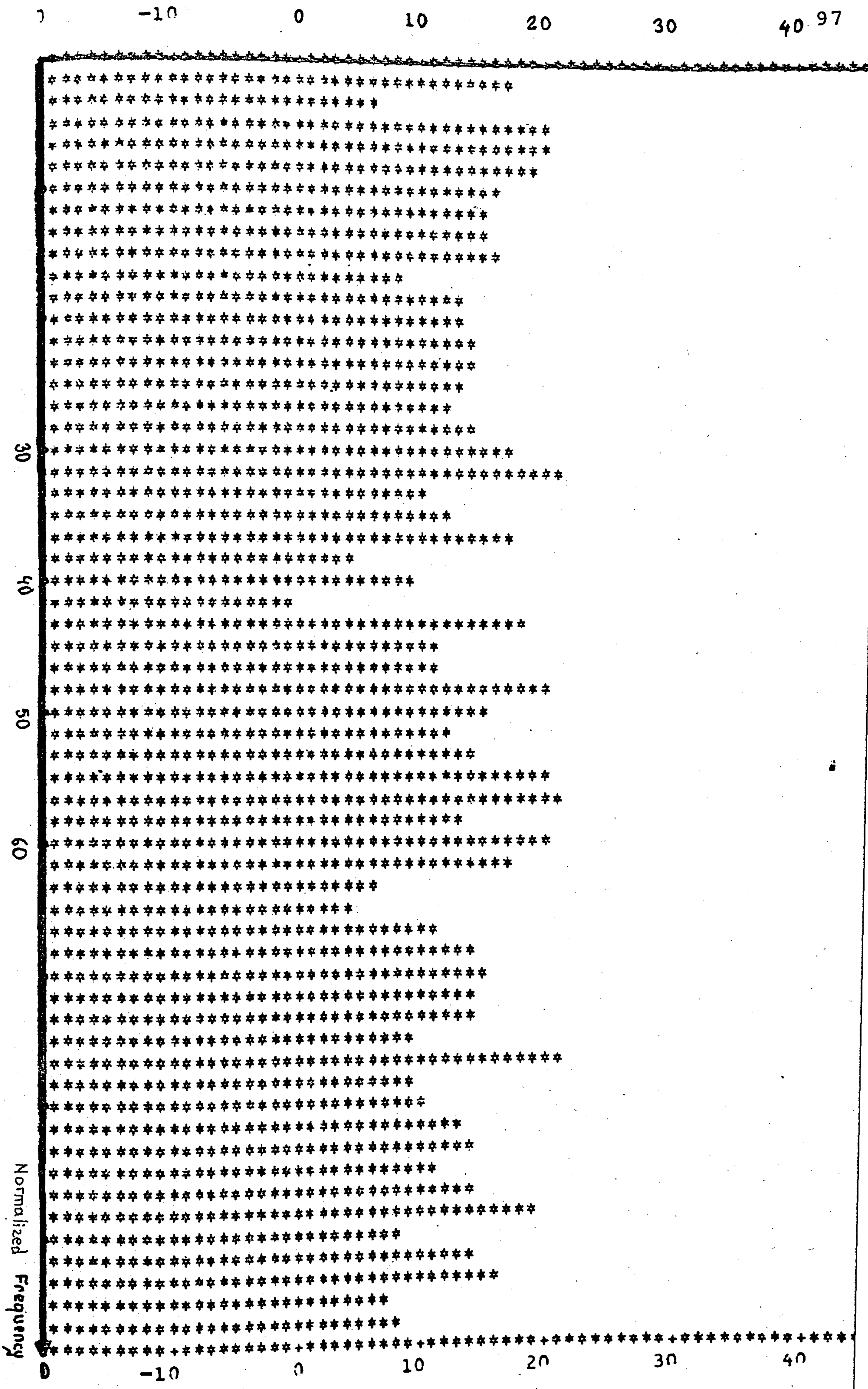


Fig.5.4. Linear prediction filter with order eight


-10
0
10
20
30
40
DECIBEL
30
40
50
60
Normalized frequency


-10
0
10
20
30
40
30
40
50
60
Normalized Frequency

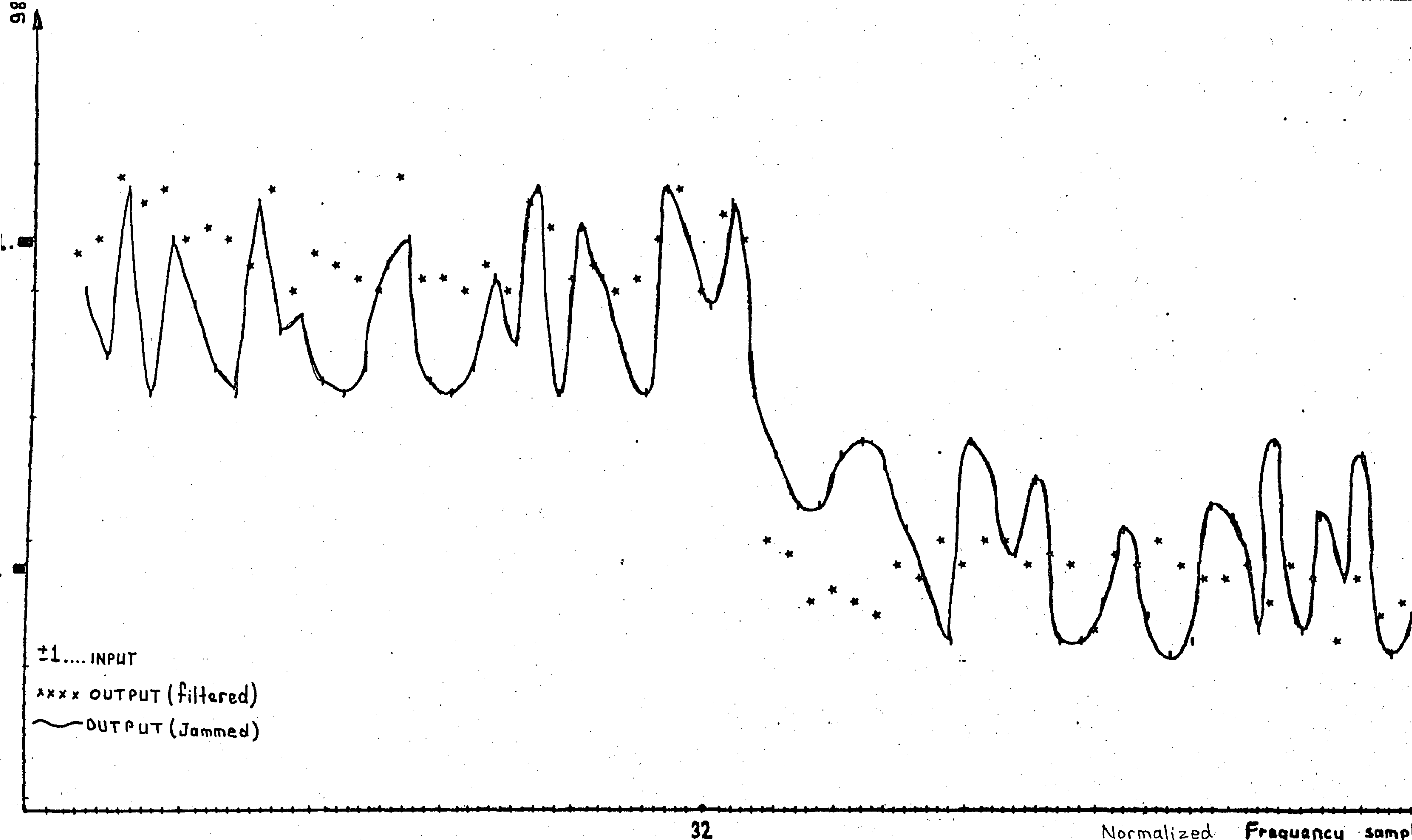


Fig.5.7.Input and output values for DS system.

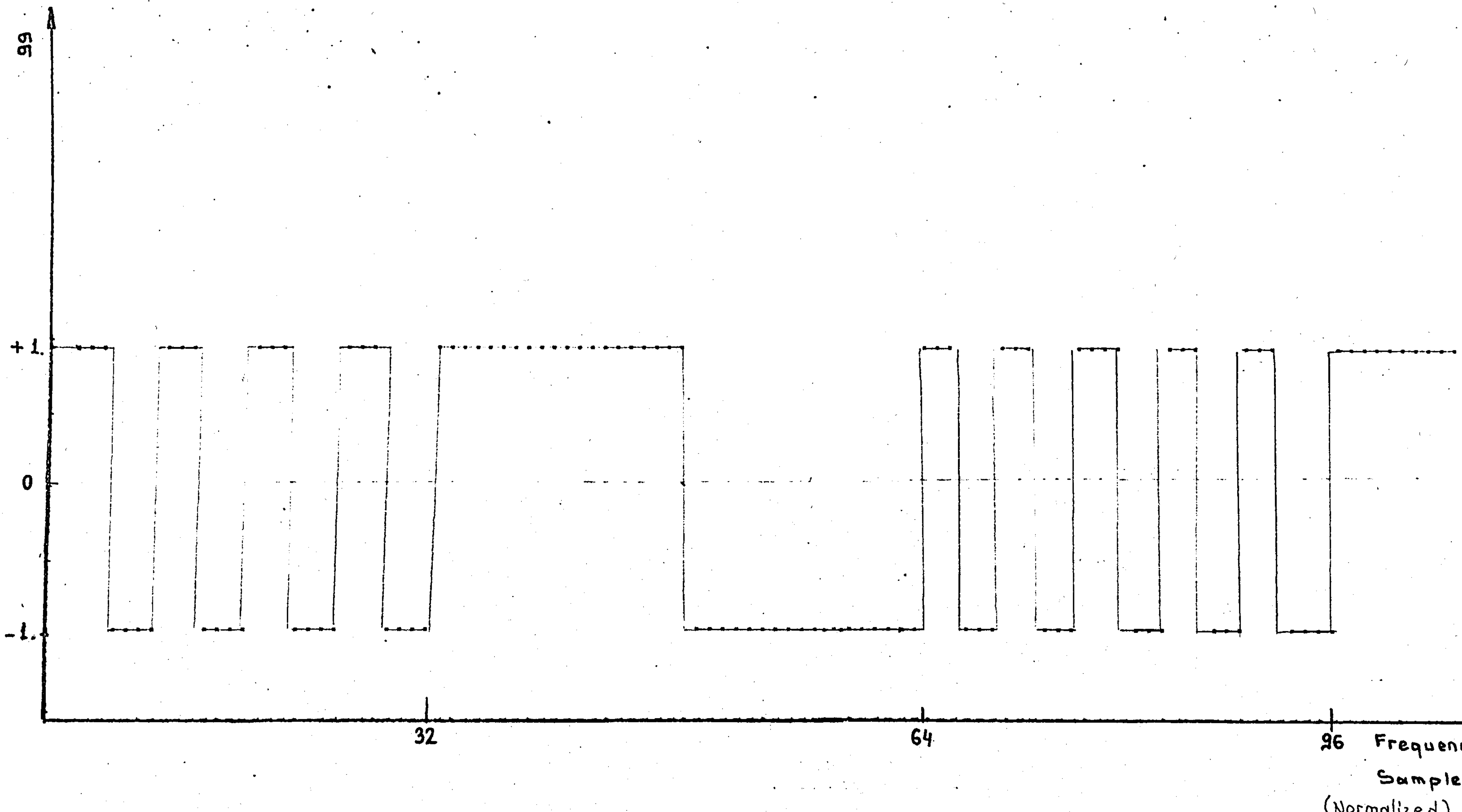
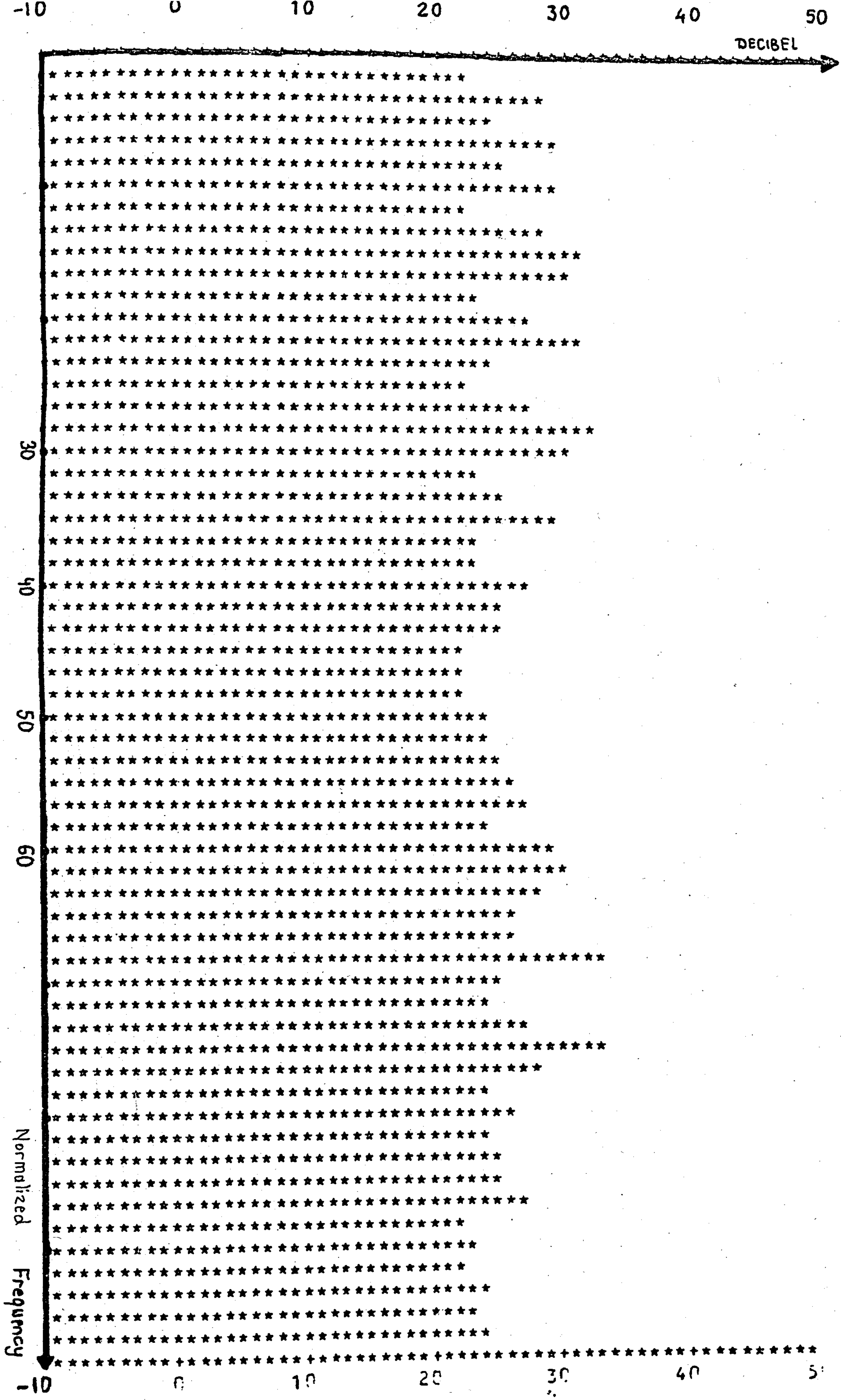


Fig.5.8. FH. Code sequence values.


-10
0
10
20
30
40
50
DECIBEL
30
40
50
60
Normalized Frequency

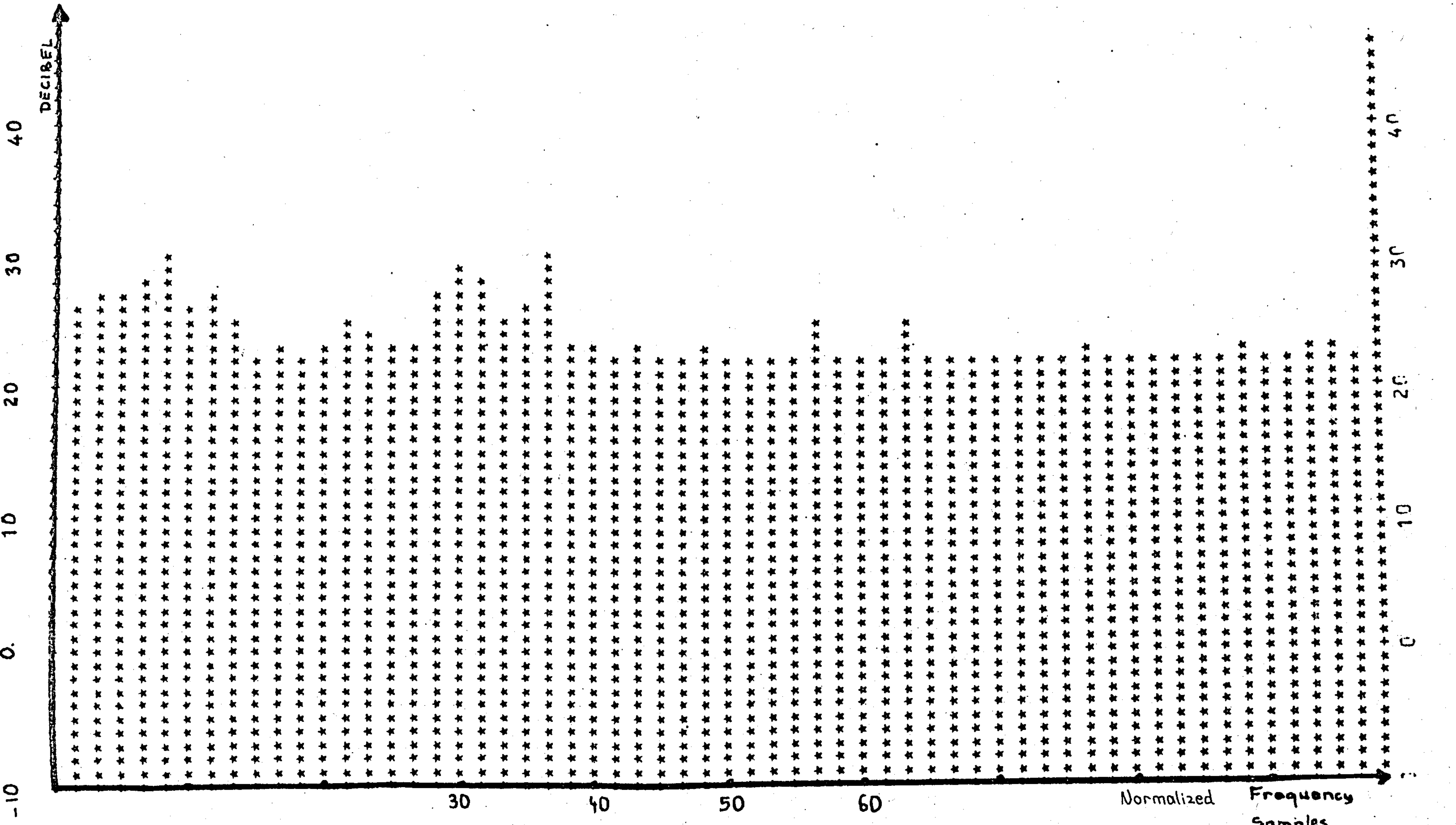
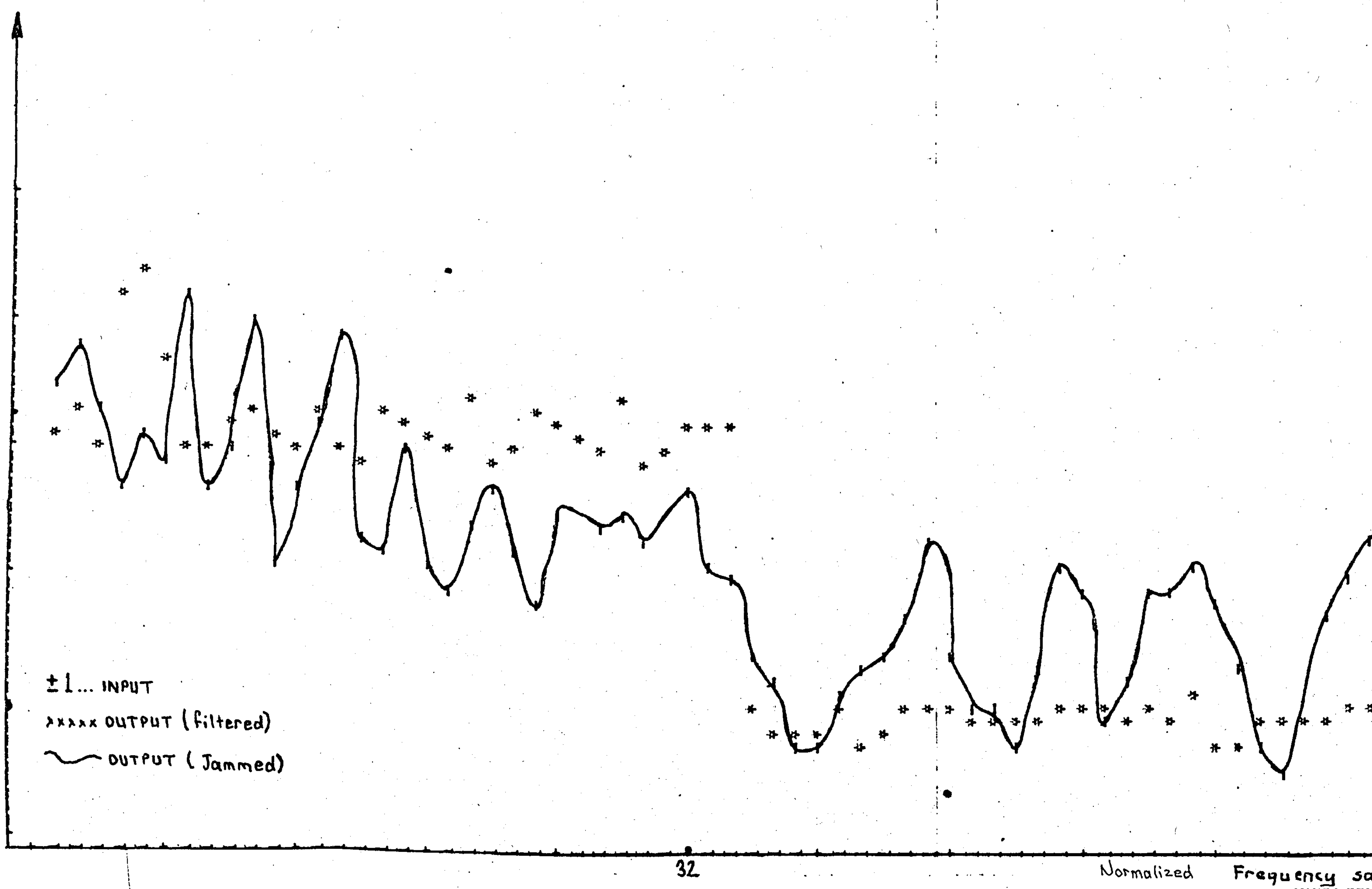
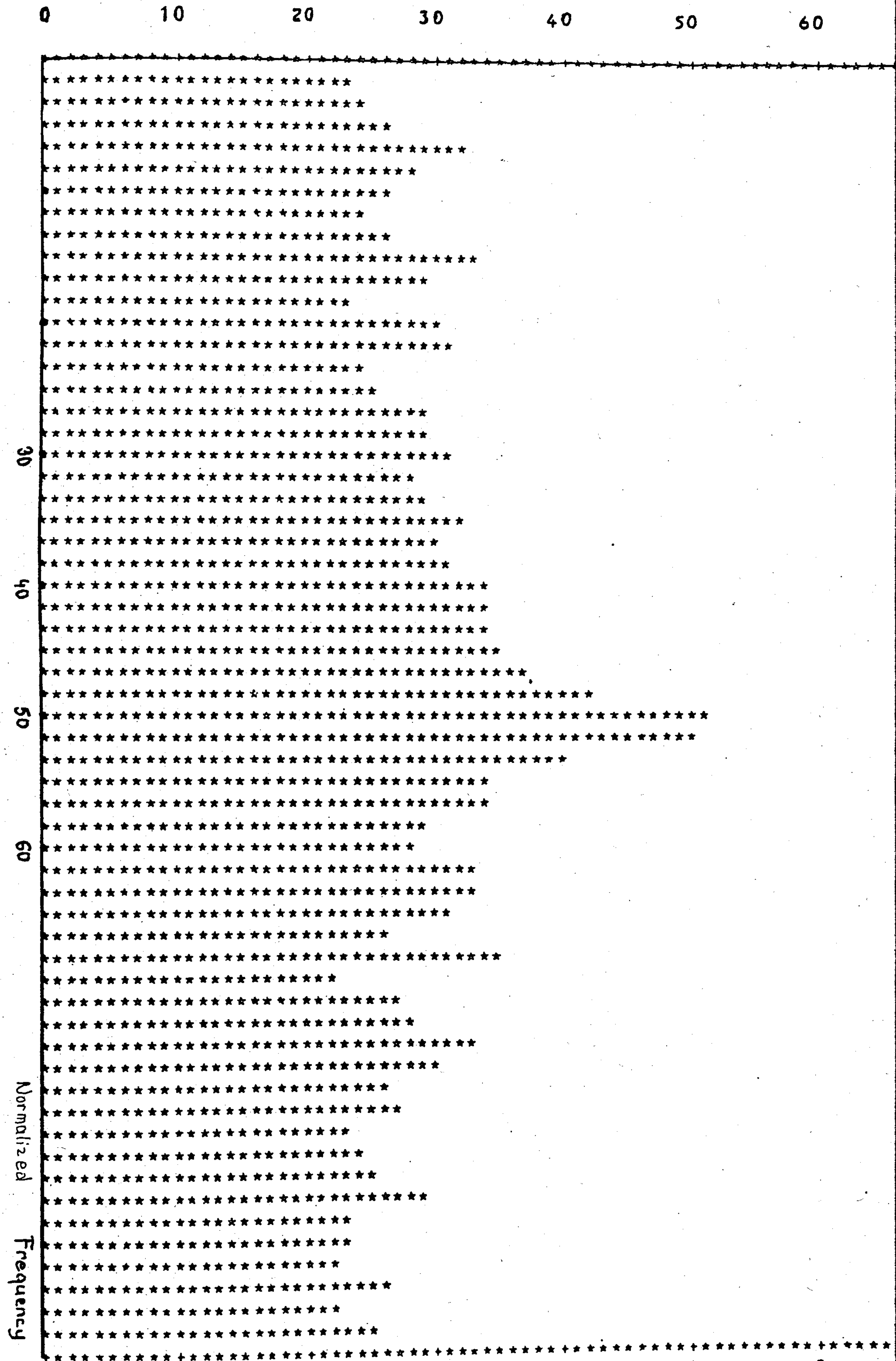


Fig.5.10.Power spectrum of FH signal with noise.


+1.
-1.
±1... INPUT
xxxxx OUTPUT (filtered)
OUTPUT (Jammed)
32
Normalized Frequency sam

POWER SPECTRUM



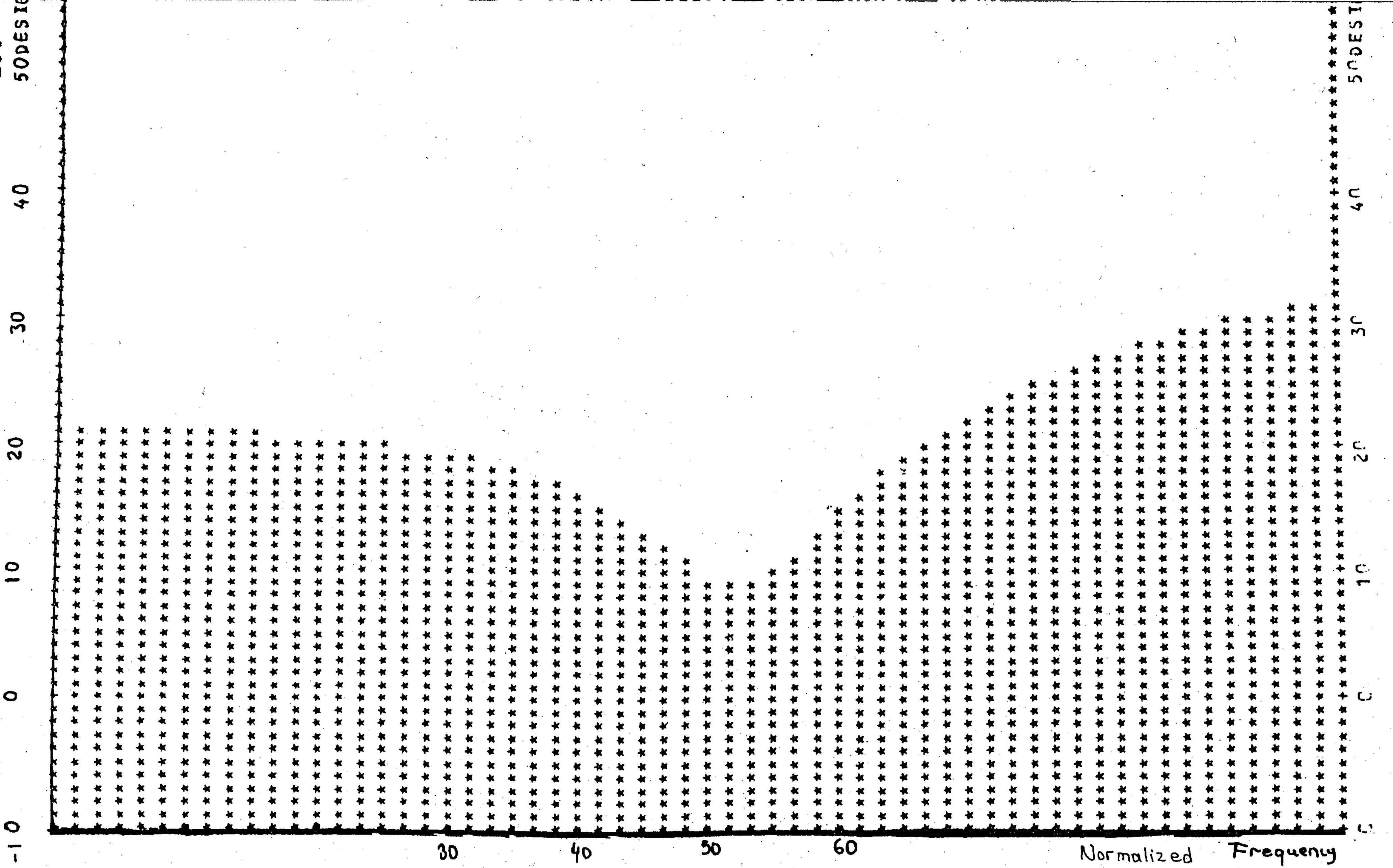
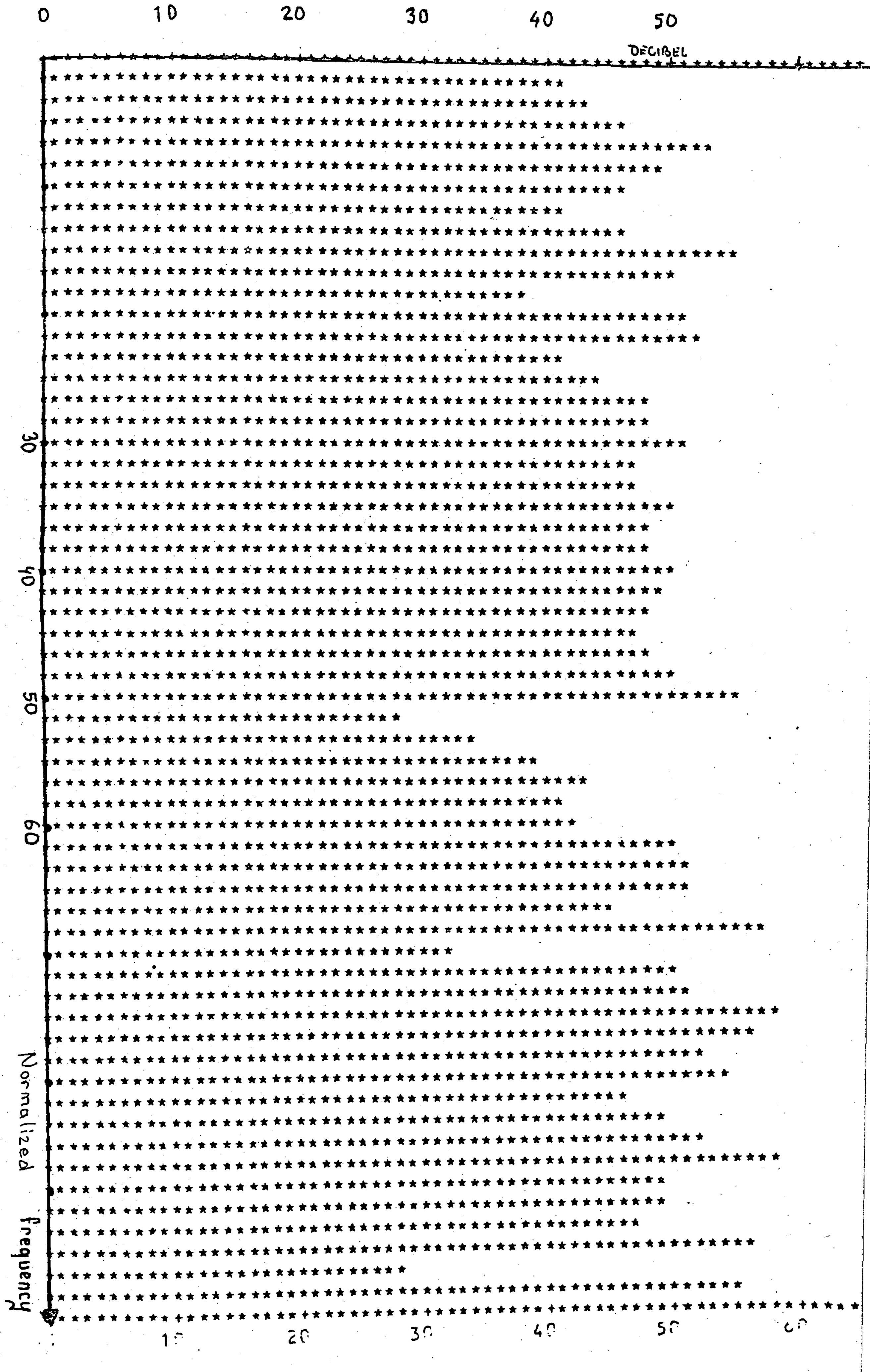


Fig.5.13.Linear prediction filter with order eight.


0
10
20
30
40
50
DECIBEL
30
40
50
60
Normalized frequency

```
      PROGRAM BEKIR(WAA,GEL,TAPE5=WAA,TAPE6=GEL)
      COMMON /A/X(2000),Y(2000),B(2000),H(2000)
      COMMON /A/C(2000),E(500),F(2000),Z(2000)
      COMMON /A/D(2000),G(500),R(2000),AL(2000)
      COMMON /A/P(2000),V(2000),W(2000),GI(2000)
      COMMON /A/DE(100),U(500),S(800),XE(300)
      COMMON /A/ANPT(300),CE(300),A(20,20)
      COMMON /A/RR(40),ZE(300),YE(300),UE(300)
      COMMON /A/RE(300),PE(300),SE(300),VE(300)
      COMMON /A/CC(300),COEF(50),BE(300),GE(300)
      COMMON /CPROG/IWRT,ANKT,ERROR,KERRO,SNR1
      COMMON /B/SNR2,PRFH,SNR,T,M,N,SS,ERRO
      COMMON /B/AC,MM,BA,RV
      COMMON /CSIGNL/ND,NP,NP1,TS,AG,AP,DELF,DELT
*************************************************
*                                              **
*   GENERATION  OF  CODE  SIGNALS   Z(I)       **
*                                              **
*************************************************
C     SS=1  FREQUENCY  HOPPING  SIMULATIONS
C     SS=0  DIRECT  SEQUENCE  SIMULATIONS
      SS=1.
      IF(SS.EQ.1.) GO TO 2001
      YY=0.
      ISEED=256
      CALL RANSET(ISEED)
      DO 60 I=1,1024
      T=RANF()
      IF (T.GE.0.5) T=1.
      IF (T.LT.0.5) T=-1.
      Z(I)=T
      V(I)=Z(I)
   )  CONTINUF
      WRITE(6,102) (Z(I),I=1,20)
   2  FORMAT(//,20X,'FIRST  CODE  VALUES',///,10
     +(10F6.1,//))
      IF(YY.EQ.1) WRITE(6,103) (Z(I),I=1001,1020)
   3  FORMAT(//,10X,'CODE  VALUES  BETWEEN 1001 &
     +1020',///,10(10F6.1,//))
*************************************************
*                                              **
*   GENERATION  OF  DATA  SIGNALS    D(I)      **
*                                              **
*************************************************
      DO 1 I=1,32,2
      D(I)=1.
   1  D(I+1)=-1.
      WRITE(6,15)  (D(I),I=1,32)
   5  FORMAT(//,20X,'DATA VALUES',///,10
     +(10F6.1,//))
*************************************************
*                                              **
*   MODULATION  (DATA*CODE)          Z(I)      **
*                                              **
*************************************************
      J=1
      ILK=1
      ISON=32
   0  CONTINUE
      DO 51 I=ILK,ISON
      T=D(J)
      IF(T.EQ.-1.) GO TO 52
      Z(I)=-Z(I)
   2  Z(I)=-Z(I)
   1  CONTINUE
   3  IF(ISON.EQ.1024) GO TO 501
      J=J+1
      ILK=ILK+32
      ISON=ISON+32
      GO TO 500
   1  CONTINUE
      WRITE(6,56) (Z(I),I=1,20)
   6  FORMAT(//,20X,'FIRST  MODULATED  VALUES',//
     +10(10F6.1,//))
      IF(YY.EQ.1)  WRITE(6,57) (Z(I),I=1001,1020
   7  FORMAT(//,10X,'MODULATED  VALUES  BETWEEN
     +& 1020',//,10(10F6.1,//))
```

```
**************************************************
**                                              **
**           D E M O D U L A T I O N            **
**                                              **
**   MULTIPLAYING OF OUTPUT & CODE VALUES       **
**                                              **
**************************************************
      DO 13 I=1,1024
  13  P(I)=Z(I)*V(I)
      WRITE(6,74) (P(I),I=1,20)
  74  FORMAT(//,20X,'DEMODULATED  VALUES',///,10
     +(10F6.1,//))
      ILK=1
      ISON=32
1019  CONTINUE
      DO 1031 I=ILK,ISON
1031  X(I)=P(I)+1.
      ILK=ILK+64
      ISON=ISON+64
      IF(ISON.LT.1026) GO TO 1019
      ILK=33
      ISON=64
1032  CONTINUE
      DO 1033 I=ILK,ISON
1033  X(I)=P(I)-1.
      ILK=ILK+64
      ISON=ISON+64
      IF(ISON.LT.1026) GO TO 1032
      XAX=5.
      YAX=13.
      N=128
      DO 737 I=1,N
 737  B(I)=I
C     IF(YY.EQ.0.) CALL PLOT(XAX,YAX,128,X,B,'*')
      J=0
      DO 14 I=1,1024,32
      J=J+1
  14  D(J)=P(I+6)
      WRITE(6,75) (D(I),I=1,8)
  75  FORMAT(//,20X,'NEW  DATA  VALUES',///,10
     +(10F6.1,//))
**************************************************
**          SQUARE   WAVE   INPUT    S(I)       **
**************************************************
      DO 3 I=1,127
   3  S(I)=0.
      DO 4 I=384,512
   4  S(I)=0.
      DO 2 I=129,383
   2  S(I)=1.
**************************************************
**       SPECIAL    CODED   SEQUENCE   U(I)     **
**************************************************
      K1=1
      K2=20
  81  CONTINUE
      DO 80 I=K1,K2
  80  U(I)=-1.
      K1=K1+20
      K2=K2+20
      DO 89 I=K1,K2
  89  U(I)=1.
      K1=K1+20
      K2=K2+20
      IF(K1.GT.300) GO TO 82
      GO TO 81
  82  CONTINUE
**************************************************
**        SECOND   CODED   INPUT     B(I)       **
**************************************************
      ISEED=256
      CALL RANSET(ISEED)
      DO 61  I=1,1024
      T=RANF()
      IF(T.GE.0.5) T=1.
      IF(T.LT.0.5) T=-1.
      B(I)=T
  61  CONTINUE
      IF(YY.EQ.1)  WRITE(6,62) (B(I),I=1,20)
  62  FORMAT (//,15X,'SECOND  CODE  VALUES,'///,10
     +(10F6.1,/))
```

```
****************************************************
        RANDOM    SIGNAL  (CODED)  F(I)         **
****************************************************
   ISEED=956
   CALL RANSET(ISEED)
   DO 70 I=1,1024
0  F(I)=RANF()
   IF(YY.EQ.1) WRITE(6,71) (F(I),I=1,20)
1  FORMAT(//,20X,'RANDOM  INPUT  VALUES,'///,10
  +(10F6.3,//))
****************************************************
        RANDOM     DATA   (CODED)  W(I)         **
****************************************************
   ISEED=956
   CALL RANSET(ISEED)
   DO 20 I=1,256
   U(I)=RANF()
0  W(I)=U(I)
   IF(YY.EQ.1.) WRITE(6,72) (U(I),I=1,10)
2  FORMAT(//,20X,'RANDOM  DATA  VALUES (W),'///,
  +10(10F6.3,//))
****************************************************
     F(I)*U(I)     MODULATION        E(I)       **
****************************************************
   N1=1
   N2=32
   J=1
3  CONTINUE
   DO 21 I=N1,N2
1  E(I)=F(I)*U(J)
   IF(N2.EQ.1024) GO TO 22
   N1=N1+32
   N2=N2+32
   J=J+1
   GO TO 23
2  CONTINUE
   IF(YY.EQ.1) WRITE(6,73) (E(I),I=1,10)
3  FORMAT(//,20X,'MODULATED  VALUES',///,10
  +(10F6.3,//))
****************************************************
                                                **
      RECEIVED   SIGNAL     (IDEAL)             **
                 ( Z(I) )                       **
****************************************************
   DO 151 I=1,256
   Y(I)=0.
1  X(I)=Z(I)
   CALL FFTA(X,Y,8,1)
   DO 152 I=1,256
2  G(I)=X(I)**2+Y(I)**2
   AG=100.
   CALL LPLOT(60,G,80,8,0.,0.100,200.)
****************************************************
                                                **
  RECEIVED  SIGNAL    WITH   INTERFERENCE       **
                ( Z(I)+C(I) )                   **
****************************************************
   PI=3.141585
   AC=5.
   MM=256
   DO 121 I=1,MM
   C(I)=AC*(SIN(2*PI*1000*0.0001045*FLOAT(I-1)))
   CE(I)=C(I)
1  CONTINUE
   DO 122 I=1,256
   X(I)=Z(I)+C(I)
   Y(I)=0.
2  CONTINUF
   CALL FFTA(X,Y,8,1)
   DO 65 I=1,256
   AL(I)=X(I)
   D(I)=Y(I)
5  H(I)=X(I)**2+Y(I)**2
   IF(YY.EQ.1) WRITE(6,66) (H(J),J=1,50)
6  FORMAT(///,15X,'OUTPUT VALUES',///,10(5F12.4,/
   DO 28 I=1,256
8  GI(I)=H(I)
   AG=100.
   CALL LPLOT(60,GI,80,8,0.,0.100,200.)
   DO 590 I=1,256
0  C(I)=15.*(SIN(2*PI*I/1.99))
```

```
*****************************************************
**                                                 **
**          INTERFERENCE    ESTIMATION             **
**                 (LEVINSON)                      **
*****************************************************
      ND=256
      C0=93.
      YY=4000.
      TS=1./8000.
      IWRT=0
      DO 531 I1=8,8
      NP=I1
      ML=NP
      DO 591 I=1,256
591   XE(I)=Z(I)+C(I)
      IF(NP.EQ.8) GO TO 699
      C0=78.773
      ND=256
      YY=1.
699   CONTINUE
      CALL AUTOCO(XE,DE)
      WRITE(6,600) ND
600   FORMAT(1X,'NUMBER OF DATA SAMPLES',I10)
      WRITE(6,601) NP
601   FORMAT(1X,'ORDER OF PREDICTOR',I10)
*****************************************************
**        PRACTICAL  SNR  (WITHOUT FILTER)         **
*****************************************************
      DO 801 K=1,256
      ANPT(K)=Z(K)**2+ANPT(K-1)
801   ANKT=ANPT(K)*2
      ERROR=(AC)**2/ANKT
      ERROR=ABS(ERROR)
      SNR=10.*ALOG10(ERROR)
      WRITE(6,802) SNR
802   FORMAT(1X,'PRACTICAL SNR (DB)',10F10.3)
*****************************************************
**                MINIMUM  ERROR                   **
*****************************************************
      DO 811 I=1,ML
811   ANPT(I)=DE(I)*RR(I)+ANPT(I-1)
      ERRO=RR(1)+ANPT(ML)
      ERRO=ABS(ERRO)
      ERRO=10.*ALOG10(ERRO+0.01)
      WRITE(6,812) ERRO
812   FORMAT(1X,'THEORITICAL MINIMUM ERROR (DB)',10F8
      DO 603 K=1,256
      E(K)=0.
603   G(K)=0.
      DO 605 K=1,NP
605   E(K)=DE(K)
      CALL FFTA(E,G,8,1)
      DO 602 I=1,256
602   S(I)=E(I)**2+G(I)**2
      IF(YY.EQ.0) WRITE(6,597) (S(I),I=1,50)
597   FORMAT(///,10X,'SECOND OUTPUT',///,10(5F12.4,/
      AG=10.
      CALL LPLOT(60,S,80,8,0.,0.100,10.)
*****************************************************
**             INTERFERENCE  SUPPRESSION           **
*****************************************************
      NP=256
      DO 556 I=1,NP
      U(I)=0.
556   R(I)=(H(I)/9.)*(S(I)/9.)
      AG=100.
      CALL LPLOT(60,R,80,8,0.,0.100,500.)
      DO 561 I=1,NP
      X(I)=(E(I)*AL(I))-(G(I)*D(I))
561   Y(I)=(AL(I)*G(I))+(E(I)*D(I))
      CALL FFTA(X,Y,8,-1)
      DO 569 I=1,256
      P(I)=X(I)
569   X(I)=X(I)/C0
      WRITE(6,563) (X(I),I=1,10)
563   FORMAT(//,20X,'INVERSE  FFT',//,5(10F8.3,//))
      DO 567 I=1,256
      Y(I)=0.
567   F(I)=ABS(Z(I)-X(I))
      CALL FFTA(F,Y,8,1)
      DO 568 I=1,256
568   H(I)=F(I)**2+Y(I)**2
      CALL LPLOT(60,H,80,8,0.,0.100,1.)
C
C
C
C
C
C
```

```
THE LAST  CORRELATION  (DEMODULATION)
   J=1
   ILK=1
   ISON=32
   DO 1001 I=1,256
   X(I)=P(I)
   H(I)=V(I)*X(I)
01 B(I)=H(I)/93.822
   WRITE(6,1004) (B(I),I=1,10)
04 FORMAT(//,20X,' FILTERING',//,10(10F8.3,////))
02 CONTINUE
   DO 1005 I=ILK,ISON
   X(I)=H(I)**2+X(I-1)
   D(J)=X(ISON)/32
   G(I)=ABS(B(I))
05 G(I)=G(I)+1.
   D(J)=D(J)/YY
   ISO=32+ISON
   X(ISO)=0.
   J=J+2
   ILK=ILK+64
   ISON=ISON+64
   IF(ISON.LT.257.) GO TO 1002
   J=2
   ILK=33
   ISON=64
08 CONTINUE
   DO 1009 I=ILK,ISON
   X(I)=H(I)**2+X(I-1)
   X(I)=ABS(X(I))
   D(J)=-1.*(X(I)/32.)
   G(I)=-1.*(ABS(B(I)))
09 G(I)=G(I)-1.
   D(J)=D(J)/YY
   ISO=32+ISON
   X(ISO)=0.
   J=J+2
   ILK=ILK+64
   ISON=ISON+64
   IF(J.LT.9)   GO TO 1008
   WRITE(6,1003) (D(I),I=1,8)
03 FORMAT(//,20X,'NEW DATA ',//,10(8F10.3,////))
   WRITE(6,1017) (G(I),I=1,10)
17 FORMAT(//,20X,'DEMODULATED',//,10(10F10.3,///,
   DO 1018 I=1,256
18 GI(I)=G(I)
   XAX=5.
   YAX=13.
   N=128
   DO 736 I=1,N
36 B(I)=I
   CALL PLOT(XAX,YAX,128,GI,B,'*')
*****************************************************
          PERFORMANCE  CRITERIA
*****************************************************

 CALCULATING  SNR2

(N)=AVERAGE POWER
   DO 401 I=1,256
1  X(I)=Z(I)+CE(I)
   M=256
   N=ML
   DO 402 I=1,M-1
2  X(I)=X(I)**2+X(I-1)
   V(1)=X(M-1)/M
   DO 403 K=2,N
   V(K)=V(K-1)*(1-A(N,N)**2)
3  AP=V(K)
   ERRO=ABS(AP)
   ERRO=10.*ALOG10(ERRO+0.01)
   WRITE(6,404) ERRO
4  FORMAT(1X,'AVERAGE POWER (DB)',10F10.2)
   BA=128.
   DELF=1000./256.
   DELT=0.0001941
   DO 406 I=1,N
6  COEF(I)=DE(I)**2+COEF(I-1)
   ERROR=BA/(COEF(N)+MM*AP*DELF*DELT)
     ERROR=ABS(ERROR)
     SNR=10.*ALOG10(ERROR+0.01)
     WRITE(6,408)SNR
408  FORMAT(1X,'SNR2 (DB)',10F8.3)

     CALCULATING  SNR1
     AP=MM*(AC)**2
     ERRO=BA/AP
     ERRO=ABS(ERRO)
     SNR=10.*ALOG10(ERRO+0.01)
     WRITE(6,409) SNR
409  FORMAT(1X,'SNR1 (DB)',10F8.3)

     PERFORMANCE  CRITERIA
     PRFM=ERROR/ERRO
     PRFM=10.*ALOG10(PRFM+0.01)
     WRITE(6,411)PRFM
411  FORMAT(1X,'PERFORMANCE (DB)',10F8.3)
531  CONTINUE
     IF(SS.EQ.0.) GO TO 2002
2001 SS=1.
*********************************************************
****           FREQUENCY    HOPPING                *****
****                  SYSTEM                       *****
*********************************************************
*    FH  CODE  SIGNAL  GENERATION
*
     PI=3.1415
     K=1
     Q=1.
     M=1
     N=32
     DO 3001 I=M,N
     FT=FLOAT(M)/64.
     DO 3002 J=1,32
     D(J)=SIN(2*PI*FT*FLOAT(J))
     IF(D(J).GE.0.) T=1.
     IF(D(J).LT.0.) T=-1.
     D(K)=T
     K=K+1
3002 CONTINUE
     IF(M.LT.8.) GO TO 3005
     IF(M.GT.8.) M=M-7.
     GO TO 3006
3005 M=M+8.
     IF(M.EQ.6) GO TO 3004
3006 Q=1.
3001 CONTINUE
3004 XAX=5.
     YAX=13.
     DO 3003 I=1,256
     C(I)=D(I+32)
     E(I)=C(I)
     PE(I)=E(I)
     ZE(I)=C(I)
3003 B(I)=I
     IF(Q.EQ.1.) CALL PLOT(XAX,YAX,128,E,B,'*')
*********************************************************
********************** DATA *****************************
*********************************************************
     DO 3008 I=1,32,2
     E(I)=1.
3008 E(I+1)=-1.
     WRITE(6,3011) (E(I),I=1,8)
3011 FORMAT(//,20X,'DATA  VALUES',///,10(12F6.1,//))
```

```
********************************************
********** MODULATION *************************
*********************************************
   WRITE(6,3011) (C(I),I=1,10)
   J=1
   K=1
   L=32
12 CONTINUE
   DO 3009 I=K,L
   T=E(J)
   IF(T.EQ.1.) GO  TO 3013
   X(I)=-C(I)
   GO TO 3014
13 X(I)=C(I)
14 Q=1.
09 CONTINUE
   IF(L.EQ.256.) GO TO 3015
   J=J+1
   K=K+32
   L=L+32
   GO TO 3012
15 CONTINUE
   WRITE(6,3011) (X(I),I=1,10)
   DO 3025 I=1,256
   P(I)=X(I)
   ZE(I)=X(I)
   XE(I)=P(I)
25 Y(I)=0.
   CALL FFTA(X,Y,8,1)
   DO 3026 I=1,256
26 D(I)=X(I)**2+Y(I)**2
   IF(Q.EQ.1.) CALL LPLOT(60,D,80,B,0.,0.100,200
*********************************************
***************** DEMODULATION ****************
*********************************************
   DO 3016 I=1,256
   Y(I)=P(I)*C(I)
16 X(I)=Y(I)
   IF(Q.EQ.0.)  WRITE(6,3017) (Y(I),I=1,128)
17 FORMAT(//,20X,'DEMODULATED',///,10(10F7.1,//))
   IF(Q.EQ.0.) CALL PLOT(XAX,YAX,128,Y,B,'*')
   J=1
   K=1
   L=256
18 CONTINUE
   DO 3019 I=K,L,32
   V(J)=X(I+10)
   J=J+1
19 CONTINUE
   WRITE(6,3020) (V(I),I=1,4)
20 FORMAT(//,20X,'NEW DATA',///,10(10F10.3,//))
*********************************************
                  J A M M I N G              **
*********************************************
   IFQ=1000
   IDT=0.000375
   ISEED=1000000000
   DO 3103 J=1,128
   T=0.
   DO 3104 I=1,12
   CALL RANSET(ISEED)
   RV=RANF()
   ISEED=ISEED-510
   T=T+RV  .
04 CONTINUE
   FT=T-6.0
   S(J)=FT
03 CONTINUE
   DO 3171 I=1,128
71 E(I)=3*S(I)
   ISEED=500000000
   DO 3105 J=1,128
   HT=0.
   DO 3106 I=1,12
   CALL RANSET(ISEED)
   RV=RANF()
   ISEED=ISEED-200
   HT=HT+RV
06 CONTINUE
   T=HT-6.0
   V(J)=T
   C(J)=3*V(J)
05 CONTINUE
   FT=0.350
```

```
     DO 3031 I=1,128.
     G(I)=S(I)*(COS(2.*PI*FT*FLOAT(I)))-V(I)*
    +(SIN(2*PI*FT*FLOAT(I)))
     Z(I)=G(I)
3031 CONTINUE
     XAX=5.
     YAX=13.
     DO 3032 J=1,128
3032 B(J)=J
     IF(Q.EQ.0.) CALL PLOT(XAX,YAX,128,G,B,'*')
     DO 3107 I=1,128
     R(I)=P(I)+Z(I)
     VE(I)=R(I)
3107 Y(I)=0.
***************** TONE  JAMMING ******************
     DO 3201 I=1,256
     CE(I)=5.*SIN(2*PI*0.1190*I)
     YE(I)=15.*SIN(2*PI*I/1.99)
     BE(I)=XE(I)+YE(I)
     ZE(I)=XE(I)+CE(I)
3201 RE(I)=ZE(I)
     DO 3500 I=1,256
     UE(I)=SIN(2*PI*0.1190*FLOAT(I-1))
     CC(I)=3*XE(I)+UE(I)
3500 GE(I)=CC(I)
********** SPECTRUM OF RECEIVED SIGNAL ************
     CALL FFTA(R,Y,7,1)
     DO 3108 I=1,128
     C(I)=0.
3108 D(I)=R(I)**2+Y(I)**2
     IF(Q.EQ.1.) CALL LPLOT(60,D,80,B,0.,0.100,200.)
     IF(Q.EQ.0.) CALL PLOT(XAX,YAX,128,D,B,'*')
     IF(Q.EQ.0.) GO TO 3177
*****************  SPECTRUM OF INTERFERENCE ******
     CALL FFTA(Z,C,7,1)
     DO 3172 I=1,128
3172 X(I)=Z(I)**2+C(I)**2
     IF(Q.EQ.1.)CALL LPLOT(60,X,80,B,0.,0.100,200.)
3177 Q=1.
**************  DEMODULATION  WITH  INTERFERENCE  **
     DO 3178 I=1,128
     Y(I)=VE(I)*ZE(I)
3178 P(I)=Y(I)
     IF(Q.EQ.0.)  WRITE(6,3179)  (Y(I),I=1,128)
3179 FORMAT(//,20X,'DEMODULATION',///,10(10F7.1,//))
     XA=5.
     YA=13.
     DO 3191 I=1,128
3191 B(I)=I
     IF(Q.EQ.1.)  CALL PLOT (XA,YA,128,Y,B,'*')
     IL=1
     IS=32
     X(1)=0.
     K=1
3181 CONTINUE
     T=0.
     DO 3180 I=IL,IS
3180 T=T+P(I)
     Z(K)=T/32.
     K=K+2
     IL=IL+32
     IS=IS+32
     IF(IS.LT.130.) GO TO 3181
     IL=33
     IS=64
     K=2
3182 CONTINUE
     T=0.
     DO 3183 I=IL,IS
3183 T=T+P(I)
     Z(K)=-(T/32.)
     K=K+2
     IL=IL+64
     IS=IS+64
     IF(IS.LT.130.)GO TO 3182
     IF(Q.EQ.1.) WRITE(6,3020)(Z(K),K=1,4)
```

```
*************************************************
EEEEEEEEE INTERFERENCE  ESTIMATION EEEEEEEEEEEE
*************************************************
  DO 3502 I=1,256
2 B(I)=0.
  CALL FFTA(CC,B,8,1)
  DO 3501 I=1,256
1 UE(I)=CC(I)**2+B(I)**2
  DO 3202 I=1,256
2 Y(I)=0.
  CALL FFTA(RE,Y,8,1)
  DO 3203 I=1,128
3 R(I)=RE(I)**2+Y(I)**2
  CALL LPLOT(60,R,80,B,0.,0.100,200.)
  ND=256
  TS=1./8000.
  IWRT=0
  NP=8
  NM=NP
  CALL AUTOCO(BE,DE)
  WRITE(6,3204) ND
4 FORMAT(1X,'NUMBER OF DATA SAMPLES ',I10)
  WRITE(6,3205) NP
5 FORMAT(1X,'ORDER OF PREDICTOR',I10)
  DO 3206 I=1,256
  E(I)=0.
6 G(I)=0.
  DO 3207 K=1,NP
7 E(K)=DE(K)
  CALL FFTA(E,G,8,1)
  DO 3208 I=1,256
8 P(I)=E(I)**2+G(I)**2
  CALL LPLOT(60,P,80,B,0.,0.100,10.)
********** SUPPRESSION  OF  INTERFERENCE  ******
  NP=256
  DO 3210 I=1,NP
0 S(I)=R(I)*P(I)
  CALL LPLOT(60,S,80,B,0.,0.60,1000.)
  DO 3211 I=1,NP
  X(I)=(E(I)*RE(I))-(G(I)*Y(I))
1 B(I)=(RE(I)*G(I))+(E(I)*Y(I))
  CALL FFTA(X,B,8,-1)
  DO 3212 I=1,NP
  G(I)=X(I)
2 X(I)=X(I)/90.
  WRITE(6,3563) (X(I),I=1,10)
3 FORMAT(//,20X,'INVERSE FFT',//,5(10F8.3,//))
**********DIFFERENCE (IDEAL-JAMMED)************
**********   AS  SPECTRUM  ********************
  DO 3213 I=1,NP
  B(I)=0.
3 D(I)=ABS(ZE(I)-X(I))
  CALL FFTA(D,B,8,1)
  DO 3214 I=1,NP
4 E(I)=D(I)**2+B(I)**2
  IF(Q.EQ.1.) CALL LPLOT(60,E,80,B,0.,0.100,0.1)
************ DEMODULATION  AFTER  FILTERING ****
  J=1
  IL=2
  M=60.
  IS=32
  X(1)=0.
  DO 3215 I=1,NP
  P(I)=G(I)*PE(I)
5 S(I)=P(I)
  WRITE(6,3563) (P(I),I=1,25)
6 CONTINUE
  DO 3217 I=IL,IS
  R(I)=ABS(P(I))/M
7 X(I)=R(I)+X(I-1)
  D(J)=X(IS)/30.
  ISO=IL+62
  X(ISO)=0.
  M=30.
  J=J+2
  IL=IL+63
  IS=IS+64
  IF(IS.LT.260.) GO TO 3216
  J=2
  M=30.
  X(33)=0.
      IL=34
      IS=64
 3219 CONTINUE
      DO 3218 I=IL,IS
      R(I)=-1*(ABS(P(I)))/M
 3218 X(I)=R(I)+X(I-1)
      D(J)=X(IS)/30.
      ISO=IL+62
      X(ISO)=0.
      J=J+2
      IL=IL+63
      IS=IS+64
      IF(IS.LT.260.) GO TO 3219
      WRITE(6,3903) (D(I),I=1,8)
 3903 FORMAT(//,20X,'NEW DATA',//,10(8F10.3,////))
      DO 3571 I=1,128
 3571 CC(I)=R(I)
      IF(Q.EQ.0.)   WRITE(6,3569)(R(I),I=1,128)
 3569 FORMAT(///,20X,'DEMODULATED',//,10(10F10.3,//
      J=1
      IL=1
      IS=32
 3223 CONTINUE
      DO 3221 I=IL,IS
      R(I)=ABS(S(I))/M
      E(I)=R(I)
 3221 R(I)=R(I)+3.
      IL=IL+32
      IS=IS+32
      DO 3222 I=IL,IS
      R(I)=(-1.)*(ABS(S(I)))/M
      E(I)=R(I)
 3222 R(I)=R(I)-3.
      IL=IL+32
      IS=IS+32
      IF(IS.LT.129.) GO TO 3223
      WRITE(6,3573) (E(I),I=1,128)
 3573 FORMAT(//,20X,'NEW VALUES',//,5(10F8.3,////))
      XAX=5.
      YAX=13.
      N=128
      DO 3220 I=1,N
      B(I)=I
 3220 V(I)=R(I)
      CALL PLOT(XAX,YAX,128,V,B,'*')
*********************************************
*****************   SNR 2   *****************
*********************************************
**    AVERAGE  POWER  **
      DO 3231 I=1,128
 3231 X(I)=ZE(I)+CE(I)
      M=128
      N=NM
      DO 3232 I=1,M-1
 3232 X(I)=X(I)**2+X(I-1)
      V(1)=X(M-1)/M
      DO 3233 K=2,N
      V(K)=V(K-1)*(1-A(N,N)**2)
 3233 AP=V(K)
      WRITE(6,3039) AS
 3039 FORMAT(1X,' AS ',10F10.3)
      ERRO=ABS(AP)
      ERRO=10.*ALOG10(ERRO+0.01)
      WRITE(6,3234) ERRO
 3234 FORMAT(1X,'AVERAGE POWER (DB)',10F10.2)
      BA=128.
      DFLF=1000./128.
      DELT=0.00050
      DO 3235 I=1,N
 3235 CC(I)=DF(I)**2+CC(I-1)
      ERROR=BA/(CC(N)+128.*AP*DELF*DELT)
      ERROR=ABS(ERROR)
      SNR=10.*ALOG10(ERROR+0.01)
      WRITE(6,3236) SNR
 3236 FORMAT(1X,'SNR2 (DB)',10F8.3)
      AP=128*(15)**2
```

```
C
C
C
*************************************************************
*************  SNR  1   *****************************
*************************************************************
      ERRO=BA/AP
      ERRO=ABS(ERRO)
      SNR=10.*ALOG10(ERRO+0.01)
      WRITE(6,3237)  SNR
 3237 FORMAT(1X,'SNR1 (DB) ',10F8.3)
*************************************************************
***********  PERFORMANCE  CRITERIA  ***************
*************************************************************
      PRF=ERROR/ERRO
      PRF=10.*ALOG10(PRF+0.01)
      WRITE(6,3328) PRF
 3328 FORMAT(1X,'PERFORMANCE (DB)',10F8.3)
*************************************************************
*************  SNR 2   FOR  FILTERED  SIGNAL *******
*************************************************************
C       NEW  AVERAGE  POWER
      DO 3240 I=1,128
 3240 X(I)=R(I)
      M=128
      N=8
      DO 3241 I=1,M-1
 3241 X(I)=X(I)**2+X(I-1)
      V(1)=X(M-1)/M
      DO 3242 K=2,N
      V(K)=V(K-1)*(1-A(N,N)**2)
 3242 AP=V(K)
      ERRO=ABS(AP)
      ERRO=10.*ALOG10(ERRO+0.01)
      WRITE(6,3243) ERRO
 3243 FORMAT(1X,'LAST POWER (DB)',10F10.2)
      DO 3244 I=1,N
 3244 CC(I)=DE(I)**2+CC(I-1)
      ERROR=BA/(CC(N)+128.*AP*DELF*DELT)
      ERROR=ABS(ERROR)
      SNR=10.*ALOG10(ERROR+0.01)
      WRITE(6,3245) SNR
 3245 FORMAT(1X,'NEW SNR2 (DB) ',10F8.3)
 2002 CONTINUE
      STOP
      END
C
C
```

```
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C*****************************************************************
C                 PLOT  ROUTINE                                   *
C*****************************************************************
      SUBROUTINE PLOT(XAX,YAX,NP,X,Y,DOT)
      DIMENSION X(0:NP),Y(0:NP),XNUM(0:15),YNUM(0:120)
      CHARACTER*1 P(0:480,0:112),DOT,HXY(2)
      CHARACTER*10 BOTLIN(15)
      CHARACTER*5 HLS(2)
      CHARACTER*3 HMM(2)
      DATA SXMAX,SYMAX,XNEAR,YNEAR/11.,45.,0.,0./
      DATA HXY/'X','Y'/,HLS/'LARGE','SMALL'/,HMM/'MAX','MIN'/
      DATA P,BOTLIN/54353*'  ',15*'---------+'/
      IF(XAX.GT.SXMAX) THEN
      WRITE(6,90) HXY(1),HLS(1),SXMAX,HMM(1)
      XAX=SXMAX
      GO TO 40
      END IF
      IF(XAX.LT.3.) THEN
      XAX=3.
      WRITE(6,90) HXY(1),HLS(2),XAX,HMM(2)
      END IF
   40 IF(YAX.GT.SYMAX) THEN
      WRITE(6,90) HXY(2),HLS(1),SYMAX,HMM(1)
      YAX=SYMAX
      GO TO 41
      END IF
      IF(YAX.LT.1.5) THEN
      YAX=1.5
      WRITE(6,90) HXY(2),HLS(2),YAX,HMM(2)
      END IF
   41 XX=XAX*10.-1.
      YY=YAX*10.-1.
      XMIN=X(0)
      YMIN=Y(0)
      XMAX=X(0)
      YMAX=Y(0)
      DO 17 I=1,NP-1
      IF(X(I).LT.XMIN) XMIN=X(I)
      IF(X(I).GT.XMAX) XMAX=X(I)
      IF(Y(I).LT.YMIN) YMIN=Y(I)
      IF(Y(I).GT.YMAX) YMAX=Y(I)
   17 CONTINUE
      IXNO=XX/10+1
      IYNO=YY/6+1
      CALL SCALE(XX,XMAX,XMIN,IXNO,12.,10.,SFX,XNUM,XNEAR,F1)
      CALL SCALE(YY,YMAX,YMIN,IYNO,8.,6.,SFY,YNUM,YNEAR,F2)
      DO 6 I=0,NP-1
      X(I)=X(I)/SFX+F1+XNEAR/SFX+0.5
      Y(I)=Y(I)/SFY+F2+YNEAR/SFY+0.5
      JXP=IFIX(X(I))
      IYP=IFIX(Y(I))
    6 P(IYP,JXP)=DOT
      WRITE(6,94)
      NPRX=IXNO*10
      NPRY=IYNO*6
      WRITE(6,83)(XNUM(I),I=0,IXNO)
      WRITE(6,82)(BOTLIN(I),I=1,IXNO)
      K=0
      L=6
      DO 8 I=0,NPRY
      IF(L.EQ.7) THEN
      L=1
      K=K+1
      END IF
      IF(L.EQ.6) THEN
      WRITE(6,80) YNUM(K),(P(I,J),J=0,NPRX)
      ELSE
      WRITE(6,81) (P(I,J),J=0,NPRX)
      END IF
    8 L=L+1
      RETURN
   94 FORMAT(1H ////1H )
   90 FORMAT(//1X,'*WARNING* SCALE FACTOR GIVEN FOR ',A1,
     &       'AXIS IS TOO     ',A5/11X,'IT IS  ',F4.1,'(',A3,'
     &        )  ASSUMED.')
   80 FORMAT(2X,E8.2,'+',115A1)
   81 FORMAT(10X,'I',115A1)
   82 FORMAT(11X,'+',11A10)
   83 FORMAT(7X,12E10.2)
      END
C
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C
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      SUBROUTINE SCALE(TT,TMAX,TMIN,INO,C1,C2,SFT,TNUM,TNEAR,F)
      DIMENSION TNUM(0:120)
      IF(TMIN.GE.0.) THEN
      SFT=TMAX/TT
      F=0.0
      DO 2 I=0,INO
    2 TNUM(I)=C2*I*SFT
      RETURN
      ELSE
      SFT=(TMAX-TMIN)/(TT-C1)
      F=ABS(TMIN)/SFT+C1
      DO 3 I=0,INO
    3 TNUM(I)=C2*I*SFT-F*SFT
      IF(TMAX.GT.0.0) THEN
      DO 4 I=0,INO
      IF(TNUM(I).GT.0.0) THEN
      TNEAR=TNUM(I-1)
      DO 5 K=0,INO
    5 TNUM(K)=TNUM(K)-TNEAR
      RETURN
      END IF
    4 CONTINUE
      END IF
      END IF
      RETURN
      END
C*****************************************************************
C             F F T    A L G O R I T H M                         *
C                                                                 *
C*****************************************************************
       SUBROUTINE FFTA(X,Y,L,IFFT)
       DIMENSION X(500),Y(500)
       NP=2**L
       IF(IFFT.GT.0) GO TO 80
       DO 81 K=1,NP
81     Y(K)=-Y(K)
80     CONTINUE
       LMX=NP
       SCL=6.283185303/NP
       DO 20 LO=1,L
       LIX=LMX
       LMX=LMX/2
       ARG=0.0
       DO 10 LM=1,LMX
       C=COS(ARG)
       S=SIN(ARG)
       ARG=ARG+SCL
       DO 10 LI=LIX,NP,LIX
       J1=LI-LIX+LM
       J2=J1+LMX
       T1=X(J1)-X(J2)
       T2=Y(J1)-Y(J2)
       X(J1)=X(J1)+X(J2)
       Y(J1)=Y(J1)+Y(J2)
       X(J2)=C*T1+S*T2
       Y(J2)=C*T2+S*T1
10     Y(J2)=C*T2-S*T1
20     SCL=2.0*SCL
       J=1
       NV2=NP/2
       NPM1=NP-1
       DO 50 I=1,NPM1
       IF(I.GE.J)  GO TO 30
       T1=X(J)
       T2=Y(J)
       X(J)=X(I)
       Y(J)=Y(I)
       X(I)=T1
       Y(I)=T2
30     K=NV2
40     IF(K.GE.J) GO TO 50
       J=J-K
       K=K/2
       GO TO 40
50     J=J+K
       IF(IFFT.GT.0) GO TO 66
       DO 70 I=1,NP
       X(I)=X(I)/NP
70     Y(I)=-Y(I)/NP
66     A=1.0
       IF(A.EQ.0.)WRITE(6,67) (X(I),I=1,50)
67     FORMAT(/,20X,'TRANSFORMED VALUES,'//,10(5F12.4,/))
       IF(A.EQ.0.)    WRITE(6,67) (Y(I),I=1,50)
       RETURN
       END
C
C
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```

```
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C
C***********************************************************************
C**                                                                   **
C**        T H E     P L O T     R O U T I N E                        **
C**                                                                   **
C***********************************************************************
      SUBROUTINE LPLOT(NSPC,SPLOT,M,FPLOT,FINT,FFIN,AG)
      DIMENSION ISARET(4),IBUF(300),IS(300),IY(300)
      DIMENSION FPLOT(500),SPLOT(500)
      DATA ISARET/1H ,1H*,1H+,1H./
      WRITE(6,90)
C*********************************************************************
C
C NSPC : NUMBER OF SPECTRAL POINTS
C SPLOT : SPECTRUM ORDINATES
C FPLOT : DATA WAVEFORM
C FINT: STARTING FREQUENCY
C FFIN : ENDING FREQUENCY
C M : NUMBER OF TIME DOMAIN SAMPLES
C ********************************************************************
     PLOTX=900.
     K=1
  11 K=K+1
     IF(K.GT.NSPC) GO TO 12
     IF(SPLOT(K).GT.PLOTX) GO TO 11
     PLOTX=SPLOT(K)
     GO TO 11
  12 DO 13 K=1,NSPC
  13 SPLOT(K)=SPLOT(K)-PLOTX
     DO  10 I=1,NSPC
10   IS(I)=IFIX(10.*ALOG10(SPLOT(I)+AG))
     IBIG=IS(1)
     DO 20 I=2,NSPC
20   IF(IS(I).GT.IBIG)     IBIG=IS(I)
     NORM=IFIX(FLOAT(IBIG-100)/10.)*10
30   DO 40 I=1,12
40   IY(I)=NORM+(I-1)*10
     WRITE(6,100) (IY(JK),JK=1,12)
     DO 1 I=1,NSPC
     IS(I)=IS(I)-NORM
     FRQ=((FFIN-FINT)/FLOAT(NSPC))*FLOAT(I-1)+FINT
     DO 2 J=2,124
     IBUF(J)=ISARET(1)
     IF(I.EQ.1.OR.I.EQ.NSPC) GO TO 3
     IF(J.LE.IS(I).AND.J.LT.111) IBUF(J)=ISARET(2)
     GO TO 2
3    IBUF(J)=ISARET(2)
     DO  50 I1=1,121,10
50   IBUF(I1)=ISARET(3)
2    CONTINUE
     IF(I.GT.M)  GO TO 5
     IMF=IFIX(FPLOT(I)/20.)+118
     IBUF(IMF)=ISARET(4)
5    IBUF(1)=ISARET(2)
     IBUF(111)=ISARET(2)
     IBUF(125)=ISARET(2)
1    WRITE(6,110) FRQ,(IBUF(JK),JK=1,125)
     WRITE(6,100) (IY(JK),JK=1,12)
90   FORMAT(////,50X,"POWER SPECTRUM",//)
100  FORMAT(1X,I5,11I10,"DESIBEL",/)
110  FORMAT(1X,F4.3,125A1)
     RETURN
     END
C
C.
C
C
C
C
C
C
C
```

```
CEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEEE
      SUBROUTINE AUTOCO(RD,DE)
      COMMON /CSIGNL/ND,NP,NP1,TS,JF,PERDEV,IDB,P2
      COMMON/CPROG/PROB,IWRT,NFOU,LNFOU,IRUN,IRUNM,IPOLY,IMPULS
      COMMON/CFEQ/FEQ
      DIMENSION A(28,28),DE(100),SIG(28),RR(40),RD(1000)

C*************************************************************C

C THIS IS THE AUTOCORELATION METHOD OF AR PARAMETER ESTIMATION

C*************************************************************C

C CALCULATE THE SAMPLE AUTOCORRELATION COEFFICIENTS

      NP=8
      ND=256
      DO 10 KK=1,NP
      NK=KK-1
      S=0.
      KL=1
      KH=ND-NK
      DO 20 K=KL,KH
      S=S+RD(K)*RD(K+NK)
20    CONTINUE
      RR(KK)=S/ND
10    CONTINUE
      IF(IWRT.EQ.1) WRITE(6,444)
      IF(IWRT.EQ.0) WRITE(6,1020) (RR(K),K=1,NP)
1020  FORMAT(1X,'AUTOCORRELATION COEFF',10F10.3)
444   FORMAT(1X,///)

C THE LEVINSON DURBIN ALGORITHM

      A(1,1)=-RR(2)/RR(1)
      SIG(1)=(1.-A(1,1)**2)*RR(1)
      DO 30 K=2,NP
      K1=K-1
      A(K,K)=RR(K+1)
      DO 40 L=1,K1
      A(K,K)=A(K,K)+A(K1,L)*RR(K+1-L)
40    CONTINUE
      A(K,K)=-A(K,K)/SIG(K-1)
      SIG(K)=(1.-A(K,K)**2)*SIG(K-1)
      DO 50 L=1,K1
      A(K,L)=A(K-1,L)+A(K,K)*A(K-1,K-L)
50    CONTINUE
30    CONTINUE
      SG=SIG(NP)*TS

C WRITE OUT THE AR POLYNOMIAL

      DO 60 K=1,NP
60    DE(K+1)=A(NP,K)
      DE(1)=1.
      IF(IWRT.EQ.1) WRITE(6,444)
      IF(IWRT.EQ.0) WRITE(6,1010) (DE(K),K=1,NP)
1010  FORMAT(1X,'AR COEFFICIENTS      ',10F10.2)
      IF(IWRT.EQ.1) WRITE(6,444)
      DO 70 K=1,NP
      K1=K+1
      IF(IWRT.EQ.1) WRITE(6,1010) (A(K,L),L=1,K1)
70    CONTINUE
      RETURN
      END
8.09.36.UCLP, AA, P03    ,       1.296KLNS.
```

VI- CONCLUSION

The aim of this thesis to provide a tutorial treatment of the theory of spread-spectrum communications, including a computer aided analysis of system using DS and FH techniques. Spread-spectrum signals used for the transmission of digital information are distinguished by the characteristic that their bandwidth is much greater than the information rate in bits per second. The large redundancy inherent in spread spectrum signals is required to overcome the severe levels of interference that are encountered in the transmission of information over some channels. Two important elements employed in the design of spread-spectrum signals are coding and pseudo-randomness, which makes the signals appear similar to random noise and difficult to demodulate by receivers other than the intended ones.

Although the current applications for spread spectrum continue to be primarily for military communications, there is a growing interest in the use of this technique for mobile radio networks, timing and positioning systems, some specialized applications in satellites, etc. While the use of spread spectrum naturally means that each transmission utilizes a large amount of spectrum, this may be compensated for by the interference reduction capability inherent in the use of spread spectrum techniques, so that a considerable number of users might share the same spectral band.

We have demonstrated that a narrow-band interference embedded in a PN and FH spread-spectrum signal can be estimated and suppressed by means of an adaptive filter that precedes the PN and FH correlator. The estimation of the interference can be accomplished by linear prediction algorithm. We have also investigated the characteristics of the interference suppression filter based on the levinson-Durbin algorithm and derived the performance gain achieved by it. Linear prediction has the advantage of requiring fewer data points in arriving at the estimate. The results indicate that digital whitening filters can effectively suppress jamming signals or environmental interference whose bandwidth is narrow relative to the spread signal bandwidth. It is demonstrated that a spread-spectrum signal with broad-band gaussian noise can not be whitened by using this techniques, because of the power spectrum relations. But it's known that the power of noise is obliged to spread that power overall the frequencies because of the spreading principle of the system that's why a fundamental issue which is about protection against interfering signals is accomplished To make a comparison between FH and DS spread-spectrum systems as antijam techniques, we can say that FH system has some advantages and DS system has other advantages. Using the same whitening techniques it is obtained that FH system has a better performance, which is almost 4 dB., than DS system.

To be sure, there are disadvantages, but these are often outweighed by the advantages. Two prime disadvantages may be listed for spread-spectrum systems: firstly they employ more bandwidth than a conventionally modulated system and secondly they are more complex in that they must include PN signal generators; correlators, whitening filters or other subsystems not necessarily needed in conventional systems.

There are many reasons for spreading the spectrum, and if done properly, a multiplicity of benefits can occur similtaneously. Some of these are

1- Antijamming

2- Anti interference

3- Low-probability of intercept

4- Multiple user random access communications with selective addressing capability.

5- High resolution ranging

6- Accurate universal timing

This thesis does not really demonstrate the applications of spread spectrum systems for which they are useful. The techniques briefly described here will undoubtedly influence future communications system designs. We can say that the communications engineer will become familiar with spread-spectrum techniques, for they are definitely a part of his future. It is not likely that standart broadcast or television signals will ever be spread spectrum, but they could be. We may all be certain, on the other hand, that all other systems we see in the future (especially those having military application) will be carefully considered for possible application of the spread-spectrum techniques.

APPENDIX A

THE LEVINSON-DURBIN ALGORITHM

The Levinson-Durbin algorithm is an order-recursive method for determining the solution to the set of linear equations

$$\Phi_p a_p = \phi_p \tag{A.1}$$

where Φ_p is a pxp Toeplitz matrix, a_p is the vector of predictor coefficients expressed as

$$a'_p = a_{p1} a_{p2} \cdots a_{pp}$$

and ϕ_p is a p-dimensional vector with elements

$$\Phi'_p = \left[\Phi(1)\,\Phi(2)\ \ldots\ \Phi(p)\right]$$

For a first-order (p=1) predictor, we have the solution

$$\phi(0)a_{11} = \phi(1)$$

$$a_{11} = \frac{\phi(1)}{\phi(0)} \tag{A.2}$$

The residual mean square error (MSE) for the first-order-predictor is

$$\begin{aligned}\varepsilon_1 &= \phi(0) - a_{11}\phi(1)\\ &= \phi(0) - a_{11}^2\phi(0)\\ &= \phi(0)(1 - a_{11}^2)\end{aligned} \tag{A.3}$$

In general, we may express the solution for the coefficients of an mth-order predictor in terms of the coefficients of the (m-1)st-order predictor. Thus we express a_m as the sum of two vectors, namely,

$$a_m = \begin{bmatrix} a_{m1} \\ a_{m2} \\ \\ a_{mm} \end{bmatrix} = \begin{bmatrix} a_{m-1} \\ \cdots\cdot \\ 0 \end{bmatrix} + \begin{bmatrix} d_{m-1} \\ \cdots\cdot \\ k_m \end{bmatrix} \tag{A.4}$$

where the vector d_{m-1} and the scalar k_m are to be determined. Also, Φ_m may be expressed as

$$\Phi_m = \left[\begin{array}{c|c} \Phi_{m-1} & \phi^r_{m-1} \\ \hline \phi^{r}_{m-1} & \phi(0) \end{array}\right] \tag{A.5}$$

where ϕ^r_{m-1} is just teh vector ϕ'_{m-1} in reverse order.

Now

$$\Phi_m a_m = \phi_m$$

$$\begin{bmatrix} \Phi_{m-1} & \phi^r_{m-1} \\ \phi^{r'}_{m-1} & \phi(0) \end{bmatrix} \left\{ \begin{bmatrix} a_{m-1} \\ \\ 0 \end{bmatrix} + \begin{bmatrix} d_{m-1} \\ \\ k_m \end{bmatrix} \right\} = \begin{bmatrix} \phi_{m-1} \\ \\ \phi(m) \end{bmatrix} \tag{A.6}$$

From (A.6) we obtain two equations. The first is the matrix equation

$$\Phi_{m-1}a_{m-1} + \Phi_{m-1}d_{m-1} + k_m\phi^r_{m-1} = \phi_{m-1} \tag{A.7}$$

But $\Phi_{m-1}a_{m-1} = \phi_{m-1}$. Hence (A.7) simplifies to

$$\Phi_{m-1}d_{m-1} + k_m\phi^r_{m-1} = 0 \tag{A.8}$$

This equation has the solution

$$d_{m-1} = - k_m\Phi^{-1}_{m-1}\ \phi^r_{m-1} \tag{A.9}$$

But ϕ^r_{m-1} is just ϕ_{m-1} in reverse order. Hence the solution in (A.9) is simply a_{m-1} in reverse order multiplied by

$-k_m$. That is,

$$d_{m-1} = -k_m \begin{bmatrix} a_{m-1m-1} \\ a_{m-1m-2} \\ \\ a_{m-1\;\;1} \end{bmatrix} \qquad \text{(A.10)}$$

The second equation obtained from (A.6) is the scalar equation

$$\Phi^r_{m-1}a_{m-1} + \Phi^r_{m-1}d_{m-1} + \phi(0)k_m = \phi(m) \qquad \text{(A.11)}$$

We eliminate d_{m-1} from (A.11) by use of (A.10). The resulting gives us k_m. That is.

$$k_m = \frac{\phi(m) - \Phi^{r'}_{m-1}a_{m-1}}{\phi(0) - \Phi^r_{m-1}\Phi^{-1}_{m-1}\Phi^r_{m-1}}$$

$$= \frac{\phi(m) - \Phi^{r'}_{m-1}a_{m-1}}{\phi(0) - a'_{m-1}\Phi_{m-1}} \qquad \text{(A.12)}$$

$$= \frac{\phi(m) - \Phi^{r'}_{m-1}a_{m-1}}{\varepsilon_{m-1}}$$

where $_{m-1}$ is the residual MSE given as

$$_{m-1} = \ (0) - a'_{m-1}\ _{m-1} \tag{A.13}$$

By substituting (A.10) for d_{m-1} in (A.4) we obtain the order-recursive relation

$$a_{mk} = a_{m-1m} - k_m a_{m-1m-k} \qquad \begin{matrix} k = 1,2, \ldots ,m - 1 \\ m = 1,2, \ldots , p \end{matrix} \tag{A.14}$$

and

$$A_{mm} = k_m$$

The minimum MSE may also be computed recursively. We have

$$\varepsilon_m = \phi(0) - \sum_{k=1}^{m} a_{mk}\phi(k) \tag{A.15}$$

Using (A.14) in (A.15), we obtain

$$\varepsilon_m = \phi(0) - \sum_{k=1}^{m-1} a_{m-1k}\phi(k) - a_{mm} \ \ \phi(m) - \sum_{k=1}^{m-1} a_{m-1m-k}\phi(k \tag{A.16}$$

But theterm in the brackets in (A.16) is just the numerator of k_m in (A.12). Hence

$$\begin{aligned} \varepsilon_m &= \varepsilon_{m-1} - a_{mm}^2 \, \varepsilon_{m-1} \\ &= \varepsilon_{m-1}(1 - a_{mm}^2) \end{aligned} \tag{A.17}$$

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