DATA TRANSMISSION BY FREQUENCY-HOPPED MULTILEVEL FREQUENCY SHIFT KEYING

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Summary

Spread spectrum systems make use of radio frequency bandwidths which far exceed the minimum bandwidth necessary to transmit the basic message information. These systems are designed to provide satisfactory communication of the message information under difficult transmission conditions.

Frequency-hopped multilevel frequency shift keying (FH-MFSK) is one of the many techniques used in spread spectrum systems. It is a combination of frequency hopping and time hopping. In this system many users share a common frequency band using code division multiplexing. Each user is assigned an address and the message is modulated into the address. The receiver, knowing the address, decodes the received signal and extracts the message. This technique is suggested for digital mobile telephony.

This thesis is concerned with an investigation of the possibility of utilising FH-MFSK for data transmission corrupted by additive white gaussian noise (A.W.G.N.). Work related to FH-MFSK has so far been mostly confined to its validity, and its performance in the presence of A.W.G.N. has not been reported before. An experimental system was therefore constructed which utilised combined hardware and software and operated under the supervision of a microprocessor system. The experimental system was used to develop an error-rate model for the system under investigation.

The performance of FH-MFSK for data transmission was established in the presence of A.W.G.N. and with deleted and delayed sample effects. Its capability for multiuser applications was determined theoretically. The results show that FH-MFSK is a suitable technique for data transmission in the presence of A.W.G.N.

Keywords

FREQUENCY-HOPPING, SPREAD SPECTRUM
CODE MULTIPLEXING, DIGITAL COMMUNICATION, DATA TRANSMISSION

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CHAPTER ONE
INTRODUCTION

INTRODUCTION

One of the most striking of recent developments in technology has been the rapid growth of data communication systems. In such systems the transmitted signal is a waveform which may, for instance, be carried by voltage in a pair of wires, by electromagnetic radiation in the atmosphere or by light in a glass fibre [14]. The essential feature of a data communication system is that the transmitted waveform is itself composed of separate signal elements, often referred to as symbols, digits, bits or pulses, and these signal elements carry the data (information) which it is required to transmit. A signal element is thus a unit of the transmitted waveform. The data carried by an individual signal-element form a symbol or digit, which is often one of the numerals 0, 1, 2, 3, ..., and the data carried by the entire transmitted group of signal elements form a message. The symbols are themselves usually arranged in separate groups, each group forming a character (or word). Most often the transmitted message is alpha-numeric which means it is composed of a sequence of characters each of which is a letter or numeral.

There has been a rapid increase in recent years in the use of large, centrally based computers to process data originated at locations remote from the main computing centre[8]. Also, large amounts of data are being held in data bases organised by mainframe computers [9]. This is particularly the case with public services, both the police and fire services make extensive use of computers at their headquarters. Hospitals and doctor's surgeries are following suit with patient records. To be able to access this information directly from a mobile or fixed link is of major importance to the continued well-being of these services.

At the same time, the use of radio communication systems has grown very fast in the past few decades to provide a broad range of services required by modern civilisation. One consequence of this growth is that portions of the radio spectrum are very crowded with users, and yet there is a requirement to meet the demands of more users. Developing procedures for sharing, allocating and assigning the spectrum to satisfy the need for even more use of the spectrum with minimum interference are major problems[80]. Also multipath fading is a problem associated with fixed and mobile links. Spread spectrum communication systems are designed to provide solutions for the above problems and have other useful capabilities.

One of the important parameters in determining how well a communication system can perform is the signal-to-noise ratio. Many telecommunication systems must operate with low signal-to-noise ratio, such as in high interference environments. Efforts have been made in the past few years to find techniques which would permit radar, ranging and communication systems to operate under such conditions and with a high degree of resolution or accuracy of information transmission. In these efforts , advances in communication theory, coding and the development of reliable and miniaturized digital components have all played a role. All these advances are used in the design of spread spectrum systems (80). The superior performance of spread spectrum systems is not achieved without some disadvantages. The principal disadvantages (22) are increased spectrum occupancy in direct proportion to the signal processing gain, complex spectral analysis is required and the receiver code must be synchronized with the receiver. These disadvantages can be overcome by the use of surface acoustic devices, microprocessors and forms of circuit miniaturization, all of which permit very complex functions to be implemented in a very small space without the need for manual adjustments.

The means by which the spectrum is spread is crucial. Several techniques (10,23,24) exist such as "direct sequence", "frequency hopping", "time hopping" and "chirp". Hybrid combinations of these techniques are frequently used to achieve the combined advantages of the main techniques.

Frequency-hopped multilevel frequency shift keying (FH-MFSK) technique is a combined technique of time hopping and frequency hopping. The FH-MFSK technique was suggested for a multiple access spread spectrum communication system for mobile radio communication service to a large number of users [29,34,44,60,82].

Comprehensive investigations into the feasibility and practical performance of data transmission by FH-MFSK are required, since the reported work about this system has so far been mostly confined to its validity. Moreover, investigations in the presence of additive white gaussian noise (A.W.G.N.) have not been reported before.

To carry out experimental investigations into performances with different parametic values in the presence of A.W.G.N., a flexible experimental system seemed to be of great advantage.

In Chapter 2 spread spectrum systems are reviewed, in particular the frequency hopping system. The basis, advantages and techniques of spread spectrum are discussed and the main features highlighted.

Chapter 3 examines the description, performance and design procedure of an FH-MFSK system. Different address assignments for this system are discussed and compared. The noise and interference considerations are mentioned, followed by a discussion of error-rate analysis.

Different frequency synthesis and spectral analysis techniques which may be used to generate and analyze FH-MFSK signals are discussed in Chapter 4.

Implementation of the FH-MFSK experimental system based on an Intel 8085 microprocessor is explained in Chapter 5. Results obtained from this system are given and discussed in Chapter 6.

Finally, the conclusions and some suggestions for further work are to be found in Chapter 7.

CHAPTER TWO

SPREAD SPECTRUM SYSTEMS

SPREAD SPECTRUM SYSTEMS

2.1 INTRODUCTION

Spread spectrum communication systems in general employ a larger bandwidth than is normally required to transmit information[23,63]. To be classified as a spread spectrum system, the modulated signal bandwidth should be at least 10 to 100 times the information rate, and the information itself should not be a factor in setting the modulated signal bandwidth[23,24]. The band spread is accomplished by means of a code which is independent of the data. A voice signal, for example, can be sent, using amplitude modulation, in bandwidth only twice that of the information itself. Other forms of modulation, such as low deviation FM and single sideband AM, also permit information to be transmitted in a bandwidth comparable to the bandwidth of the information itself. Under the above definitions, it is seen that techniques such as wide-band FM and PCM are not spread spectrum systems, even though they employ more RF bandwidth than another system, since the spread of band is not accomplished by a means independent of the transmitted data[63].

Spread sprectrum communication systems provide secure communication, at the same time coexisting with other communication systems. They also have the ability to counteract the effects of multipaths and interference[10,23,24,63

Various techniques have been used in the past to achieve secure communication. One method is to reduce the signal power to a minimum. A second method is to vary the transmission frequency as well as transmission times, while a third method is to encode the information. These techniques also help to reduce interference. Spread spectrum systems employ all these techniques [23].

Key applications of spread spectrum communication systems are in anti-interference communication, mobile radio, low probability of intercept, direction finder in navigation systems, satellite communications, multiple user random access with selective addressing capability [10,23,24,63].

Spread spectrum communication systems are reviewed in this chapter, with particular emphasis on frequency-hopping systems.

2.2 BASIS OF SPREAD SPECTRUM TECHNOLOGY

Shannon has shown that the channel capacity is related to the bandwidth of the channel by the expression [73]:

$$C = B \log_2(1 + \frac{S}{N}) \tag{2.1}$$

where

C = capacity in bits per second

B = bandwidth in Hz

$\frac{S}{N}$ = signal-to-noise ratio.

The above equation gives the theoretical maximum channel capacity obtainable with error-free transmission, given the signal-to-noise ratio existing in the channel and the bandwidth available to transmit the information. This equation can be rewritten as

$$\frac{C}{B} = 1.44 \log_e(1 + \frac{S}{N}) \qquad (2.2)$$

For small values of $\frac{S}{N}$ ratio (typically $\frac{S}{N} \leqslant 0.01$, as in the case of interfering systems), this equation can be approximated as[24],

$$\frac{C}{B} = 1.44 \frac{S}{N} \qquad (2.3)$$

Hence, the bandwidth required to transmit the given information is given by

$$B = \frac{CN}{1.44S} \qquad (2.4)$$

From this equation one can see that, for a given signal-to-noise ratio, error-free transmission can be achieved by increasing the bandwidth of the channel used to transmit the information.

2.3 PROCESS GAIN AND JAMMING MARGIN

The process gain of a processor is defined as the difference between the output $\frac{S}{N}$ ratio of the processor and the input $\frac{S}{N}$ of the processor. A spread spectrum system develops its process gain in a sequential signal bandwidth spreading and de-spreading operation. The process gain of a spread spectrum processor can be defined as[10,23,24],

$$G_p = \frac{B_{RF}}{R_{info.}} \tag{2.5}$$

where

B_{RF} = bandwidth of the transmitted spread spectrum signal

$R_{info.}$ = data rate in the information baseband channel.

This process gain does not mean[24] that the processor can perform satisfactorily when faced with an interfering signal having a power level larger than the desired signal by the amount of the available process gain. For this reason, jamming margin, (M_j), is used to express the capability of the spread spectrum system under interference conditions[24,75,81]. Jamming margin takes into account the requirement for a useful system signal-to-noise ratio and allows for internal losses; that is,

$$M_j = G_p - \left[L_{sys} + \frac{S}{N}\right] \tag{2.6}$$

where

L_{sys} = implementation losses

$\frac{S}{N}$ = signal-to-noise ratio

This establishes that if S/N is the minimum signal-to-noise ratio needed to support a given bit-error-rate, then M_j is the maximum tolerable jamming margin for the same internal implementation losses.

2.4 ADVANTAGES OF SPREAD SPECTRUM TECHNIQUES

Advantages of spread spectrum techniques are:-

(1) Interference rejection.

(2) Code division multiplexing.

(3) Multipath tolerance capability.

(4) Low density power spectra for signal hiding.

(5) Message screening from eavesdroppers.

(6) Ranging or time delay measurement.

(1) Interference Rejection

Spread spectrum systems provide an interference rejection that cannot be matched in any other way. Both deliberate and unintentional interference are rejected by a spread spectrum receiver up to a maximum which is equal to the jamming margin[10,81].

Equation (2.6) shows how the spread spectrum achieves interference rejection. The process gain is the quantitative measure of this capability.

(2) Code Division Multiplexing

When codes are properly chosen so that cross-correlation is low, minimum interference occurs between users[23,24]. Any user assigned an unique code has access only to the transmitter transmitting this code. Thus more than one signal can be unambiguously transmitted in the same frequency band simultaneously; code division multiplexing is, thus, easily implemented in a spread spectrum system. Code division mutliplexing will be discussed in greater detail in the next chapter.

(3) Multipath Tolerance Capability

The primary characteristic of radio channels which makes their use difficult for the transmission of data is their susceptibility to fading[9]. That is, the envelope of the radio signal present at the receiver is subject to large scale amplitude variations. Fading has two primary causes. These are shadowing and multipath interference. Shadowing is some large body being interposed between the transmitter and the receiver, causing a reduction in the signal amplitude at the receiver.

Multipath interference occurs due to different propagation paths taken by radio waves (either ground wave, tropospheric or sky wave). Consequently, the signal that is presented to the receiver is a summation of many copies of the original transmitted signal. Each copy will have a different amplitude and phase depending on the distance it had travelled and any extra attenuation it has experienced. The resultant signal produced by the summation will vary with time in both amplitude and phase. The situation is further complicated if either the transmitter, the receiver, or the reflecting body is moving. The individual components will acquire a Doppler shift proportional to the relative motion of the receiver, transmitter and the reflector, if one is present.

One method used to combat multipath is the use of large time - bandwidth coded waveforms [65]. This technique is used in spread spectrum systems and is discussed in the next chapter.

(4) Low Density Power Spectra for Signal Hiding

Because of the wideband signal generated by code modulation, the power transmitted is low in any narrow frequency band. For this reason it is very easy to hide the signal.

(5) Message Screening from Eavesdroppers

Because the signals are coded, an eavesdropper cannot casually listen without decoding the message.

(6) Ranging or Time Delay Measurement

Spread spectrum is useful in ranging and velocity measurement for navigation applications[10]. The range and velocity measurements are obtained by separately tracking the phase of the code and/or the carrier frequency in separate loops used in the modulation process.

2.5 SPREAD SPECTRUM TECHNIQUES

The following techniques are used in spread spectrum systems:

1. Direct sequence
2. Frequency hopping
3. Pulsed-frequency modulation (chirp)
4. Time hopping
5. Hybrid forms.

All the techniques mentioned above require a pseudorandom noise (PN) code generator for bandwidth spreading[10]. A PN code generator produces a binary sequence which is apparently random but can be reproduced deterministically by the intended recipients. It is assumed that the message is in the form of binary data, which will be appropriately combined with PN codes for transmission. Fig 2.1 shows an example of a maximal length PN code generator which consists of a shift register with modulo 2 feedback.

2.5.1 Direct Sequence (DS)

Direct sequence spread spectrum generates a wideband noise-like signal from a PN code generator[10]. This technique is the most popular and widely used for navigation systems[10,24]. Generally, a direct sequence modulation approach is employed whereby the PN code directly balance modulates the carrier, as shown in Fig. 2.2.
Fig. 2.2, illustrates a typical direct sequence communication link. The basic form of direct sequence is that produced by a simple, biphase-modulated carrier. Normally the carrier in a spread spectrum is modulated by baseband information. The baseband information is digitised and added to the code sequence. In this chapter, it is assumed that the RF carrier has been modulated before code modulation since this assumption simplifies the block diagram for the modulation-demodulation process. The multiplication process of the RF carrier by the PN code will spread the message bandwidth to twice the code rate. (The spectrum of a rectangular pulse is twice its rate for the main lobe[80].)

After amplification, a recieved signal is correlated with a reference signal with the same PN code. Assuming that the transmitter's code and receiver's code are synchronous , the transmitted carrier inversions are removed at the receiver and the original carrier restored. This narrow-band restored carrier can then flow through a bandpass filter designed to pass only the baseband modulated carrier.

However, any receiving 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 PN code[24,80].

Because an unwanted input signal is mapped into a bandwidth that is at least as wide as the receiver's PN code signal, the bandpass filter can reject almost all the power of an undesired signal. This is the mechanism by which process gain is realised in a direct sequence system[24]. The process gain of the DS system is given by equation (2.5). That is, the receiver transforms synchronous input signals from the code modulated bandwidth to the baseband modulated bandwidth. At the same time unwanted input signals are spread at least over the code modulated bandwidth.

2.5.2 Frequency Hopping (FH)

Frequency hopping modulation is more accurately termed "multiple frequency, code selected frequency shift keying". It is nothing more than simple frequency shift keying (FSK) except that the set of frequency choices is greatly expanded[24]. Simple FSK most often uses only two frequencies; for example, f_1 is sent to signify a mark f_2 to signify a space. Frequency hoppers, on the other hand, often have thousands of frequencies available [10,24].

The main differences betweeh FH and DS techniques are in the way the transmitted spectrum is generated and in the way interference is rejected. Fig. 2.3 is a simplified block diagram of the transmission system. The frequency spectrum of this frequency hopper is shown in Fig. 2.4. Ideally, the instantaneous frequency hopper output is a single frequency. FH systems also use PN code generators at both the transmitter and receiver that are capable of producing identical codes with proper synchronization.

In the FH technique, the PN code sequence is used to switch the carrier frequency instead of directly modulating the carrier. When the frequency synthesizer is switched with a synchronized replica of the transmitted code, the frequency hops on the received signal will be removed, leaving the original modulated signal, which is then demodulated in a conventional manner.

The bandwidth over which the energy is spread is essentially independent of the code clock rate, and can be chosen by combination of the number and size of frequency hops [80]. The bandwidth of an idealised power spectrum of a FH signal is given by,

$$B_{RF} = (2^{ns}-1)\ \Delta f \qquad (2.7)$$

where

ns = Number of shift register stages used in PN code generator

Δf = Frequency separation between discrete frequencies.

Δf must be at least as wide as the information bandwidth (B_{inf}). As in the DS technique, the processing gain for the FH technique is given by,

$$G_p = \frac{B_{RF}}{B_{inf}}$$

or

$$G_p = \frac{(2^{ns}-1)\Delta f}{\Delta f}$$

$$= 2^{ns}-1 \qquad (2.8)$$

= The number of channels used.

However, the DS technique needs a high code rate (that is much higher than the message information rate) to produce the necessary spectrum spreading as mentioned in the previous section. On the other hand, FH techniques do not necessarily require such high code rates as shown in equation (2.8).

Interference from unwanted signals can occur under the following conditions [80]. First, the receiver code shifts by the same amount an unwanted signal is shifted. To minimise this interference, low cross-correlation codes have to be used. Secondly, such interference can occur if the receiving channel is nonlinear and intermodulation between interfering signals produces frequencies which, when mixed with the current value of frequency synthesizer, result in frequencies commensurate with the receiver passband. A fixed frequency signal, of course, causes interference only when the code produces a frequency which, when combined with the fixed frequency signal produces a frequency capable of passing through the receiver passband. At all other code states the fixed frequency signal is translated to a different frequency which would not coincide with that of the receiver passband.

Frequency hopping is classified, according to frequency hop rate, as either slow frequency hopping or fast frequency hopping[16]. With slow frequency hopping, the carrier frequency remains constant for time periods far in excess of the propagation time. Several milliseconds is a reasonable dwelling time in this case. This usually allows many data bits to be transmitted at each frequency, and the resulting transmitter and receiver equipment is simpler and less expensive than that for a faster frequency hop. The disadvantage of slow frequency hop is that it cannot overcome the problem of deliberate interference because the interferer has time to search and detect the transmitted frequency. On the other hand, slow frequency hop can be used in digital multiplexing, using code division multiplexing techniques.

Fast frequency hopping involves very rapid retuning of the signal and very short periods at each frequency. Generally, fast frequency hop is used to overcome deliberate interference.

The required hopping rate is determined by a number of parameters such as the type of information being sent, its rate and the amount of redundancy used[24]. Information in a FH system usually consists of some form of digitised signal, whether the information is a digitised analogue signal or data. A frequency hopping system must have a large number of frequencies usable on demand. The number required is dependent on system error-rate.

Some form of redundancy is used that allows bit decisions to be based on more than one frequency. More than one frequency sent per bit will decrease the error rate, but the hopping rate required and the RF bandwidth will increase in direct proportion. For this reason, if the synthesizer frequency bandwidth and the number of frequencies are limited, then some trade off must be made between sending a larger number of frequencies per bit and reducing the number of frequencies available.

Secondly, the required hopping rate is also determined by the signals received due to multipath or, even worse, deliberate interference. FH systems can combat the multipath signal by using M-ary signals. The threat from deliberate interference can be overcome by using a high hopping rate which allows the FH system to skip to another frequency before the interferer can respond to the previous one.

In M-ary frequency hopping systems a block of ($\log_2 M$) bits are sent by transmitting one of a set of M frequencies[24]. The reason for using an M-ary transmission format is to reduce the hopping rate. This, however, does not reduce errors because when such transmissions are used, an interfering signal needs only to hit any one of the (M-1) channels not intended for transmission to cause a block error.

2.5.3 Pulsed Frequency Modulation (chirp)

The linear frequency modulation technique of spectrum spreading was developed to improve the resolution of the radar receiver when processing return signals from adjacent targets and thus making possible a significant power reduction[24,80]. This type of spread spectrum system does not necessarily employ coding but does use a wide bandwidth.

Chirp transmissions are characterized by pulsed RF signals whose frequency varies in some known way during each pulse period as shown in Fig. 2.5a. A simple way to generate chirp signals is by using a long transmission pulse, of constant amplitude A_c and duration ΔT_c, to suitably control a linear voltage-sweep oscillator. The frequency of the transmitted pulse is linearly increased from f_1 to f_2 during the pulse. Chirp signal generation is discussed in greater detail in Chapter Four.

The receiver used for chirp signals is a matched filter, matched to the angular rate of change of the transmitted frequency-swept signal. The matched filter is a storage and summing device that accumulates the energy received over an interval. It assembles and releases the energy in one coherent burst. We can visualise the filter operation by a transmitter that sends a series of signals over a time period, using linear frequency modulation as mentioned above[24]. Each increment in frequency denotes a new signal. A received signal is stored until the entire ensemble of frequencies in one period arrives and the summed power at all frequencies is output at one time.

The transmitted chirp signal has two parameters, the amount of frequency sweep $f_2-f_1 = \Delta f_c$ and the amount of time ΔT_c used to sweep Δf_c[24]. The transmitted signal energy contained in the original long pulse is compressed into a shorter pulse of duration approximately $\frac{1}{\Delta f_c}$, the instantaneous peak power of the compressed pulse is increased by the factor Δf_c over that of the long transmitted pulse ΔT_c. For frequency modulated signal spread spectrum then, a processing gain, $G_p = \Delta f_c . \Delta T_c$, is attained[80].

The chirp system has been used in data transmission communication systems[46].

2.5.4 Time Hopping (TH)

Time hopping systems control their transmission time with a code sequence in the same way as frequency hoppers control their frequency. TH can be viewed as pulse modulation under code sequence control. Fig. 2.6 shows a block diagram for a time-hopping system. From this block diagram one can see the simplicity of the TH modulator. Any pulse-modulated signal source that is controlled by a code sequence is suitable to generate TH waveforms. The demodulator includes a code sequence generator similar to that used in the previous spread spectrum system, followed by a converter which changes the pulse-modulated wave to a train of weighted impulses (or the realisable equivalent)[11,24].

Time-hopping systems operate with a high transmitter peak power. The processing gain of TH is similar to the processing gain of the FH and is given by equation (2.8), and is equal to the number of available slots in a repetition frame[10].

This type of modulation may be used to reduce interference between systems in time-division multiplexing,[10,24]. Stringent timing requirements must be placed on the overall system to ensure minimum overlap between transmitters. Simple TH modulation offers little in the way of interference rejection.

TH modulation in combination with frequency hopping has found major application in systems where large numbers of widely spacially separated users with (possibly) differing transmitter powers operate simultaneously on a single link[10,24]. Such systems tend to employ simple coding, primarily for addressing, rather than to spread the spectrum. The general tendency is to design for the equivalent of a wireless telephone switching system in which random access and discrete address are the prime operational goals. For such uses, time frequency hopping is well adapted and offers a viable solution to the near-far problem. Thus, using time-frequency hopping,transmitters in many links can be programmed to transmit on different frequencies as well as at different times. Time-frequency hopping systems will be discussed in greater detail in the next chapter.

2.5.5 Hybrid Forms

Hybrids of the basic spread spectrum modulation techniques are useful in many applications[10,24]. Hybrid spread spectrum systems are derived by combining two or more of the four basic modulation techniques. These often represent the only viable approach to implementing very wide-band and/or very high process gain systems. The reason for this is that no frequency synthesizer can be expected to produce the number of frequencies that is required to cover the entire spread signal bandwidth of a system. Likewise, no code generator can be expected to run at a rate that provides the entire spectrum spreading. The most commonly used combinations for hybrid spread spectrum systems are:-

1. Direct sequence/Frequency hopping (DS/FH)

2. Direct sequence/Time hopping (DS/TH)

3. Frequency hopping/Time hopping (FH/TH)

4. Direct sequence/Frequency hopping/Time hopping (DS/FH/TH).

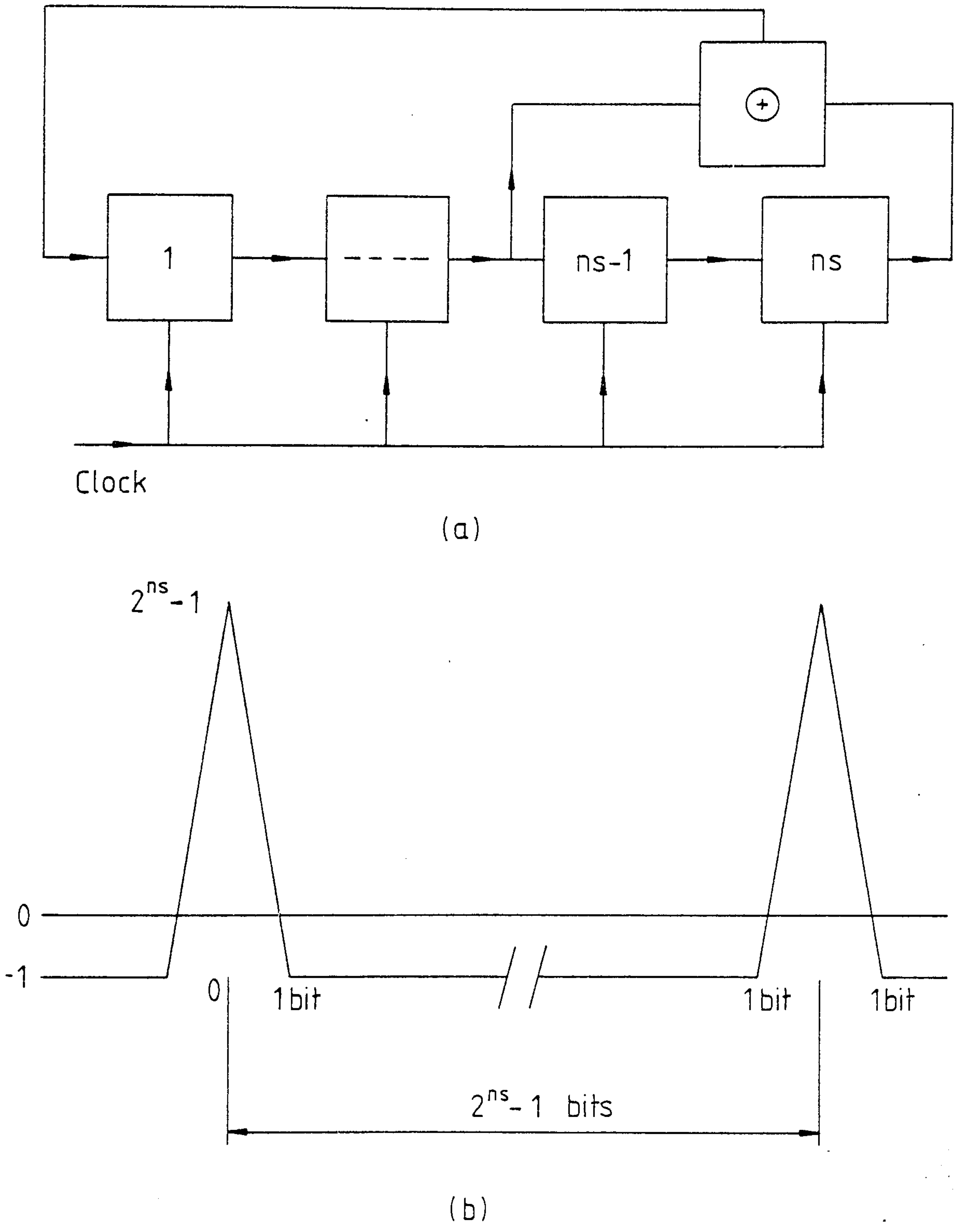


Fig 2.1 PN code generator : (a) typical maximal PN code generator : (b) autocorrelation of maximal code.

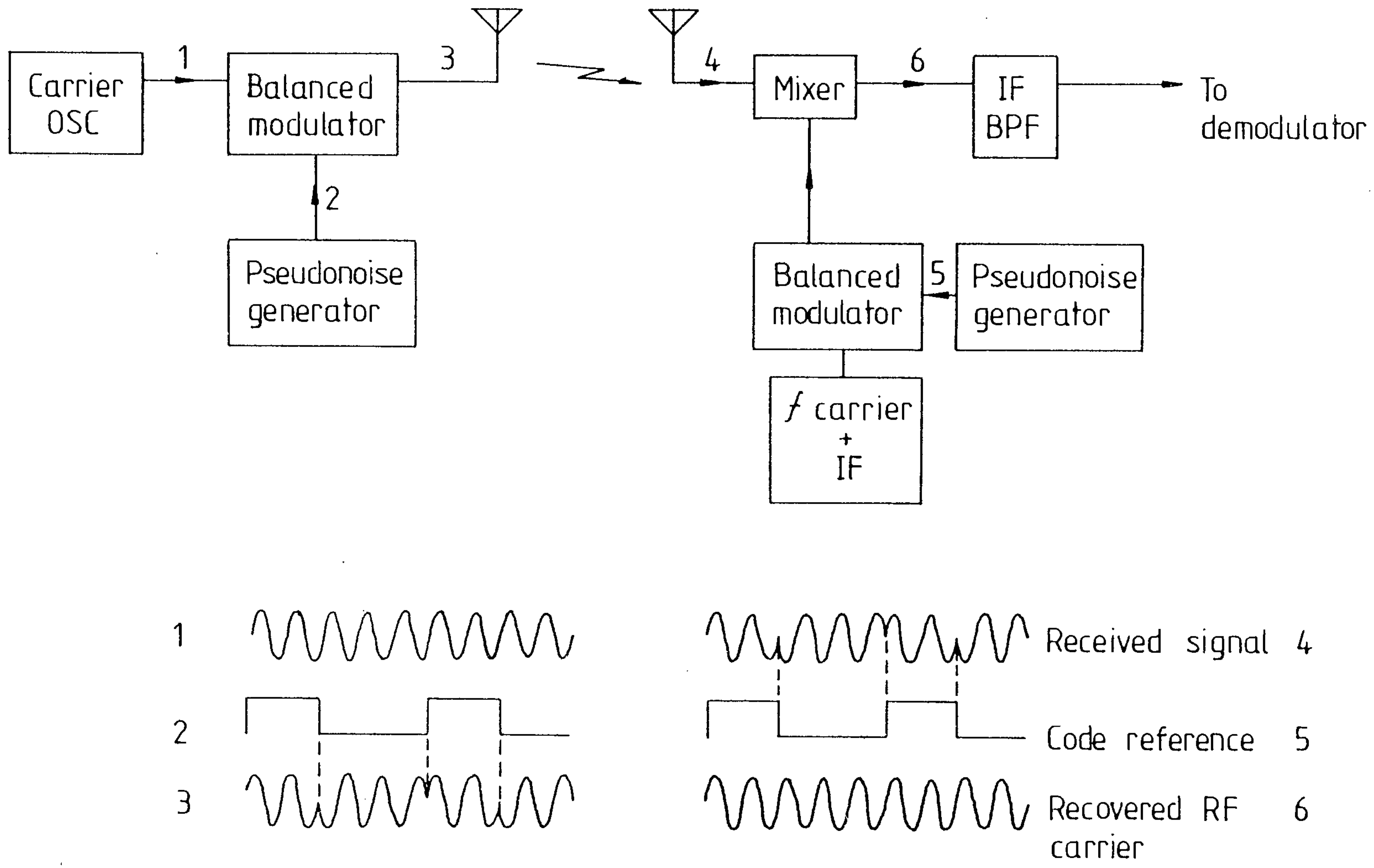


Fig 2.2 Overall direct sequence system showing waveforms.

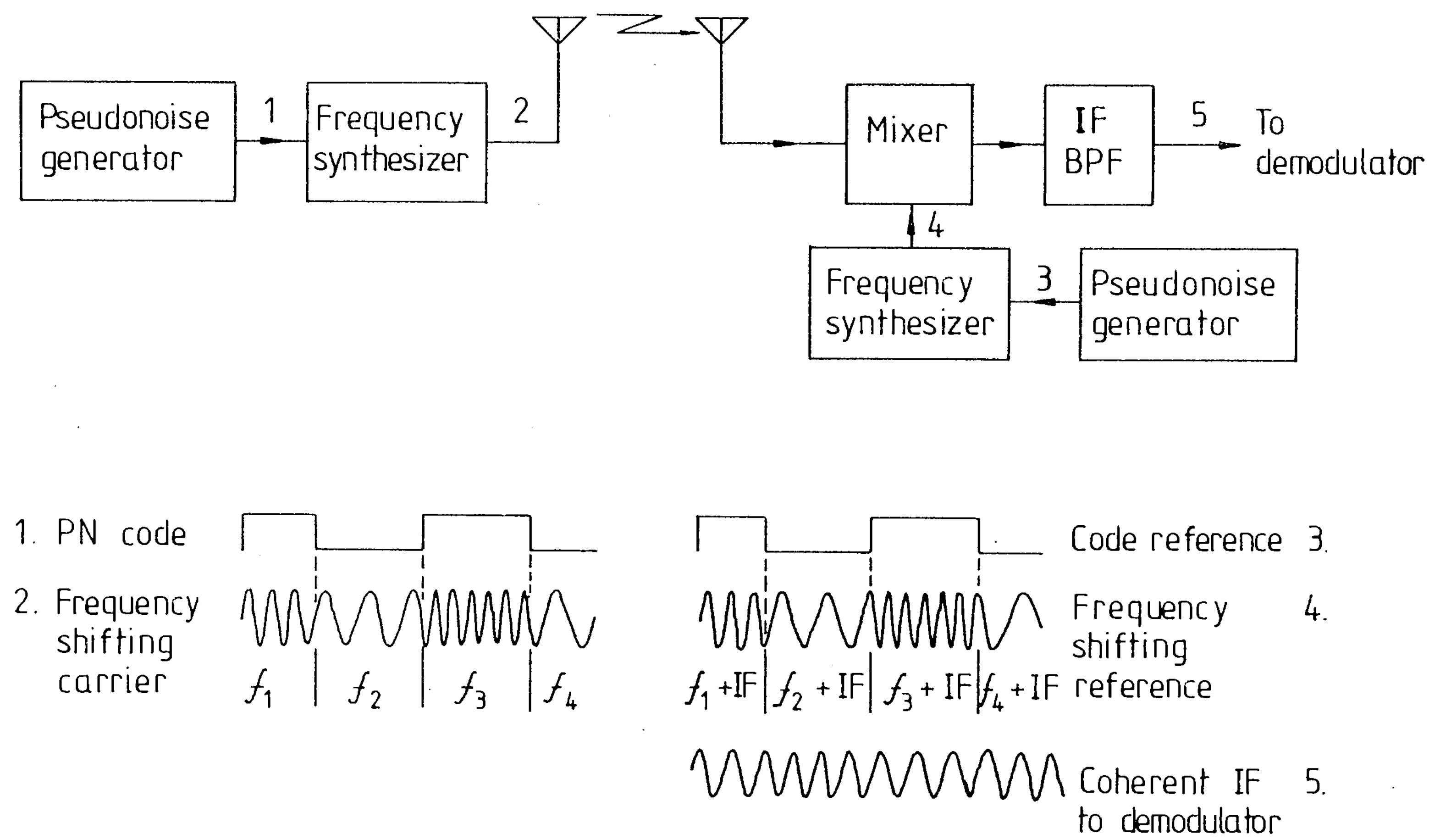


Fig 2.3 Basic frequency hopping system with waveforms.

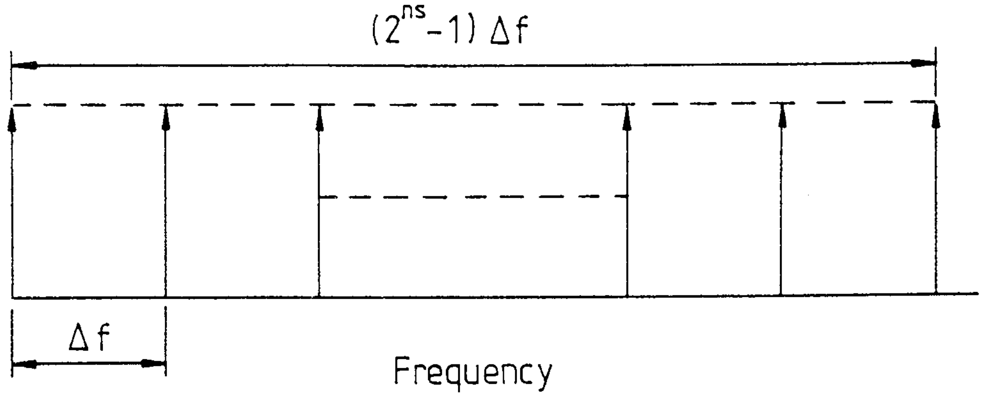


Fig 2.4 Ideal frequency - hopping spread spectrum

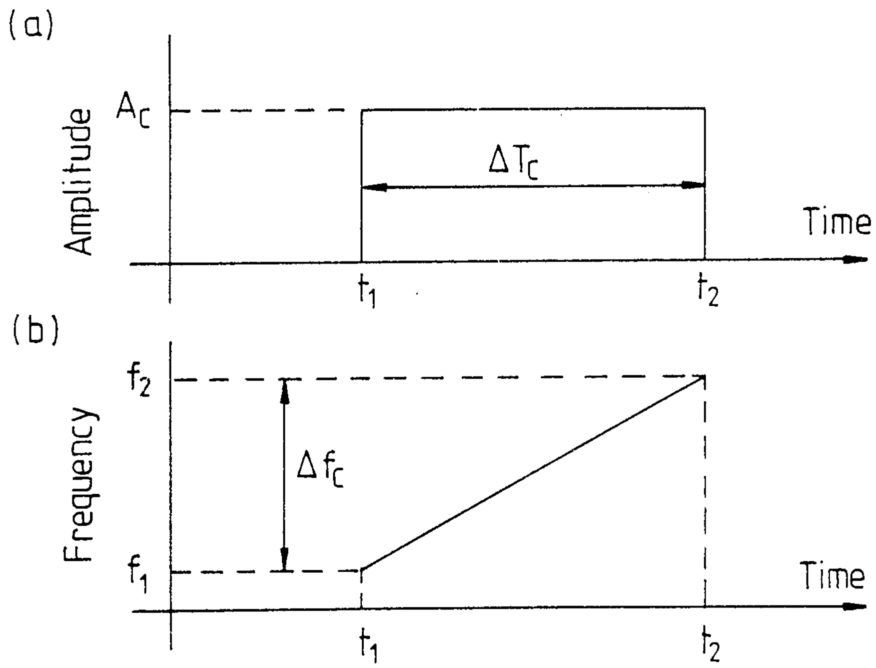
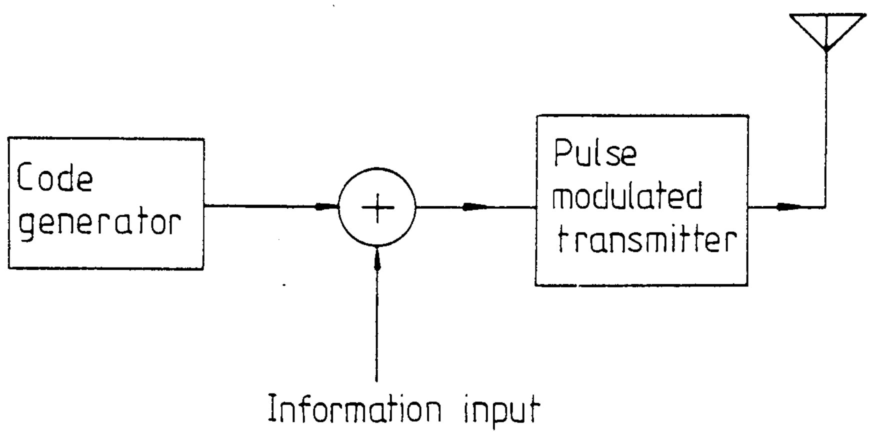
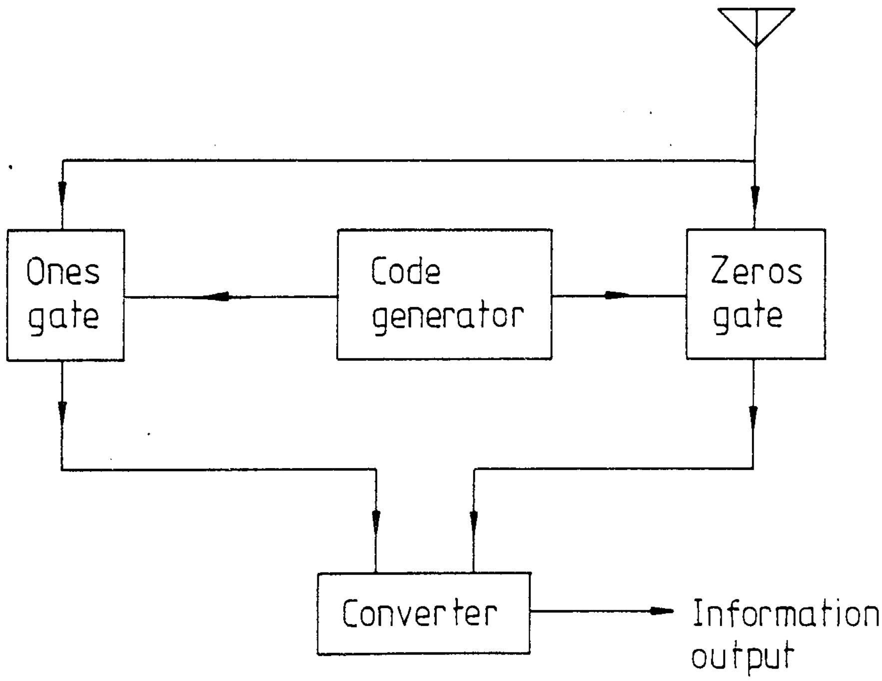


Fig 2.5 Linear FM spread spectrum :
(a) Transmitted pulse of amplitude A_c and duration ΔT_c
(b) Linear frequency variation over bandwidth Δf_c of transmitted pulse.



(a)



(b)

Fig 2.6 Simple time hopping system :
(a) Transmitter (b) Receiver

CHAPTER THREE

FREQUENCY-HOPPED MULTILEVEL FREQUENCY SHIFT KEYING SYSTEM

FREQUENCY-HOPPED MULTILEVEL FREQUENCY SHIFT KEYING SYSTEM

3.1 INTRODUCTION

In this chapter the FH-MSK system is discussed. Multiplexing and its techniques are briefly mentioned in Section 3.2. In Section 3.3 a description of the FH-MSK system is given and its performance is discussed. This is followed by the design procedure in Section 3.4. Different address assignments and the algebra of finite fields are briefly mentioned in Section 3.5. Lastly, in Section 3.6, noise, interference and error-rate analysis are discussed.

3.2 Multiplexing Techniques

The term multiplexing has long been used in telecommunications to refer to any technique which permits more than one independent signal to share one physical facility[11,57,69,38]. There are three well known basic multiplexing techniques:

1. Frequency division
2. Time division
3. Random access discrete address.

3.2.1 Frequency Division Multiplexing (FDM)

In FDM, the frequency bandwidth available is divided up into small bands. Each signal occupies one of these bands as shown in Fig. 3.1. The individual baseband signals $s_1(t)$, $s_2(t)$,....,$s_\lambda(t)$, each bandlimited to f_{bs}, are applied to individual modulators. Each input signal individually modulates the carrier waveform of frequency f_{c_1}, f_{c_2},...,f_{c_λ} . Most importantly, the carrier frequencies are selected so that the ranges of the modulator-output signals do not overlap. To reduce this coupling, the modulated message spectra are spaced out in frequency by guard bands into which the filter transition regions can be fitted. The combined output of all the modulators is applied to a common communications channel[11,69].

At the receiving end, the composite signal is applied to each of a group of bandpass filters. They are then applied to individual demodulators which extract the baseband signals from the carrier.

The major practical problem of FDM is crosstalk, the unwanted coupling of one message into another.

3.2.2 Time Division Multiplexing (TDM)

In TDM, the available time is divided into small slots and each of these slots is occupied by one message to be sent. The essentials of TDM are quite simple as shown in Fig. 3.2. The several input signals, all bandlimited in f_{bs}, are sequentially sampled at the transmitter by a rotary switch. The switch makes one complete revolution in $T_s \leqslant \frac{1}{2f_{bs}}$ seconds, extracting one sample from each input. Hence, the rotary switch output is a pulse amplitude modulation containing the individual message samples periodically interlaced in time. If there are λ inputs, the pulse to pulse spacing is $\frac{T_s}{\lambda}$, while the spacing between successive samples from any one input is of course T_s. At the receiver there is a similar rotary switch, called the demodulator, which distributes the pulses to a bank of low pass filters, which in turn reconstruct the original messages.

A critical aspect of TDM is synchronization. However, the implementation of TDM is somewhat simpler than FDM[11]. Actual pulse shapes have decaying tails, and tend to overlap. This can be reduced by providing guard times between pulses.

3.2.3 Random Access-Discrete Address (RADA) or Code Division Multiple Access

In RADA, each user has access to the entire system bandwidth,[24, 52, 53] thus avoiding the limited channel problem when more potential users exist than there are channels available. RADA is an efficient technique of extending the capability of multiple access systems[24,38,52,53,82], many users accessing a common channel simultaneously without co-ordination. Thus, the RADA technique could be used in satellite communications[57,82] and also for emergency wideband channel mobile users[52]. This technique will be discussed in more detail in the following sections.

3.2.3.1 RADA System Description

The RADA system has many transmitters that radiate signals more or less simultaneously, and each receiver must pick out of this mixture of signals those signals which are intended for it. Thus we use the term random access. In order to accomplish this, each transmitter addresses each and every signal element which it transmits with the code of the intended receiver so that the receiver may select the coded signal addressed to it. Hence, the term discrete address[38].

Also, a system which separates the signals by coding is called a code division multiplexing (CDM) system[30,33].

A typical RADA system[30,33] will be a time sharing system in which each user transmits short RF pulses occupying a small portion of the time (this small portion of time is called a time slot). The transmissions are not synchronized, so the pulses from different users may coincide, causing a certain amount of self-interference that is characteristic of the RADA system. Self-interference is the interference due to other users of the same RADA system.

The available frequency spectrum of RADA, instead of being divided into many narrowband channels and assigned to each user, is made available as a common single wideband channel to all users. The wideband channel is subdivided into a few subchannels called frequency slots as shown in Fig. 3.3. Each user of RADA transmits short pulses in these frequency slots according to its address. Thus, a time-frequency (T-F) matrix is formed. Each entry of the T-F matrix is called a chip. This system will be discussed in Section 3.3.

3.2.3.2 Advantages of RADA System

(1) The most important advantage of RADA is that each user can contact any other user without delay by simply setting his transmitter to the address of the other user. This can be done at any time without correlation with others or checking to see if the channel is free and without any central control[53,83].

(2) RADA systems offer one viable solution to the near-far problem[24,53], as mentioned in Section 2.5.4. This is achieved by programming all transmitters to transmit on different frequencies as well as at different times. This problem occurs in those systems where a large number of users at widely variable distance or transmitted power are to operate simultaneously on a single link.

Consider two receiving and transmitting links whose receivers and transmitters are spaced as shown in Fig. 3.4 and in which the receiver for each link is positioned so that the other link transmitter causes interference. The problem in this system lies in the difference in distance between a receiver and its desired transmitter and the nearby transmitter. The interference rejection capacity of spread spectrum systems aids in such a situation[24] but unfortunately this is not enough under most conditions to overcome the disparity seen in nearby and remote signals. The near-far problem can be overcome by using a RADA technique.

(3) Better frequency spectrum utilisation,is possible in mobile emergency communication systems[52,53]. In such systems, delays cannot be tolerated. If narrowband channels are used, they can be utilised only for a small percentage of the time so as to permit free access to a channel without delay in case of emergency. For narrowband channels, there must be sufficient guard band between channels in order that the strong undesired signals on the nearby channels do not interfere with the reception. Also in VHF and UHF, frequency stability is a problem requiring additional guard bands. In practical systems, 50-100 kHz spacing between channels is generally provided. If one considers several such channels, which under emergency consideration can only be used a small fraction of the time, it turns out that a RADA system will provide much better service with no delays or loss of frequency spectrum.

(4) Another advantage of RADA is its anti-interfering quality, which is common to wideband systems[10,24,53].

3.2.3.3 RADA modulation methods

The transmitted RADA waveform has to carry the address in addition to the modulation that is used in the other commonly used communication systems[53,72]. The addressing or multiple access modulation is designed to accomplish two main purposes:-

1. It must make possible separate reception of each transmitted message by another user's receiver.

2. It must interfere as little as possible with the transmission of each user.

As mentioned previously, a RADA system uses frequency and time slots to form a T-F matrix. The T-F matrix technique requires that the modulation be in the form of a pulse train (modulation pulses), because each of the modulation pulses is transmitted as several much shorter RF pulses, selected from the T-F matrix. The possible modulation schemes that are compatible with T-F matrix addressing can be divided into two classes[53,72], analogue pulse modulation and digital pulse modulation.

(1) Analogue Pulse Modulation

If a message is described by its sample values, it can be transmitted via analogue pulse modulation, wherein the sample values directly modulate a periodic pulse train with one pulse for each sample. There are four types of pulse modulation[11,69], usually designated as pulse-amplitude modulation (PAM), pulse frequency modulation (PFM), pulse width modulation (PWM) and pulse position modulation (PPM). The modulated pulse parameter - amplitude, frequency, width or position - is varied in direct proportion to the sample values as shown in Fig. 3.5. PAM is not appropriate for RADA applications because it is very sensitive to RADA self-interference[53]. Even a relatively small interfering pulse, when added to or subtracted from the amplitude modulated pulse, will cause a considerable amount of interference to the detected output.

PWM is not compatible with the T-F address matrix because the transmitted pulse width variation is sensitive to the path of transmission. PPM and PFM are often considered for RADA systems.

The advantages of the usable analogue pulse modulation schemes is that a relatively small number of pulses (only one modulation pulse per voice sample) are transmitted. Its disadvantage is its relatively high sensitivity to interfering pulses. Another disadvantage of analogue modulation shows up when repeaters are used. Not only is the design of pulse repeater difficult for this application but also each repeater adds noise and distortion to the modulation[53].

(2) Digital Modulation

In digital modulation the message is represented by a coded group of digital signals before transmission. Information is expressed, therefore, as a binary sequence of logical 1 and 0 pulses. This could be derived either from conversion of analogue signals or from any other source of digital information [11,56,69]. When transmitting binary data, it is not necessary to preserve the signal waveform to ensure perfect message reproduction. It is sufficient to prevent each binary signalling waveform from becoming so distorted that it can be mistaken for the other. The use of binary signalling means that the receiver implements the simplest possible decision, that is, the decision concerning which of two possible signalling waveforms is received. Digital modulation also provides a uniform and reliable output. Fixed signal-to-noise ratio at the system output is obtained. This is because it is a sharp transition threshold system and the only noise present is the so called quantizing noise, which results from converting the analogue signal to digital form. There are two binary digital modulation systems compatible with the T-F matrix; pulse code modulation (PCM) and delta modulation (DM). DM has several advantages over PCM for RADA applications. One is the simplicity of the DM modulators and demodulators, and another is the relative insensitivity to transmission error caused by self-interference, which always exists in RADA systems.

In DM, each error in the pulse train will cause only a two-quantizing level error at the receiver output. On the other hand, an error in PCM may cause an output error of several quantizing levels, depending on its position in the pulse group.

An additional advantage of all digital systems for RADA applications (53) is that simple pulse-by-pulse repeaters can be used and the messages can be repeated several times over several repeaters without accumulating distortion and noise.

Thus, from the above discussion, the possibility can be seen of data communication using RADA techniques to gain the advantages of RADA which have been discussed in Section 3.2.3.2.

3.3 FH-MFSK SYSTEM DESCRIPTION AND PERFORMANCE

The FH-MFSK system is an example of a RADA system, using a T-F matrix, as mentioned in the previous section.

An FH-MFSK system with a majority logic receiver is investigated here for possible application in data transmission. This system uses an MFSK signal to modulate FH carriers. It also uses the same time-frequency matrix in the transmission and detection process.

In FH-MFSK systems (29,34,44,60), each user m can be thought of as transmitting every T seconds K message bits X_{mt} where X_{mt} in the set $(0,1,2,...,Q-1)$, where Q is the number of frequencies available. Figure 3.6 shows the operation of the FH-MSFK transmitter. Assume for the moment only one user m, and the modulo Q adder does nothing. Then X_{mt} appears at the adder output, where it selects one of the Q different frequencies available from the tone generator. This means that with only one user, a message X_{mt}, which is transmitted via MFSK, requires only one chip duration T.

At the receiver, the spectrum of each T second transmission is analyzed to determine which frequency, and hence which K-bit word X_{mt}, is sent. Figure 3.7 shows the operation of the reciever.

The Q-ary MFSK system just described is clearly not suitable for multiple-user operation. If a second transmitter were to generate $X_{m't}$, neither receiver m' nor receiver m would know whether to detect X_{mt} or $X_{m't}$. By using address generators and modulo Q adders and subtractors, many links can share the same spectral band. Thus, with a number of users m>1, L chips are used and frequency hopping is employed to implement the communication.

Now every T seconds the user transmits his information in blocks of a K bit word X_{mt} by a transmitting sequence of L orthogonal frequencies, each with a duration of τ seconds, chosen from Q possible frequencies. The sequence is uniquely determined by the user's address and the K bit message. Users are assigned L word addresses, A_{m1}, A_{m2},..., A_{mL}, where $A_{m\ell}$ is a K bit word corresponding to one of Q available frequencies. This sequence $A_{m\ell}$ is also referred to as the address vector of user m. These address sequences are generated by address generators at a rate of one word every $\tau = \frac{T}{L}$ seconds. The transmitted sequence $Y_{m\ell}$ is then formed as follows:-

$$Y_{m\ell} = X_{mt} + A_{m\ell} \tag{3.1}$$

At the receiver, after each τ seconds, the spectrum of transmission is analyzed to determine which frequency has been sent. To decode the message X_{mt}, the receiver performs the inverse operation of equation (3.1) as follows:-

$$Z_{m\ell} = Y_{m\ell} - A_{m\ell} \tag{3.2}$$

Thus after T seconds, assuming no channel impairments, a duplicate of the transmitted matrix will be formed at the receiver with only one row. If that is the only full row, decoding is complete. However, with many users, the possibility of an error exists. This is because the frequencies of other users can combine to produce a complete row other than X_{mt} in the received matrix. Also, due to channel impairments, noise and multipath propagation, a frequency in the received matrix may be detected when none has been transmitted, giving rise to false alarm probability (P_F). In addition, the receiver can omit a transmitted frequency from the received matrix, thus causing a deletion probability (P_D).

This causes an incomplete row in the received matrix. We therefore use the majority logic decision rule[34] to choose the most likely transmitted code word. This rule chooses the code word associated with the row containing the greatest number of entries.

Under this decision rule, an error will occur when insertions combine to form a row with more entries than the row corresponding to the transmitted code word. An error can also occur when two or more rows contain the same number of entries as in the transmitted row.

3.4 DESIGN PROCEDURE OF FH-MFSK PARAMETERS

The modulation scheme requires a bandwidth that supports Q frequencies each of duration τ seconds. For a given message block size of K bits per user we have

$$Q = 2^K \tag{3.3}$$

The mutual orthogonality of the 2^K frequencies over τ seconds requires a bandwidth of

$$W_H = \frac{2^K}{\tau} \quad \text{Hz} \tag{3.4}$$

Since K bits are transmitted every $T = L\tau$ seconds, the data rate is

$$R = \frac{K}{L\tau} \quad \text{bit/sec} \tag{3.5}$$

Ordinarily, W_H and R will be specified for a particular system application. The goal of the design procedure is to find the values of K and L which maximize the number of users served by the system for a given tolerable bit error rate[34]. A useful quantity for characterizing various spread-spectrum systems is the processing gain (G_p) as mentioned in Section 2.3, then

$$G_p = \frac{W_H}{R} \tag{3.6}$$

Substituting for W_H and R, and using equations (3.4) and (3.5) we have

$$G_p = \frac{2^K L}{R} \tag{3.7}$$

Let M be the maximum number of users which can simultaneously share the system for a given error probability. The efficiency of spectrum utilisation is another parameter characterizing FH-MFSK and is defined as the total rate transmitted through the system per unit bandwidth[44,60],

$$\eta = \frac{MR}{W_H}$$

or

$$\eta = \frac{M}{G_p} \tag{3.8}$$

3.5 ADDRESS ASSIGNMENT FOR FH-MFSK SYSTEM

In a RADA system, the message intended for a particular user is distinguished by a specific signal pattern called the address[53]. Here, address assignement for FH-MFSK will be reviewed and various methods compared. Let us consider a system where an address is a sequence of L frequencies, chosen from Q possible frequencies. Then an L×Q time-frequency matrix will be formed.

Einarsoon[29] and Haskell[44] have studied the problem of assigning addresses to users in FH-MFSK systems using an algebraic approach which provides Q distinct addresses that guarantee minimum interference among M or fewer system users. There are three types of addresses used in FH-MFSK systems and these are explained in the following sections.

3.5.1 Random Address

The system has M simultaneous users, each with an address that is a sequence of words each consisting of K-bits. We assume that each word of an address is selected at random (with equal probability over the 2^K possibilities) and that all addresses in the system are selected independent of each other. If a large number $Q = 2^K$ of frequencies are available for transmission, they will provide sufficient diversity to overcome the fading problem and also reduce the interference between users to a very small amount.

Viterbi[82] has shown that an upper bound on the average bit-error-rate for random addressing is given by

$$P_b \lesssim 2^{K-1}\left[1-(1-2^{-K})^{M-1}\right] \qquad (3.9)$$

Goodman[34] has derived a complete set of formulae for random addressing, which give the upper bound on average bit-error, rate in Rayleigh fading.

Examination of the de-hopping operation (described in Section 3.4) for user m reveals that another user m' can contribute heavily towards filling an erroneous row. This happens if $A_{m\ell} + A_{m'\ell}$ gave the same result for several values of ℓ resulting in the formation of one or more rows (i.e. if the address vectors for users m and m' have the same difference for several chips). In order to reduce this interference, sophisticated address assignments,such as chirp or finite field address, are used.

3.5.2 Chirp Address

In order to minimize the interference mentioned above, one addressing method assigns each user a sequence that increases linearly along the chips with an unique slope to each user[44]. This is called a chirp address. With linearly increasing sequences, the difference between two user addresses also increases linearly and, therefore, cannot have the same value for several chips, which is the desired property. The chirp addresses are constructed having integer slopes between 0 and (Q-1). That is, for a user address having slope m

$$A_{m\ell} = (\ell-1)m, \qquad \ell = 1,2,...,L \tag{3.10}$$

Haskell[44], has shown that the performance of FH-MFSK with chirp address can be estimated using a result derived by Einarsson[29], for finite field address. This will be discussed in the following section. It is also seen that chirp addresses are better than random choices. However, the improvement diminishes as the bandwidth increases[44].

3.5.3 Einarsson Finite Field Address

Einarsson[29], has derived address sequences based on finite field arithmetic which, like the chirp sequences, have differences which are distinct from chip to chip. All arithmetic, in both cases, is defined appropriately for GF(Q). The symbol GF(Q) denotes the Galois finite field of Q elements. A summary of basic features of the finite fields will be described in the next section.

3.5.3.1 Algebra of Finite Field

A finite field GF(Q) is a set of Q elements with rules for addition and multiplication consistently defined[5,29]. As a consequence, a finite field has a zero element and a unit element, both being unique. The zero element has the property $b + 0 = b$ and for the unit element $b.1 = b$ for all $b \in GF(Q)$, (means b is an element of the GF(Q)).

A fundamental result of higher algebra[5] is that there exist finite fields only for Q equal to a prime or the power of a prime number P, i.e. $Q = p^K$. This means that Q = 2,3,4,5,7,8,9,11,13, etc. are permissible but there is no finite field with, for instance, 10 elements.

3.5.3.1.1 Q Equal to a Prime Number

When Q is equal to a prime number, the rules of addition and multiplication in GF(Q) are defined by modulo Q arithmetic[5, 29]. This means that the sum or product between two elements is defined as in usual algebra of integer numbers, with the results reduced modulo Q (i.e. equal to remainder after dividing by Q).

Let Q = 7. We then have, for example,

$3.2 = 6$, $4+2 = 6$, $4.2 = 1$ ($= 8 \bmod 7$),

$5+2 = 0$ ($=7 \bmod 7$)

A non-zero element, $b \in GF(Q)$ is called a primitive element if $b^N = 1$, where $N = Q-1$. Examples: In GF(7), the element b=2 has the powers $b^0, b^1, b^2, b^3,\ldots = 1,2,4,1,2\ldots$ The order of b=2 is thus N=3, (A non-zero element b is said to be of order N if N is the lowest non-zero integer such that $b^N=1$). The element b=3 has the powers $b^0, b^1, b^2, b^3,\ldots = 1,3,2,6,4,5,1,3,\ldots$ shows that b=3 is primitive element i.e. of order 6. It is clear from the examples that the powers of a primitive element are all the non-zero elements of a finite field.

3.5.3.1.2 Q Equal to the Power of a Prime Number

When Q is equal to the power of a prime number P, rules of addition and multiplication are different (5, 29). Consider the case P=2, which is important since binary operations can be used in the field. The elements of the field are represented as p-ary numbers or vectors, of length K. As an example, for $Q=2^5$, the elements are expressed as five digit binary numbers.

Addition is now defined as the binary addition: 1+0 = 1 and 1+1 = 0, which gives 00011+00001 = 00010. This is easily implemented by using an Exclusive-OR circuit.

To specify multiplication in $GF(P^K)$, the elements are transformed into polynomials in Z of degree K-1 by letting the first digit be the coefficient of Z^{K-1}, the second digit the coefficient of Z^{K-2}, etc. The number (11111) corresponds to $Z^4+Z^3+Z^2+Z+1$, 00011 to $Z+1$ and so on.

Addition and multiplication of polynomials are defined as in ordinary algebra using the mod-P rule for the coefficients (5,29). The multiplication rule of $GF(P^K)$ is polynomial multiplication modulo an irreducible polynomial P(Z) of degree K. A polynomial is irreducible if it is not possible to factor it into a product of polynomials of lower degree.

As an example, consider P=2 and K=5, the polynomial $P(Z) = Z^5+Z^3+1$ is irreducible, and the element $b = Z = 010$ is primitive having the table of powers shown in Table 3.1. A derivation of Table 3.1 can be found in Appendix A. Multiplication $GF(2^5)$ is most easily performed by using such a list of the non-zero elements expressed as the power of a primitive element. For instance $8=b^3$ and $19=b^{10}$ giving $8.19=b^{13}=21$, $20.10=b^{30+29}=b^{59}=b^{28}=5$.

3.5.3.2 Einarsson Finite Field Address Assignment

Einarsson has proposed the following way of generating a set of Q addresses[29]. Let

$$A_m = (\gamma_m, \gamma_m b, \gamma_m b^2, \ldots, \gamma_m b^{L-1}) \tag{3.11}$$

where γ_m is an element of GF(Q) assigned to user m and b is a fixed primitive element of GF(Q). He has also shown that, with appropriate choice of b, two transmitted words with different addresses will coincide in, at most, one chip independent of the information transmitted. Since this proposed method can have no more than one coincidence, it is optimum in terms of the amount of interference between users. For this reason Einarsson's finite field address is selected for FH-MFSK. The maximum number of addresses obtained is equal to Q .

TABLE 3.1

Powers of b=2

	b^n	f(Z)	Binary	Decimal
1	b^0	1	00001	1
2	b^1	Z	00010	2
3	b^2	Z^2	00100	4
4	b^3	Z^3	01000	8
5	b^4	Z^4	10000	16
6	b^5	$Z^5 = Z^3+1$	01001	9
7	b^6	Z^4+Z	10010	18
8	b^7	$Z^5+Z^2 = Z^3+Z^2+1$	01101	13
9	b^8	Z^4+Z^3+Z	11010	26
10	b^9	$Z^5+Z^4+Z^2 = Z^4+Z^3+Z^2+1$	11101	29
11	b^{10}	$Z^5+Z^4+Z^3+Z = Z^4+Z+1$	10011	19
12	b^{11}	$Z^5+Z^2+Z = Z^3+Z^2+Z+1$	01111	15
13	b^{12}	$Z^4+Z^3+Z^2+Z$	11110	30
14	b^{13}	$Z^5+Z^4+Z^3+Z^2 = Z^4+Z^2+1$	10101	21
15	b^{14}	$Z^5+Z^3+Z = Z+1$	00011	3
16	b^{15}	Z^2+Z	00110	6
17	b^{16}	Z^3+Z^2	01100	12
18	b^{17}	Z^4+Z^3	11000	24
19	b^{18}	$Z^5+Z^4 = Z^4+Z^3+1$	11001	25
20	b^{19}	$Z^5+Z^4+Z = Z^4+Z^3+Z+1$	11011	27
21	b^{20}	$Z^5+Z^4+Z^2+Z = Z^4+Z^3+Z^2+Z+1$	11111	31
22	b^{21}	$Z^5+Z^4+Z^3+Z^2+Z = Z^4+Z^2+Z+1$	10111	23
23	b^{22}	$Z^5+Z^3+Z^2+Z = Z^2+Z+1$	00111	7
24	b^{23}	Z^3+Z^2+Z	01110	14
25	b^{24}	$Z^4+Z^3+Z^2$	11100	28
26	b^{25}	$Z^5+Z^4+Z^3 = Z^4+1$	10001	17
27	b^{26}	$Z^5+Z = Z^3+Z+1$	01011	11
28	b^{27}	Z^4+Z^2+Z	10110	22
29	b^{28}	$Z^5+Z^3+Z^2 = Z^2+1$	00101	5
30	b^{29}	Z^3+Z	01010	10
31	b^{30}	Z^4+Z^2	10100	20
32	b^{31}	1	00001	1

3.6 NOISE AND INTERFERENCE CONSIDERATION FOR FH-MFSK WITH EINARSSON FINITE FIELD ADDRESS

3.6.1 Noise

Electrical noise can be defined as unwanted signal which is always present in a communication system. It tends to impede the reception of the wanted signal and is usually the limiting factor in signal detection [15].

There are many sources of noise and these may be classified as natural or artificial [9, 15]. Artificial or man-made noise arises from electrical equipment, e.g. electrical motors, sparking plugs and ignition systems, faulty switches, electric shavers, etc. The effect is impulsive in nature and usually consists of damped sinewaves with defined periodicity.

The natural forms of noise are those due to atmospheric noise and circuit noise. Atmospheric noise is usually regarded as that produced by lighting discharges, the noise level being a function of the radiated frequency and the proximity of the discharge. This type of noise decreases rapidly with frequencies above 50 MHz [9]. The other principal component of natural noise, cosmic noise, has a somewhat greater bandwidth and is produced by extra-terrestrial bodies. It is greatest in the range 10 to 300 MHz and is usually the limiting factor in the mid-VHF region in the absence of man-made noise.

The most troublesome form of naturally occurring noise is circuit noise. It is caused by thermal and other noise generated by individual components in the signal path. This type of noise cannot be limited because it is generated due to the nature of electronic components. It can, however, be minimised. Natural noise is essentially gaussian in character if taken over the limited bandwidths found in communication channels [9].

3.6.2 Interference

Interference is caused by unwanted signals, such as those from neighbouring stations. Noise and interference play a somewhat similar role in communication systems, but they are dissimilar in nature in one important respect. It is usually found that noise is composed of randomly-occurring voltages which are unrelated in phase or frequency and may sometimes be of a very peaky nature. Interference, on the other hand, is usually periodic and regular in form and can also be minimised by correct design, construction and maintenance of the transmitter, receiver and aerial system [9,15].

Since there are a limited number of channels available for point-to-point microwave systems, the same channels must be used over and over again. Re-use of microwave frequency bands is enhanced by the directivity of the antennas and the general need for line-of sight reception. In many large cities there is so much traffic converging into one particular area it is impossible to completely isolate two systems using the same channel. This type of intereference is referred to as co-channel interference[4,51].

The separation between adjacent channels and the assignment of frequency channels within specified geographic areas is limited by the parameters for avoidance of adjacent channel interference. Energy from an adjacent channel that does not get rejected is referred to as adjacent channel interference[4]. To achieve a satisfactory frequency channel assignment plan it is necessary to fully study the effects of co-channel and adjacent channel interference on communication systems.

Near-far end interference occurs when the distance between a mobile receiver and the base-station transmitter becomes critical with respect to another mobile transmission that is close enough to override the desired base station signal. This situation usually occurs when a mobile unit is relatively far from its desired base station transmitter[23,24]. Near-far end interference can be overcome by using a RADA technique as explained in Section 3.2.3.2.

The intermodulation interference problem is usually experienced in multi-channel applications. The reason for this type of interference is the non-linearity of power amplifiers or any other devices characterized by non-linearity. When two RF signals of slightly different frequency f_1 and f_2 are amplified in the same amplifier then intermodulation can occur due to inherent characteristic non-linearities. Unwanted outputs of the form $2f_1$, $2f_2$, $f_1 \pm f_2$, $f_1 \pm 2f_2$, $2f_1 \pm f_2$, $2(f_1 \pm f_2)$, $3f_1$, etc., will be produced. High-order products will be attenuated by tuned circuits but products of the form $2f_1 - f_2$, $f_1 - 2f_2$, etc., can be very close to the unwanted frequencies, and it may not be possible to filter them out. This type of interference can be overcome by using linear devices(9).

When digital signals are transmitted over a channel with limited bandwidth or with variable attenuation/frequency characteristics, intersymbol interference is produced. This type of interference can be minimized by using digital coding circuits, equalisers or techniques to reshape the pulse waveforms of the transmitted signal(15, 51, 56).

3.6.3 Error Rate Analysis

In this analysis we assume that the transmission to each square in the received matrix is noncoherent on-off keying. Since the detection of MFSK signals is via a set of M bandpass filters, each centre on one of the possible frequencies followed by an envelope detector. The following assumptions are also made:-

(1) The system is synchronous, which means that the signals from all users are aligned in time.

(2) All sources of errors are additive white gaussian noise A.W.G.N. in the channel and interference from other users.

(3) All signals transmitted in the system are received by every user, creating additional entries in his signal matrix.

The receiver signal is contaminated at the receiver input with A.W.G.N. originating either as pick-up by the antenna or as receiver thermal noise in the RF amplifiers or mixer stages. There are many radio channels characterised by A.W.G.N., mainly of the line-of-sight variety [74]. Short range UHF or VHF or HF ground-wave transmissions are examples. Others are ground-to-air or air-to-ground line-of-sight links, air-to-air links and ground-to-satellite and vice-versa.

When A.W.G.N. is present in the channel, deletion as well as insertions may occur [34,71,74]. That is, a frequency from the desired user may not appear in the detection matrix because the noise may cause the decision statistics to fall below the threshold level. This deletion is estimated by the deletion probability, P_D, and is defined as the probability of no frequency detected in some matrix position, conditional on a frequency transmitted to that position. Also, insertions or false alarms in one or other $(Q-1)$ frequency chip may be caused by noise as well as by other users. The insertion or false alarm probability P_F is defined as the probability of detecting a frequency when no such frequency was sent by any user of the system.

The two probabilities P_D and P_F are given by Stein and Schwartz[71,74]. These formulae depend on the signal-to-noise ratio and threshold level and are given by:

$$P_D = 1-Q_m(\sqrt{2\gamma},\ b_o) \tag{3.12}$$

and

$$P_F = e^{-b_o^2/2} \tag{3.13}$$

where,

$$\gamma = \frac{\mu^2}{2N} \tag{3.14}$$

is the input signal-to-noise ratio (S/N) at the sampling instant, and

$$b_o = \frac{b_t}{\sqrt{N}} \tag{3.15}$$

is the threshold level normalized to the rms noise voltage and $Q(\sqrt{2\gamma},\ b_o)$ is the Marcum function[55,71,74] given by:

$$Q_m(\alpha,\beta) = \int_\beta^\infty t\ I_o(\alpha t)\ e^{-(t^2+\alpha^2)/2} dt. \tag{3.16}$$

where

$I_o(\alpha t)$ can be identified in terms of the defining integral for the modified Bessel function of the first kind and zero order,

$$I_o(\alpha t) = \frac{1}{2\pi} \int_o^{2\pi} e^{\alpha t\cos\theta}\ d\theta \tag{3.17}$$

Another degradation in the system performance is due to the mutual interference between users[29,34,44,60,63].

Any one of the (Q-1) frequency chips on which the desired user did not transmit may be filled incorrectly simply as a result of a frequency transmitted by an interfering user.

An error will occur when interfering signals combine in such a way that one or more erroneous messages occur in the signal matrix. The upper-bound of the error probability for a particular sequence of the desired user will be formed by the M-1 interfering signals from the other users in the system[(29)].

Einarsson[(29)], has derived the error probability for the noiseless case with finite field address by studying the probability of a chip being filled conditional on the previous chip being occupied. In this section the error probability in an environment with A.W.G.N. is developed. Let I_ℓ denote the number of interfering signals occupying chip number $\ell, \ell = 1,2,\ldots,L$, of the fixed pattern Y, of user 1.

Let P_R be the probability of M-1 interfering users and A.W.G.N. producing a complete erroneous row of length L at a specific location in the detection matrix. Then the probability P_R is given by the product of conditional probabilities

$$P_R = P(I_1 \neq 0)\, P(I_2 \neq 0 \mid I_1 \neq 0) \ldots\ldots \times P(I_L \neq 0 \mid I_1 \neq 0, I_2 \neq 0, \ldots, I_{L-1} \neq 0) \tag{3.18}$$

The conditional probabilities in equation (3.18) are due to the fact that, because of the address coding, each transmitted word depends on the previous transmitted word. Each user has an unique set of L address words that is used to code each transmitted word.

The first term in equation (3.18) is the overall error probability for one chip ℓ in a spurious row j due to other users and A.W.G.N.

$P(I_1 \neq 0)$ can be calculated as follows by considering one chip (j,ℓ) and one interfering user. The probability of that user not sending a tone to that chip in the detection matrix is

$$1 - \frac{1}{Q} \quad , \tag{3.19}$$

and the probability that none of the M-1 interfering users send a tone to chip (j,ℓ) is

$$(1 - \frac{1}{Q})^{M-1} \tag{3.20}$$

The probability of a frequency transmitted to chip (j,ℓ) is given by

$$P_x = 1-(1- \frac{1}{Q})^{M-1} \tag{3.21}$$

P_x, P_D and P_F can be assumed as independent, though not mutually exclusive events [34]. Then the overall error probability for chip (j,ℓ) [12] is

$$P(I_1 \neq 0) = P_x + P_D + P_F - P_x P_D - P_x P_F - P_D P_F + P_x P_D P_F \tag{3.22}$$

and hence

$$\therefore P(I_1 \neq 0) = P_x \left[1 - (P_D + P_F - P_D P_F)\right] + P_D + P_F - P_D P_F$$

or

$$P(I_1 \neq 0) = \left[1 - \left(1 - \frac{1}{Q}\right)^{M-1}\right]\left[1 - (P_D + P_F - P_D P_F)\right] + P_D + P_F - P_D P_F \tag{3.23}$$

The first two terms of equation (3.18) can be expanded as

$$P(I_1 \neq 0) P(I_2 \neq 0 \mid I_1 \neq 0) = \sum_{i=1}^{M} P(I_1 = i) P(I_2 \neq 0 \mid I_1 = i) \tag{3.24}$$

when $I_1 = i$, there are M-i-1 interfering signals which can contribute to I_2. Then,

$$\therefore P(I_2 \neq 0 \mid I_1 = i) = \left[1 - \left(1 - \frac{1}{Q-1}\right)^{M-i-1}\right]\left[1 - (P_D + P_F - P_D P_F)\right]$$

$$+ P_D + P_F - P_D P_F \tag{3.25}$$

where $\frac{1}{Q-1}$ is the probability of chip 2 being filled by any of the M-i-1 signals not occupying chip 1. This reduction of possible choices from Q to Q-1 is due to the finite field addresses which have the property that two transmitted vectors with different addresses coincide in at most one chip(29).

From equation (3.25), $P(I_2 \neq 0 \mid I_1 = i)$ takes its largest value when i=1. Under this condition equation (3.24) will overbound by

$$P(I_1 \neq 0)\,P(I_2 \neq 0 \mid I_1 \neq 0)$$

$$< \left[\left[1-(1-\frac{1}{Q})^{M-1}\right] \left[1-(P_D+P_F-P_D P_F)\right] + P_D+P_F-P_D P_F \right]$$
$$\times \left[\left[1-(1-\frac{1}{Q-1})^{M-2}\right] \left[1-(P_D+P_F-P_D P_F)\right] + P_D+P_F-P_D P_F \right] \quad (3.26)$$

The next term in equation (3.18) is bounded in the same way by replacing

$$P(I_3 \neq 0 \mid I_1 = i, I_2 = j) = \left[1-(1-\frac{1}{Q-2})^{M-i-j-1}\right]$$
$$\left[1-(P_D+P_F-P_D P_F)\right] + P_D+P_F-P_D P_F \quad (3.27)$$

by its largest value, which is given by i=j=1. Repeating the same procedure we arrive at

$$P_R < \prod_{\ell=0}^{L-1} \left[\left[1-(1-\frac{1}{Q-\ell})^{M-\ell-1}\right] \left[1-(P_D+P_F-P_D P_F)\right] + P_D+P_F-P_D P_F \right] \quad (3.28)$$

The bit error rate P_b is given by[71]

$$P_b = \frac{2^{K-1}}{Q-1} P_R \quad (3.29)$$

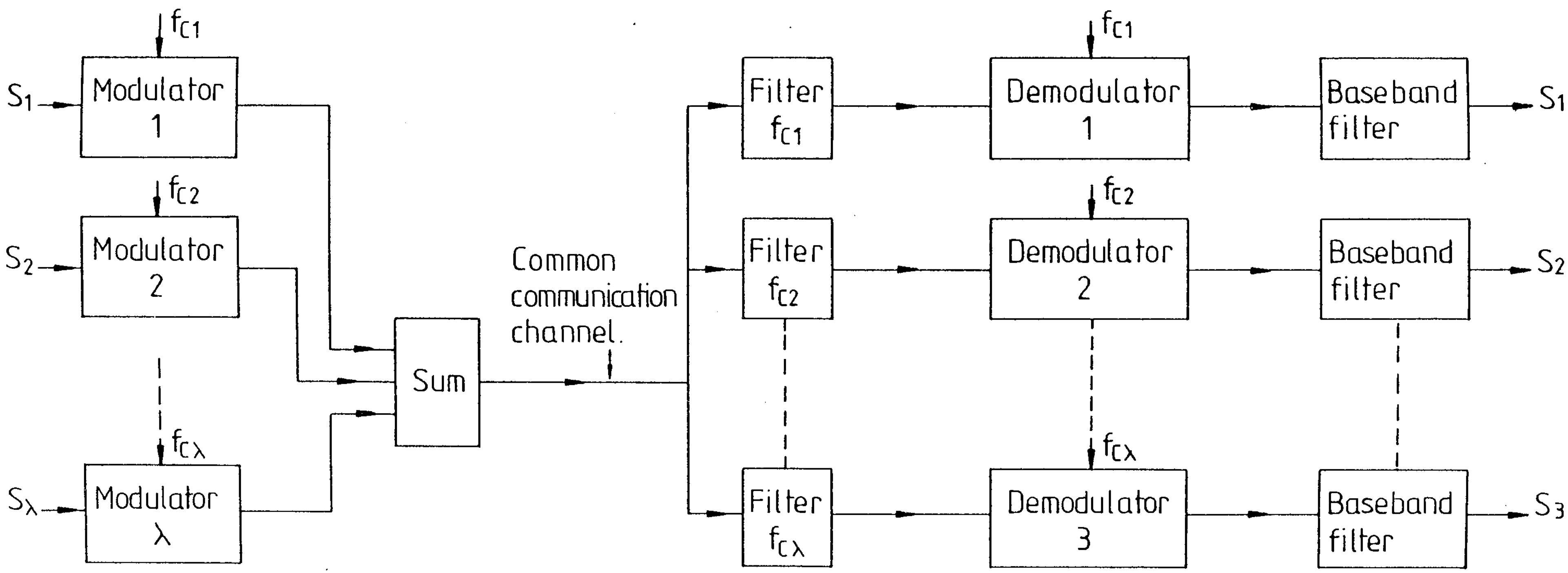


Fig 3.1 Frequency division multiplexing system.

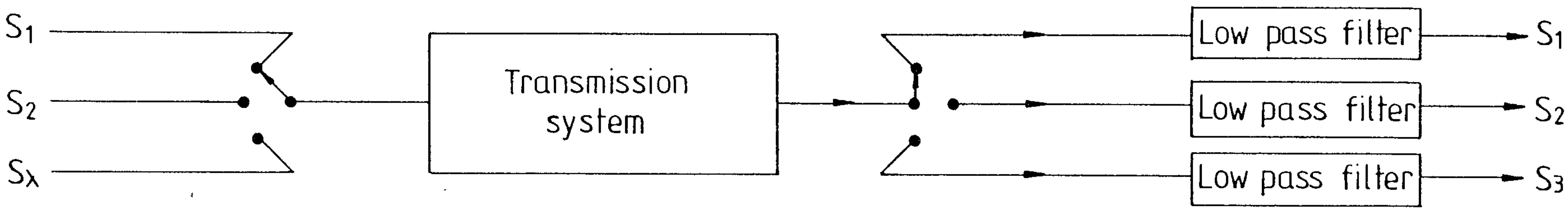


Fig 3.2 Time division multiplexing system.

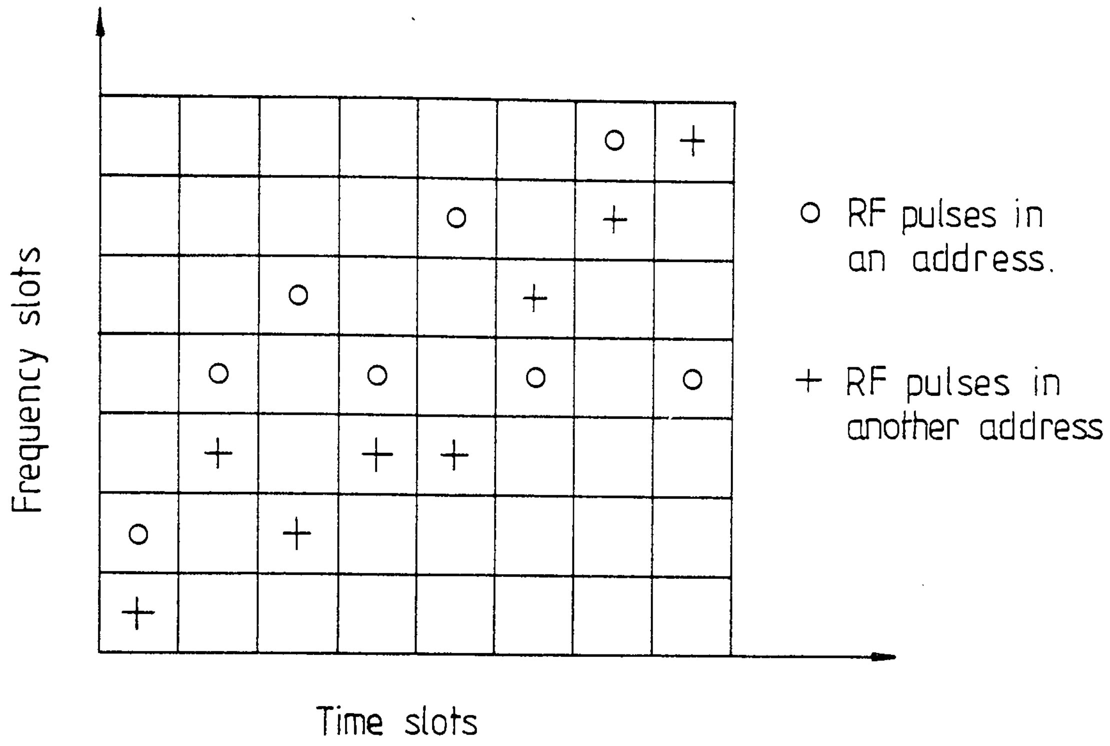


Fig 3.3 RADA address assignment

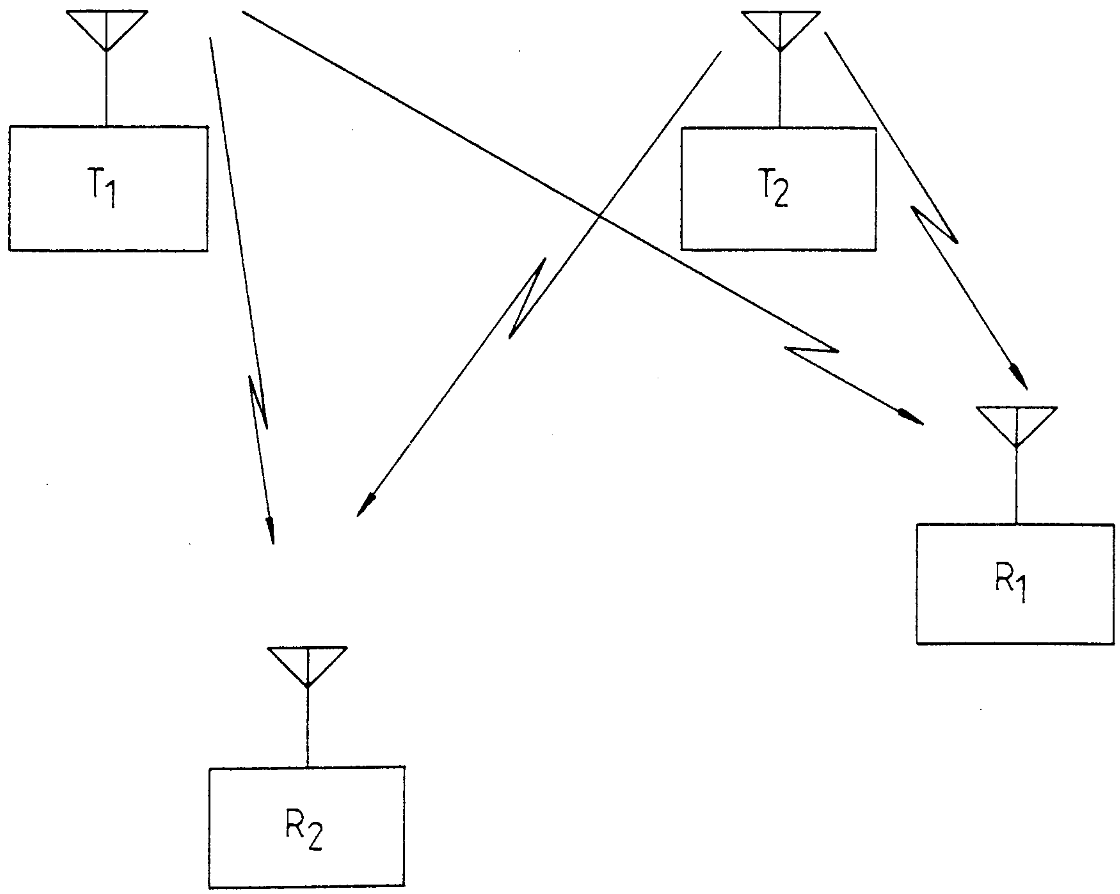


Fig 3.4 Two link communication system illustrating the near - far problem.

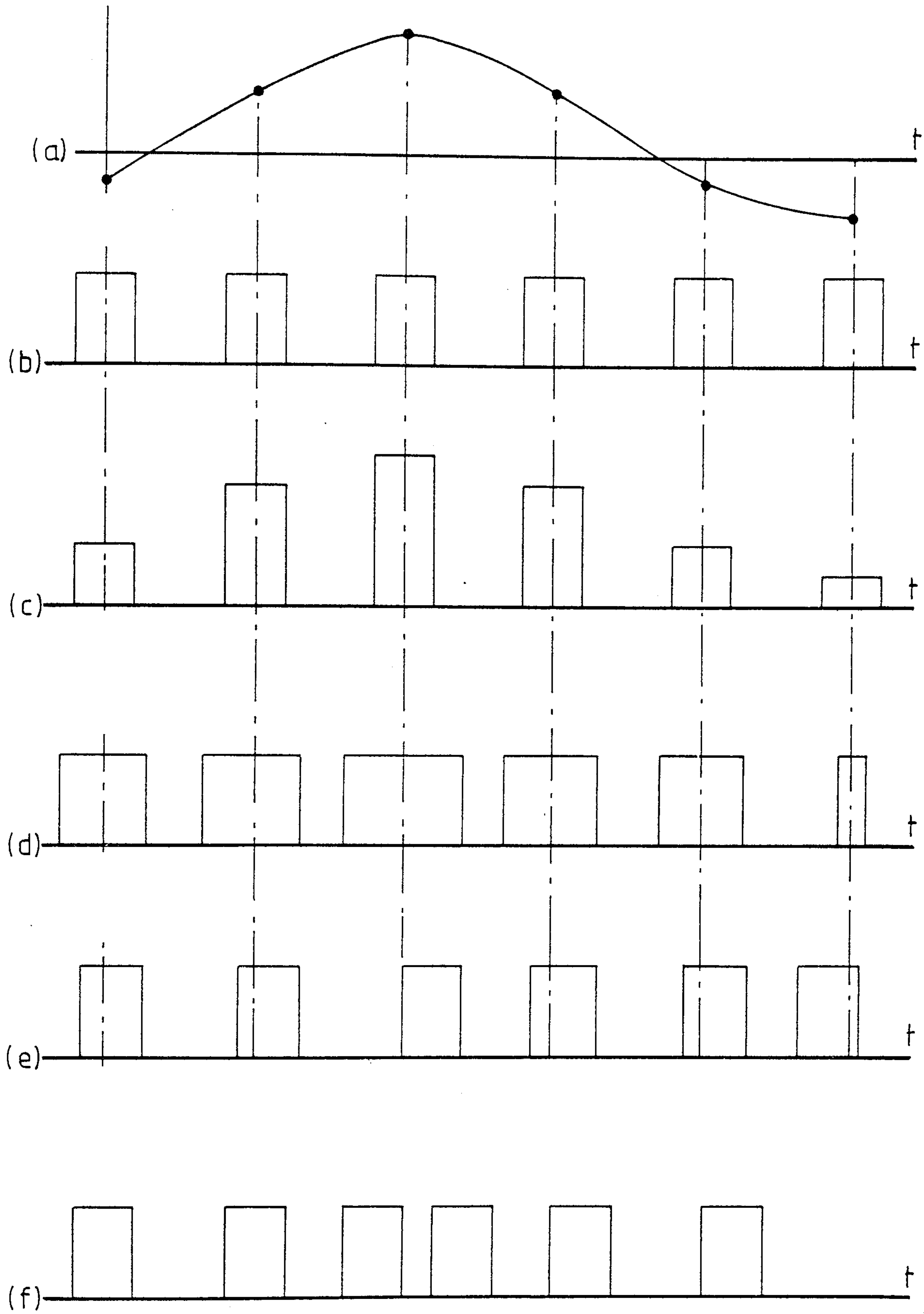
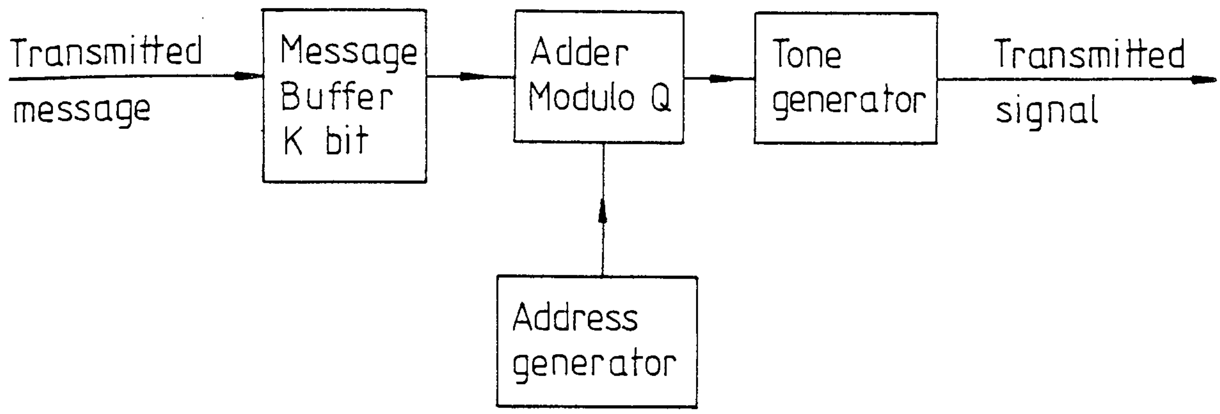


Fig 3.5 Types of analogue pulse modulation :
(a) Modulating signal (b) Carrier signal
(c) PAM (d) PWM (e) PPM (f) PFM



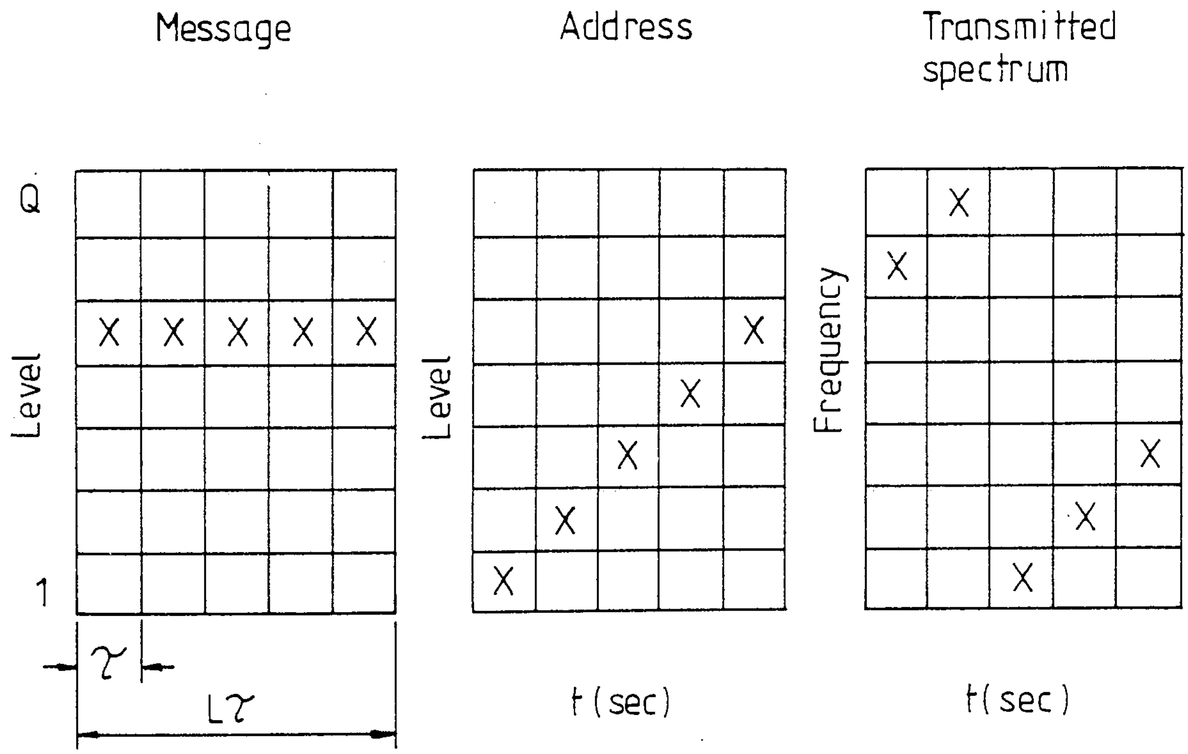


Fig 3.6 FH-MFSK transmitter block diagram and signal matrices.

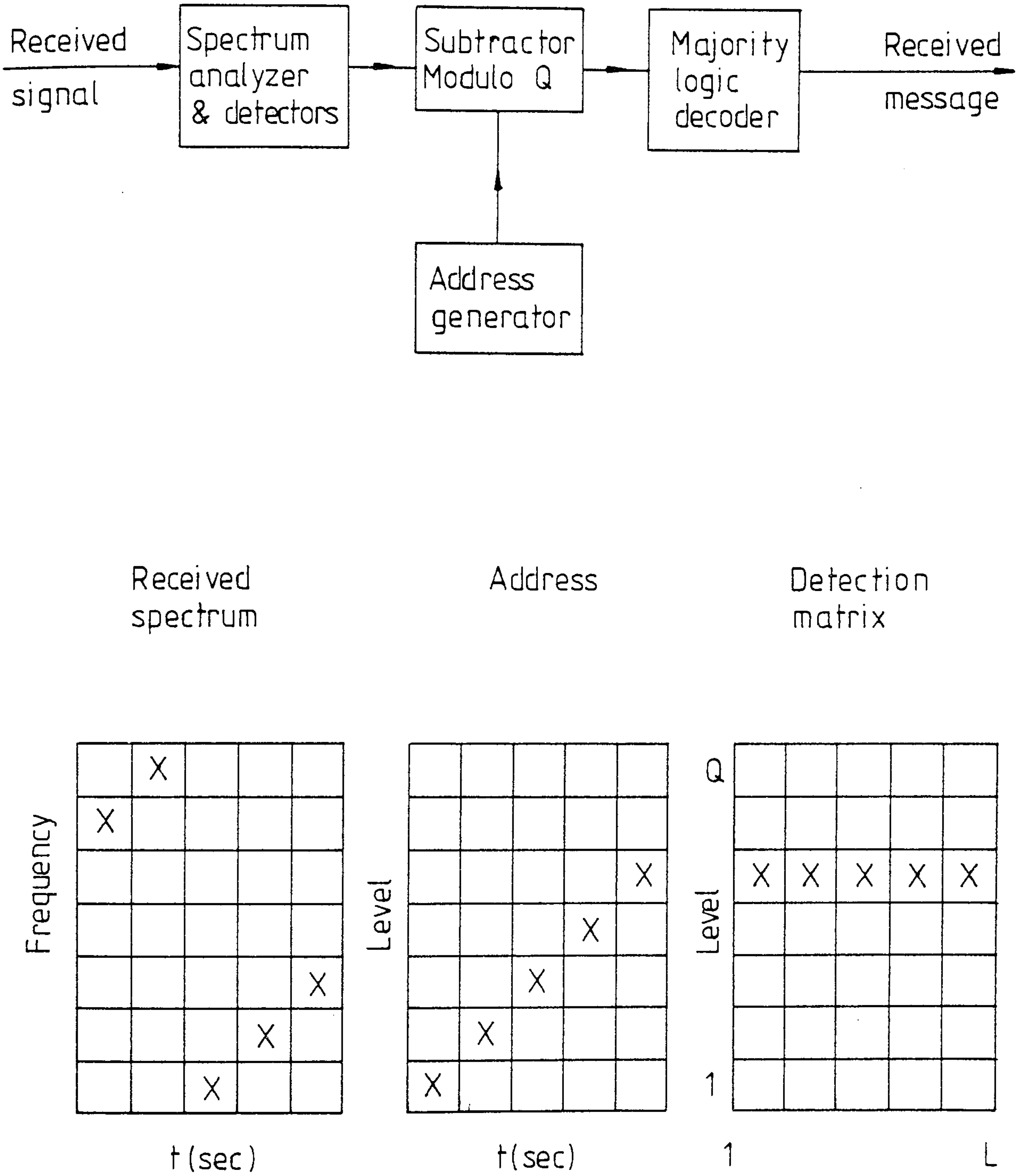


Fig 3.7 FH-MFSK receiver block diagram and signal matrices.

CHAPTER FOUR

FREQUENCY SYNTHESIS AND SPECTRAL ANALYSIS

FREQUENCY SYNTHESIS AND SPECTRAL ANALYSIS

4.1 INTRODUCTION

The frequency synthesis section of an FH-MFSK system is the heart and the most critical part of the system[24]. The number of frequencies generated, the hopping rate and the total spread spectrum bandwidth are the quantities that determin the system capabilities. The frequency hopping synthesizer is designed to switch quickly from one frequency to another, whereas a conventional synthesizer may be required to change frequencies only at long and irregular intervals.

Since the received signal of an FH-MFSK system is mapped from a large set of orthogonal signals spread over a wide band, spectral analysis is required. A fast processor must be used for the spectral analysis of the FH-MFSK signal in order to satisfy the information rate and bandwidth requirements.

Various techniques for frequency synthesis and spectral analysis are dicussed in this chapter.

4.2 FREQUENCY SYNTHESIZER PROPERTIES

A frequency synthesizer is an instrument which generates one or more frequencies from one or several reference sources[54]. Many different factors must be considered to determine the cost, weight and power consumption of a synthesizer in a particular application[26,32 37,62,66]:

(1) Resolution

Resolution is the minimum frequency change between two adjacent output frequencies. It depends on the application and in some instances may be 100 KHz or as fine as 0.01 Hz.

(2) Switching Time

Switching time is the time needed to switch the synthesizer from one frequency to another frequency. All synthesizers use some kind of filtering, active or passive. In many cases, the narrowest filter bandwidth affects the switching time of the system. It is important, therefore, to be able to estimate the longest switching time associated with the narrowest bandwidth of the filters, chosen either to attenuate certain spurious outputs or to reduce phase noise.

Switching time also determines the hop rate for a FH synthesizer. The usual rule for determining maximum hopping rate for a frequency synthesizer is that the period of transmission for one frequency should be 10 times the time required for setting to that new frequency[24].

(3) Number of Frequencies

The number of frequencies that can be generated by the synthesizer is an important factor for a FH synthesizer as it determines the capability of the system as mentioned in section 2.5.2.

(4) Spurious Level

A very important factor in the design of a synthesizer is the purity of the generated signal or the absence of unwanted sidebands (noise, intermodulation). Each part of the system must be examined for this distortion. It is advantageous to use circuits that give theoretically little in the way of unwanted sidebands. In practice it is possible to reach quite low levels of spurious signals with proper construction.

(5) Frequency Standard

Frequency standard is the deviation of the reference source frequency per day or per year from the nominal centre frequency over a specific temperature range, e.g., 1 part in 10^9 variation in output frequency per day over 40^{o} to $100^{o}C$ temperature range. All composing frequencies must be derived from the same reference to maintain the frequency stability of the output signal at the same level as the frequency standard.

(6) Possibility of Microminaturization

In communication systems, especially for mobile use, volume and weight are of prime importance.

(7) Programmability

In recent remote control systems, complete automatic synthesizers are used. The frequency synthesizer used for this purpose must be digitally controlled[66].

4.3 FREQUENCY SYNTHESIS TECHNIQUES

In general there are two main techniques for frequency synthesis, namely coherent and incoherent synthesis. The difference between them is the number of frequency sources used in the process of frequency generation[54]. In incoherent techniques a set of frequency sources are used, whereas in coherent techniques only one frequency source is used. The stability and accuracy of the output frequency in a coherent synthesis are therefore the same as the stability and accuracy of the reference source.

Coherent synthesis can be divided into two general types[24,54]. The first is direct coherent synthesis. This includes all those techniques that generate the desired output frequencies from one source by mixing, multiplying, dividing or any other means that do not include phase-locked loops. The second is indirect coherent synthesis. This includes any synthesizer employing a phase locked loop.

4.3.1 Incoherent Synthesis Technique

The output frequencies generated from the input frequencies vary in incoherent synthesis, depending on the application. Factors mentioned in Section 4.2 govern the choice of the synthesizer which use this technique. The main goal of this technique remains the same in all cases; that is, to minimize the number of crystals and the basic building blocks used in synthesis, such as oscillators, mixers, and filters in order to reduce cost. Fig. 4.1 shows a block diagram of an incoherent synthesizer [54].

This technique uses additive heterodyning. The frequency stability, accuracy, and noise of the output are therefore the sums of the stabilities, accuracies and noises of the individual frequency sources. The important advantage of incoherent synthesis is low cost, but special attention should be paid to the spurious outputs generated by the mixing process.

4.3.2 Coherent Direct Synthesis Techniques

The brute-force technique [66], is the simplest method of coherent direct synthesis. It is only used when a small number of frequencies are required to be generated but can be used whenever these frequencies have to be generated simultaneously.

Recently, considerable interest has been directed towards the realisation of coherent direct synthesizers using surface acoustic wave (SAW) technology [20,21,36,39,59]. The digitally-controlled coherent synthesizer, employing a SAW chirp filter, is one attractive realisation for obtaining a fast switched (nano seconds), phase coherent output with wide bandwidth, these being the main requirements of a FH synthesizer.

4.3.2.1 Brute-Force Technique

This is customarily used for the simultaneous generation of a small number of frequencies [54,66]. The basic building blocks utilized in a brute-force technique are frequency multipliers and dividers as shown in Fig. 4.2. The main problem associated with this technique is spurious outputs generated in multiplication and division.

4.3.2.2 Frequency Hop Synthesizer Based on Chirp Mixing Using SAW Technology

This section reviews SAW technology and its application to frequency hopping generation based on mixing two linear frequency modulated (chirp) waveforms.

4.3.2.2.1 SAW Technology

Acoustics is defined as the generation, transmission and reception of energy in the form of vibrational waves in matter [48]. As the atoms or molecules of a fluid or solid are displaced from their normal configurations, an internal elastic restoring force arises. It is this elastic restoring force, coupled with the inertia of the system, that enables matter to participate in oscillatory vibrations and thereby generate and transmit acoustic waves.

Acoustic phenomena are generally associated with the sensation of sound. A vibrational disturbance is interpreted as sound if its frequency lies in the range of about 20 Hz to 20 KHz. However, in a broader sense, acoustics also includes the Ultrasonic (frequencies above 20 KHz) and the Infrasonic (frequencies below 20 Hz)[31].

SAW technology is a new technology[59]. It is basically an analogue device which uses Rayleigh wave propagation on a quartz crystal. A Rayleigh wave[58] is a surface wave on the free plane surface of an elastic half space. Surface waves on plane boundaries are waves whose amplitude changes exponentially with the normal distance from the surface. SAW devices can be used to perform many signal processing operations. Among their important applications are those of convolving time limited signals with each other and Fourier transforming time limited signals[59]. Also bandpass and pulse compression filters and delay lines can be designed using SAW technology[19,43,59,77].

The major advantages of SAW devices are, their simplicity compared to digital integrated circuits, and their larger useable bandwidth, up to 100 MHz[59].

4.3.2.2.2 Generating the Linear Frequency Modulation or Chirp Signal

There are two basic approaches to the generation of linear FM signals. One approach is known as active signal generation since it is based on controlling the frequency of an oscillator with a voltage derived from an appropriate function generator[17]. The alternate approach is known as passive generation of the linear FM signal since an impulse of the approximate length of the desired compressed pulse is generated and inserted into a dispersive delay device which has a linear time delay versus frequency characteristic[61]. The waveform from the dispersive delay device is a linear FM signal. The compressed pulse for a linear FM wave can be obtained by passing a narrow pulse through a rectangular filter. The passive approach based on SAW technology is more attractive as will be seen from the next section.

4.3.2.2.3 Description of an FH Synthesizer Based on Chirp Mixing

An FH synthesizer based on chirp mixing[20,21,36,39,59], is shown in Fig. 4.3. Chirp signals can be generated by energising the SAW filters with impulses as mentioned in the previous section. This is a passive method for generating chirp signals. A constant frequency output can be obtained at either the sum or difference component of the mixer output. The selection of either the sum or the difference frequency depends on the filter. Gating was used to restrict the output pulse length to prevent the extra product terms which are generated at the unwanted difference frequency. This restriction in the output pulse length permits the design of a synthesizer which can hop over half the bandwidth of the chirp filter. Because of gating, a second channel identical to the first, except for timing difference in code generator and gating, is required to generate pulses occupying the space between the pulses from channel 1,[36]. For perfect signals, the two chirp waveforms $S_{c1}(t)$, $S_{c2}(t)$ can be expressed as[20,39]

$$S_{c1}(t) = \cos[\omega_{c1} t + \tfrac{1}{2}\mu_c t^2]. \qquad 0<t<T_c \qquad (4.1)$$

$$S_{c2}(t) = \cos[\omega_{c2}(t-\tau_c) + \tfrac{1}{2}\mu_c (t-\tau_c)^2]. \qquad \tau_c<t<T_c+\tau_c \qquad (4.2)$$

where

ω_c = starting frequency

μ_c = the dispersive slope

T_c = chirp duration

τ_c = the time delay between the impulses for both channels.

The product of these waveforms is given by

$$\begin{aligned} S_{co}(t,\tau_c) = \tfrac{1}{2}\cos\Big[& [\omega_{c1}+\omega_{c2}-\mu_{c2}\tau_c]t - \omega_{c2}\tau_c \\ & - \tfrac{1}{2}\mu_{c2}\tau_c^2 + \tfrac{1}{2}t^2(\mu_{c1}+\mu_{c2})\Big] \\ & + \tfrac{1}{2}\cos\Big[[|\omega_{c1}-\omega_{c2}| + \mu_{c2}\tau_c]t + \omega_{c2}\tau_c \\ & + \tfrac{1}{2}\mu_{c2}\tau_c^2 + \tfrac{1}{2}t^2(\mu_{c1}-\mu_{c2})\Big]. \qquad \tau_c<t<T_c \qquad (4.3) \end{aligned}$$

Selecting the sum term, the instantaneous angular frequency is

$$\omega_{cs} = \omega_{c1}+\omega_{c2}-\mu_{c2}\tau_c + (\mu_{c1}+\mu_{c2})t \qquad (4.4)$$

If $\mu_{c1}=-\mu_{c2}=\mu_c$ then the residual chirp term will vanish, resulting in an output frequency which is linearly dependent on τ_c. Similarly, the instantaneous difference angular frequency is given by

$$\omega_{cd} = \left|\omega_{c1}-\omega_{c2}\right| + \mu_{c2}\tau_c + (\mu_{c1}-\mu_{c2})t \qquad (4.5)$$

The generation of FH waveforms using the sum frequency method is of particular interest for high frequency generation and is widely used in spread spectrum systems.

This synthesizer has the following error sources[21]:

(1) Impulse timing errors which arise from clock instability and impulse jitter which results in frequency error.

(2) Chirp slope mismatch which results in a residual linear FM term in equations (4.4) and (4.5).

(3) Chirp centre frequency errors which directly degrade frequency accuracy and invalidate the phase coherence condition.

(4) Gate edge timing errors which directly influence spurious levels.

(5) Spurious output from the mixer which arises through harmonic generation within the mixer due to inherent non-linearities.

Satisfactory operation can be obtained when the above errors are minimised. This synthesizer allows fast frequency hop genaration over a wide bandwidth (500 MHz) with large numbers of frequencies (4000) satisfying the FH communication equipment requirements [21].

However, this technique was not implemented because of the high cost of the chirp filters.

4.3.3 Coherent Indirect Synthesis Technique

Indirect synthesis utilizes the principle of feedback in generating frequency increments. The system analysis of indirect synthesis centres on an investigation of phase-locked loop (PLL) stability and acquisition. The acquisition is a process of bringing a loop into lock. The lock-in range of the PLL is the frequency range over which the loop acquires phase without slips [32,54]. The basic components of a PLL synthesizer are shown in Fig. 4.4. The PLL synthesizer, in effect, multiplies the reference frequency by a variable number [26]. It does so by dividing its output frequency by that variable number and adjusting the output frequency so that, after division, it is equal to the reference frequency. When the difference between these frequencies is small, the phase detector generates a slowly varying a.c. voltage which is passed by the low pass filters and pulls the VCO in lock. Under locked conditions, the output of the phase detector is a d.c. voltage, whose amplitude and polarity are determined by the amount and direction of phase between the reference and the down-converted VCO signals. The low pass filter changes the amplitude and phase of individual signals passing through it as functions of the working frequency, to achieve stable loop performance. At lock frequency

$$f_{out} = N_L \frac{f_{ref}}{M_L} \tag{4.6}$$

where

f_{out} = output frequency

f_{ref} = reference frequency

N_L = integer number

M_L = integer number.

However, the stability of the VCO at high frequency is a major problem. This technique has now been used up to 8 GHz (54). The most striking feature of the PLL synthesizer is its small size, weight and power consumption. This is because:-

1. The PLL technique makes possible frequency multiplication by very large factors in one stage only (49).

2. Generally, the VCO is the only circuit that contains an LC (Inductance Capacitance) network. Thus the wide-spread use of integrated circuits makes it possible for the advantages of small size and cheap mass production is to be realised.

4.4 SPECTRAL ANALYSIS

Spectral analysis is the analysis of signals in the frequency domain[11,69]. It is based on the Fourier series and Fourier transform. There are two basic approaches to spectral analysis[25]. The first is the swept frequency approach where the analysis is done in a serial manner. This method suffers the obvious drawback of being slow, though economical, and is only used in measurement. The second approach is the parallel analysis approach, where all the frequency bands of interest are analyzed simultaneously. The parallel analyzer is obviously a much more complex machine than the serial analyser.

In recent years digital computation methods have increased the speed by a factor of a thousand or more, while the cost of digital computation has decreased by a factor of the order of four hundred or greater. The technology of large scale integrated, (LSI), fabrication has simultaneously advanced so that it is now feasible to fabricate circuits of small dimensions at low cost. One consequence of this development in LSI technology is the rapidly increasing use of digital techniques for spectrum analysis[13,25,64]. It is at this point that the Discrete Fourier Transform (DFT) comes into its own. The DFT is a version of the Fourier analysis technique which is suitable for digital implementation. This will be discussed in the next section.

Most digital computations of spectra,[47], have a common procedure:-

1. The analogue signal is sampled every T_s time units for a time T, thus $N = \frac{T}{T_s}$ digital samples are required.

2. The data is then multiplied by a window function.

3. It is Fourier Transformed.

4. The squared magnitude of the resulting transform gives the desired estimate.

4.5 DFT AND FFT

The DFT is a relation between two finite length sequences of complex numbers. It is a powerful reversible mapping function for time series[1,76]. The DFT of a finite length sequence $(x_o, x_1, \ldots, x_{N-1})$ is $(X_o, X_1, \ldots, X_{N-1})$ and is given by:

$$X_r = \frac{1}{N} \sum_{n=0}^{N-1} x_n W_F^{rn} \tag{4.7}$$

for $r = 0, 1, \ldots, N-1$

where

W_F = rotation factor

$$= e^{-j2\pi/N} \tag{4.8}$$

When digital analysis techniques are to be used for analyzing a continuous waveform, it is necessary that the data be sampled, as mentioned in section 4.4, to produce a time series of discrete samples. This can then be fed into a digital computer. Such a time series completely represents the continuous waveform, provided this waveform is frequency band-limited and the samples are taken at a rate that is at least twice the highest frequency present in the waveform[76].

The Fast Fourier Transform (FFT) is simply an efficient method for computing the DFT[6,13,18,64]. The idea behind the FFT is to break the original N-point sequence into two shorter sequences, the DFT's of which can be combined to give the DFT of the original N-point sequence. Euqation (4.7) for the DFT involves N^2 complex multiplications and $N(N-1)$ additions. Thus, if equation (4.7) is applied to the two records of length $\frac{N}{2}$, then $2(\frac{N}{2})^2$ multiplications are needed. This reduction in the number of multiplications represents a saving of 50% in computation time. The process above can be iterated so as to reduce the computation of an $(\frac{N}{2})$ point DFT to two $(\frac{N}{4})$ DFTs and thereby effect another saving of a factor of 2[13,64]. A general result for analyzing the number of operations for N-points is that the FFT requires N complex additions and $\frac{1}{2}N$ complex multiplications per stage. Since there are $\log_2 N$ stages, the number of complex multiplications and additions required to evaluate an N-point DFT are $\frac{N}{2}\log_2 N$ and

$N\log_2 N$ respectively. Figure 4.5 shows the number of operations required by the FFT and the DFT.

There are two forms of FFT decimation[1,76]. The first form is decimation in time (DIT), in which the re-ordering is done on the input samples as shown in Fig. 4.6. The second form is decimation in frequency (DIF), in which the re-ordering is done at the final stage of computation, rather than in the initial stages, as shown in Fig. 4.8. There are two apparent differences in the computation of DITs and DIFs. The first is the re-ordering of input and output samples as mentioned above. Secondly, the DIF butterfly shown in Fig. 4.9 is slightly different from the DIT butterfly shown in Fig. 4.7, the difference being that the complex multiplication takes place after the add-subtract operation in the DIF.

The DFT can be used as a bank of narrowband filters. The attractiveness of the DFT-based filter bank is the economy of computation when realised as an FFT[41], as mentioned above.

Spectral estimation of FH-MFSK signals can be implemented by the FFT technique with envelope detection to generate the 2^K spectral estimates[82]. Since FH-MFSK is a bandlimited signal and is formed from 2^K orthogonal frequencies, FFT is a more efficient technique for spectral estimation. The DIF form is used for FFT implementation as shown in the next chapter, since it does not require re-ordering of the input samples.

4.5.1 Problems in Calculating the DFT

There are three major problems in calculating the DFT, aliassing, leakage and the picket-fence[6,13,42].

1. Aliassing

Aliassing refers to the fact that high frequency components of a time function can impersonate low frequencies if the sampling rate is too low[6,13]. This is demonstrated in Fig. 4.10, by showing a relatively high frequency and a relatively low frequency that share identical sample points. This uncertainty can be removed by demanding that the sampling rate be high enough for the highest frequency present to be sampled at least twice during each cycle[6].

2. Leakage

This is a phenomenon introduced by having to observe (multiply) the data via a finite observation window, i.e. using gating in time. To reduce this problem, a suitable window which has lower sidelobes must be selected. Windows are weighting functions applied to data to reduce the spectral leakage associated with finite observation intervals[42].

3. Picket-Fence Effect

This becomes evident when the signal being analyzed is not one of those discrete orthogonal frequencies $\frac{1}{NT_s}$ (T_s

is the sampling interval). If the signal has frequency components between the harmonic of $(\frac{1}{NT_s})$, the amplitudes of these frequency components are reduced. To decrease this problem, artificially increasing the data record length by adding zeros gives a more accurate estimate of the envelope of the DFT or, equivalently, a more accurate estimate of the frequency components between the original harmonics at $(\frac{1}{NT_s})$, [6, 13].

4.5.2 Types of FFT Processors

The implementation of an FFT processor is usually dictated by speed, accuracy, size and cost requirements. Basically there are two ways to manipulate an array of data to compute an FFT, the in-place and the double memory techniques[3, 70]. The in-place technique uses only half the memory of the other methods but requires a much more complex control. In the in-place technique, two complex words are retrieved from the memory and operated upon to generate two new complex words, which are then stored in the same memory location. In the double memory technique, two blocks of complex words are switched back and forth in the roles of sending and receiving memories. While this doubles the required memory, the control is simpler.

FFT processors can be classified according to the performance and the organisation of the arithmetic unit[7, 64].

(1) The Sequential Processor

The basic operation (i.e. an addition and a subtraction followed by complex multiplication) can be applied sequentially to the $N/2\log_2 N$ sets of data. The in-place technique can be used to store the input-output data. A simplified block diagram of this processor is shown in Fig. 4.11. The sequential processor will be characterized as having:-

(a) One arithmetic unit.

(b) $N/2\log_2 N$ operations performed sequentially.

(c) An execution time of $B_s N/2\log_2 N$ where B_s is the time required for performing one basic operation.

(2) The Cascade Processor

To improve the performance of the sequential processor, parallelism can be introduced into the block diagram by using a separate arithmetic unit for each pass. The speed of execution can be increased by a factor of $\log_2 N$. A simplified block diagram of the resulting processor is shown in Fig. 4.12. The cascade processor will be characterized as having:-

(a) K arithmetic units (if $N = 2^K$).

(b) K passes performed in parallel.

(c) $\frac{N}{2}$ operations per pass performed sequentially.

(d) Execution time is equal to NB_s.

(3) The Parallel Iterative Processor

To improve the performance of the above two processors, arithmetic units can be used to perform the operations per pass in parallel before performing the operations for the next pass. A simplified block diagram of this processor is shown in Fig. 4.13. This processor will be characterized as having:-

(a) $\frac{N}{2}$ arithmetic units.

(b) $\frac{N}{2}$ operations performed in parallel.

(c) K passes performed sequentially.

(d) Execution time is equal to $B_s \log_2 N$.

(4) The Pipeline Processor

It is possible to develop efficient processors which have parallel features wherein data samples can be read and written in parallel. A simplified block diagram is shown in Fig. 4.14. This processor will be characterized as having:-

(a) $N/2\log_2 N$ arithmetic units.

(b) $N/2\log_2 N$ operations performed in parallel.

(c) An execution time equal to B_s.

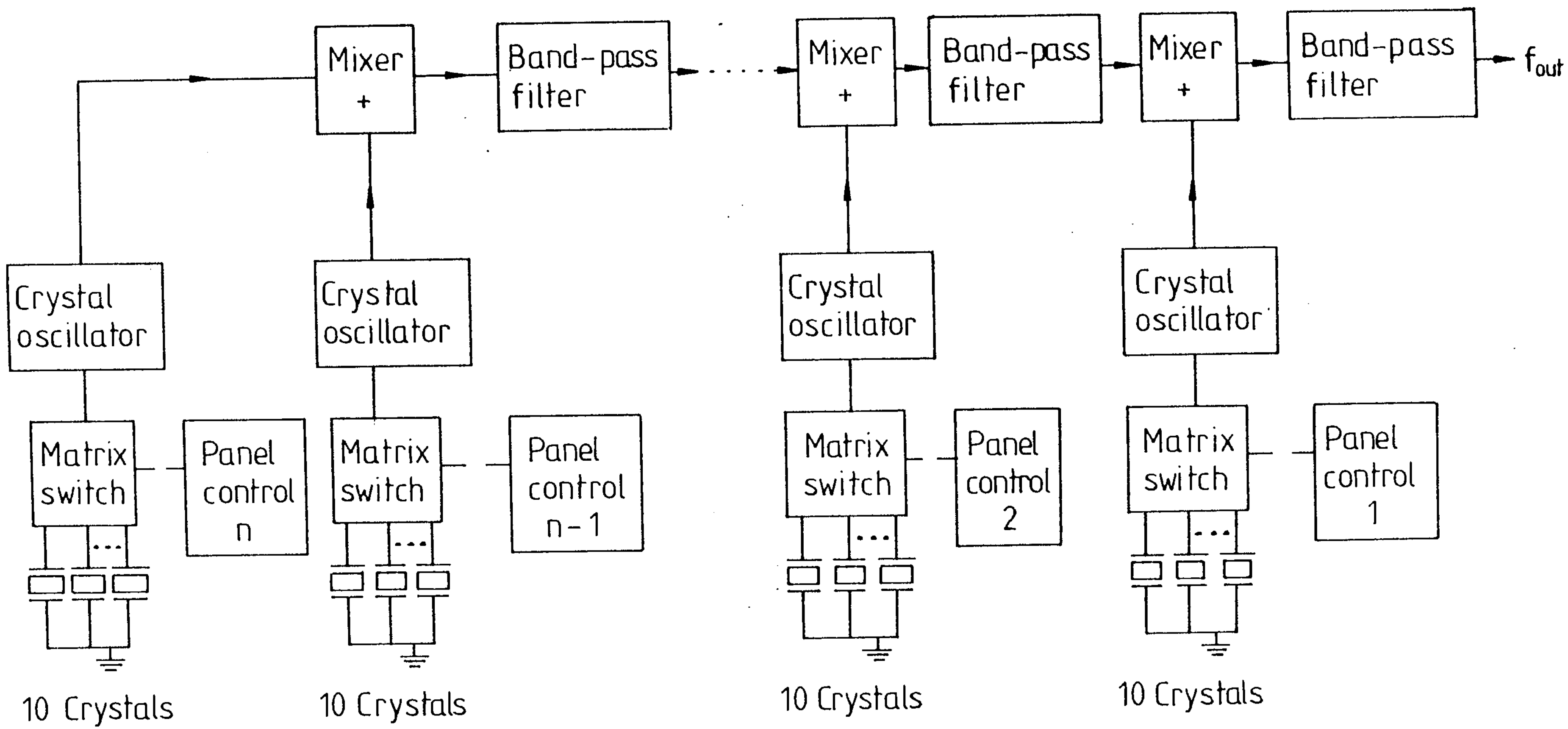


Fig 4.1 Incoherent synthesis.

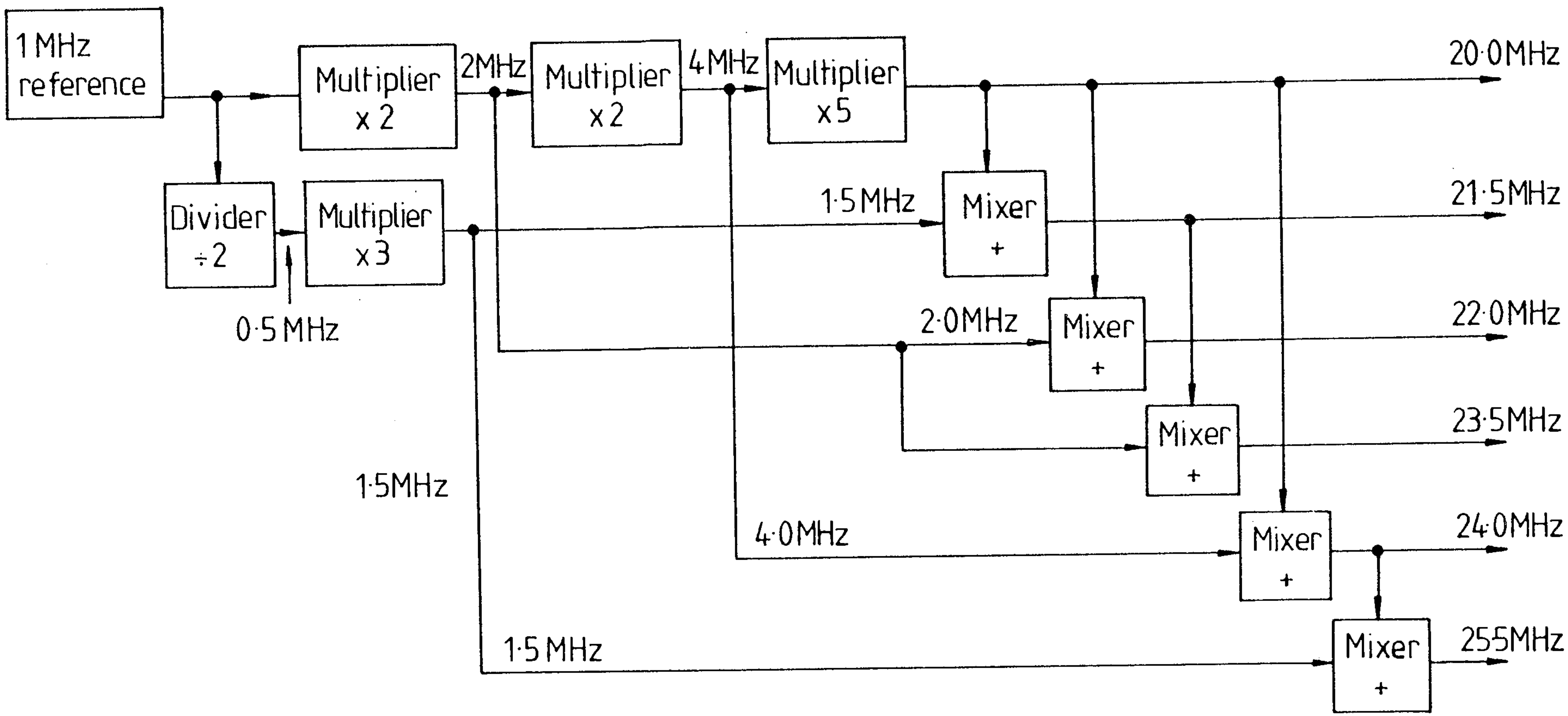


Fig 4.2 Brute-force synthesis: an example.

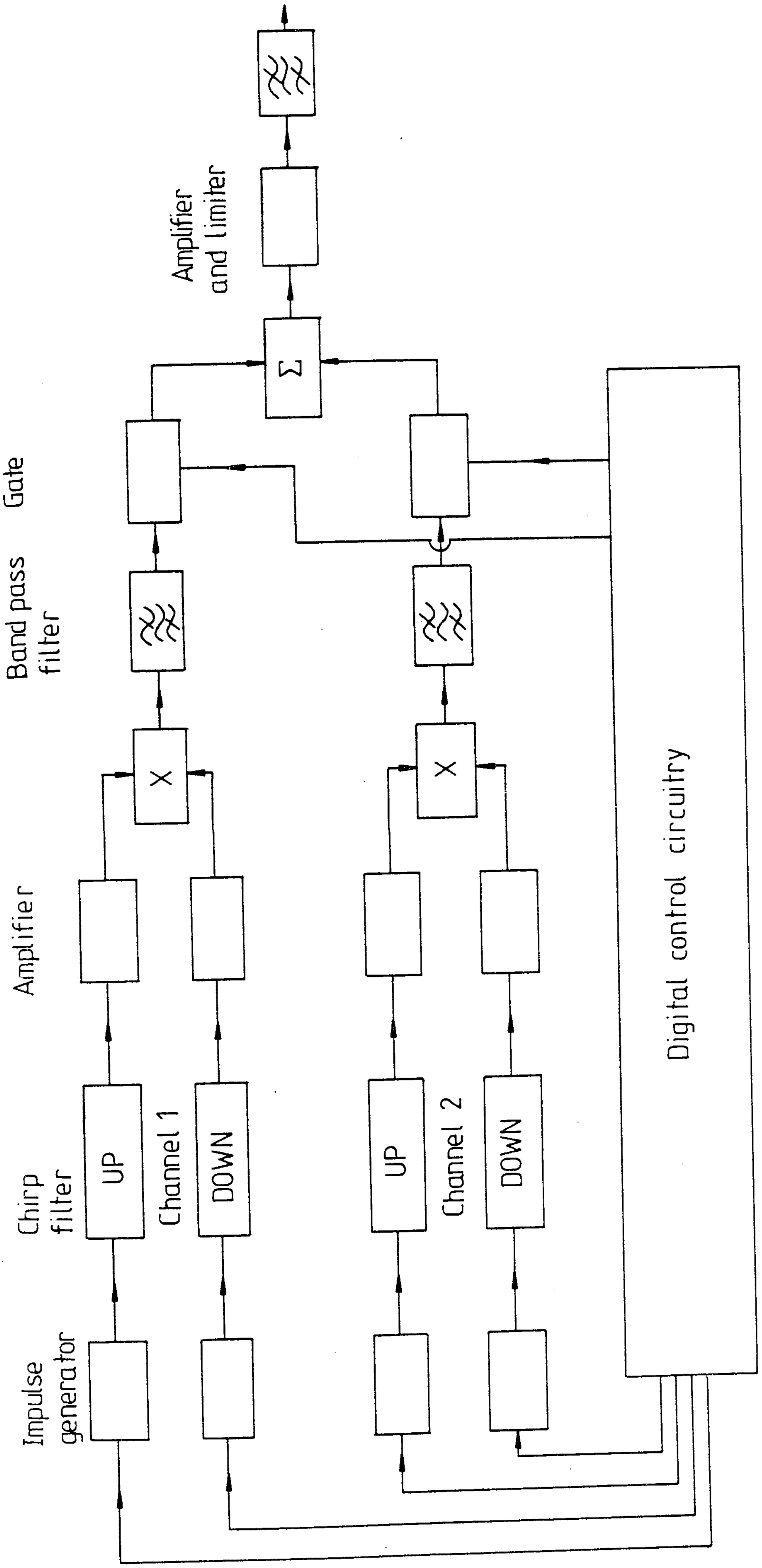


Fig 4.3 Frequency hop synthesizer based on chirp mixing.

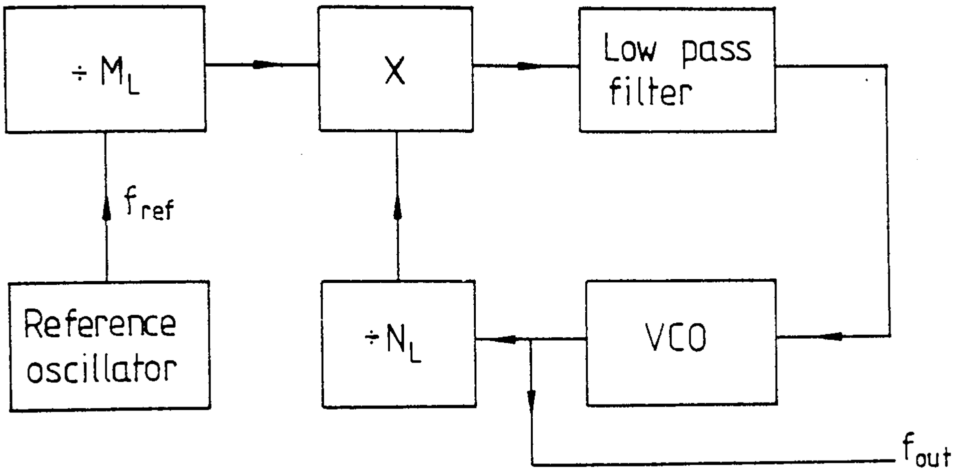


Fig 4.4 Basic indirect frequency synthesizer

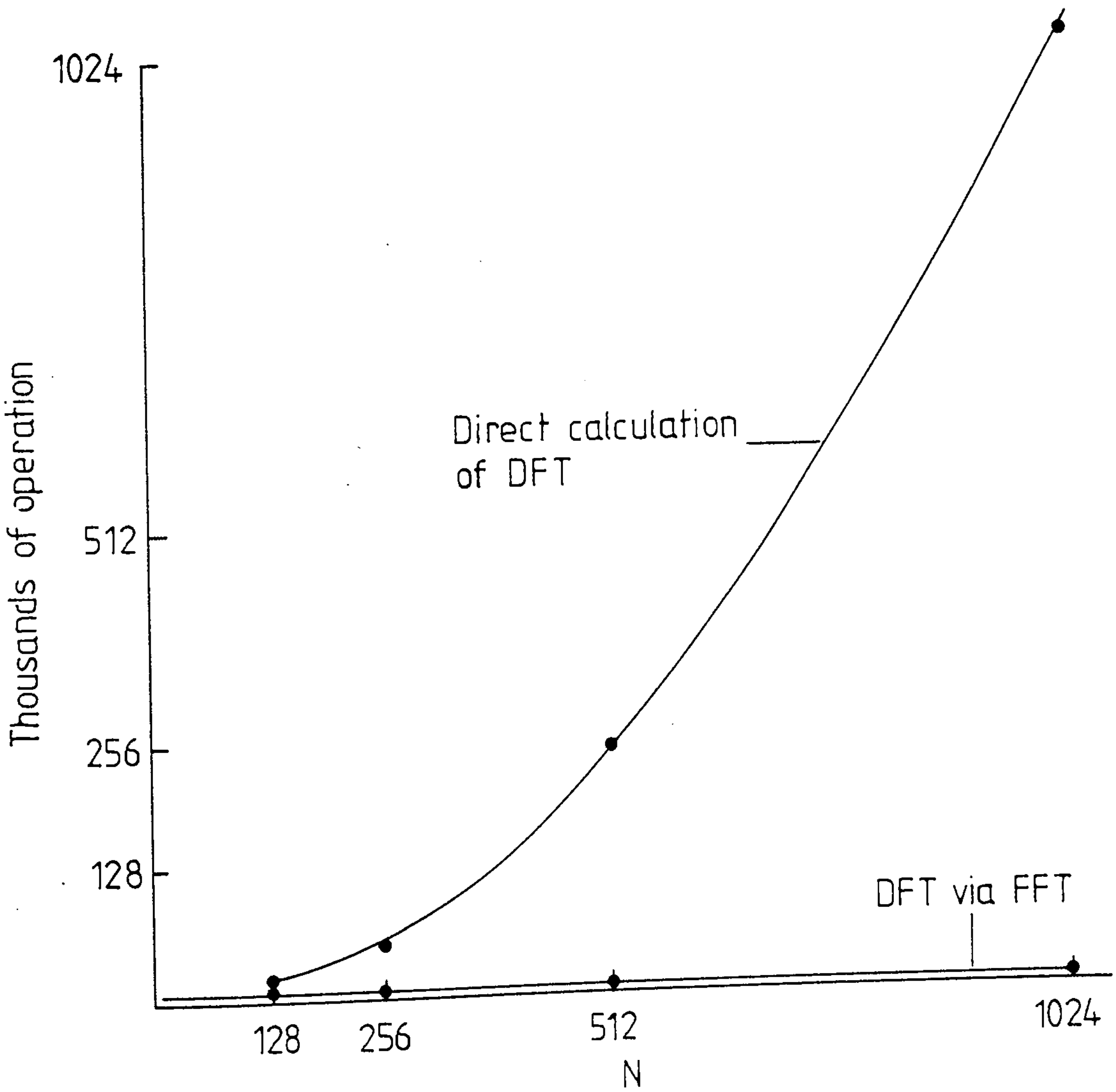


Fig 4.5 The number of operation required for computing DFT directly and by using FFT.

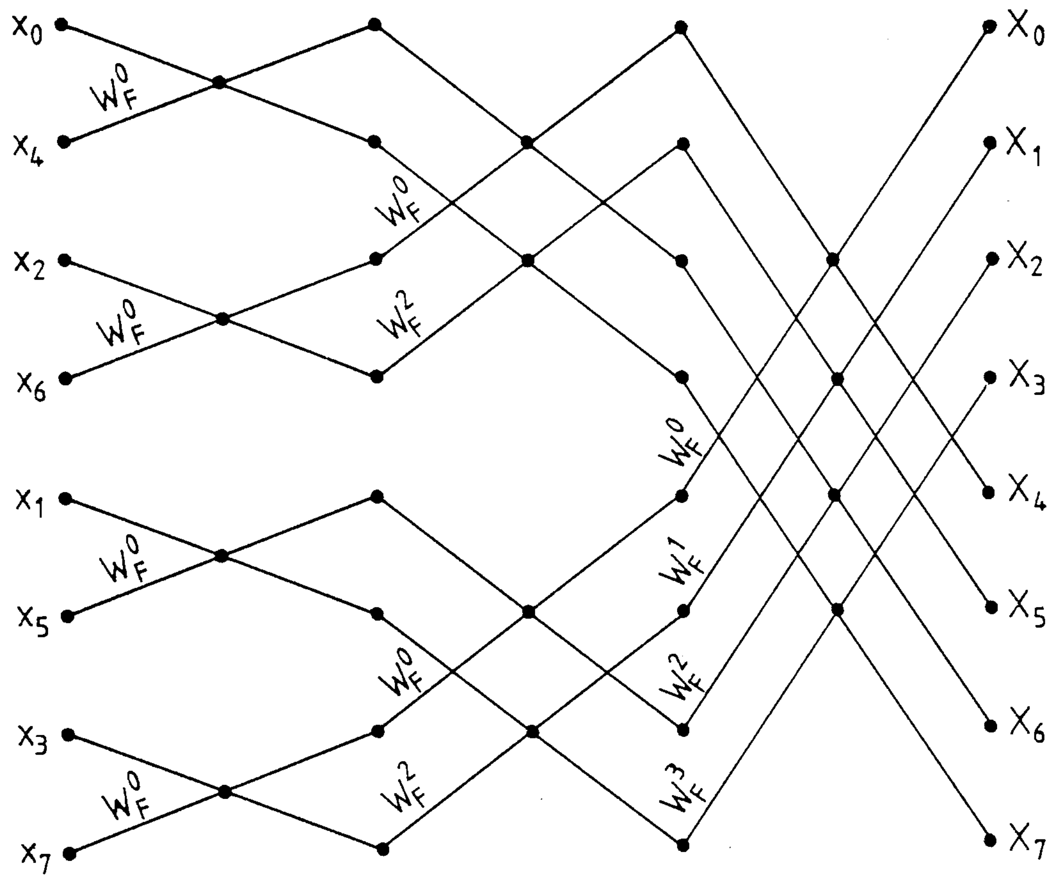


Fig 4.6 Complete eight point decimation in time FFT.

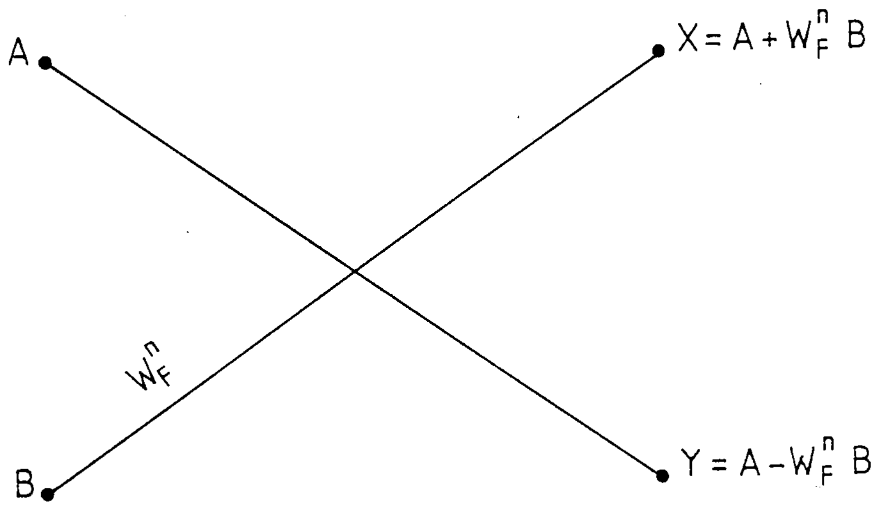


Fig 4.7 Butterfly for decimation in time.

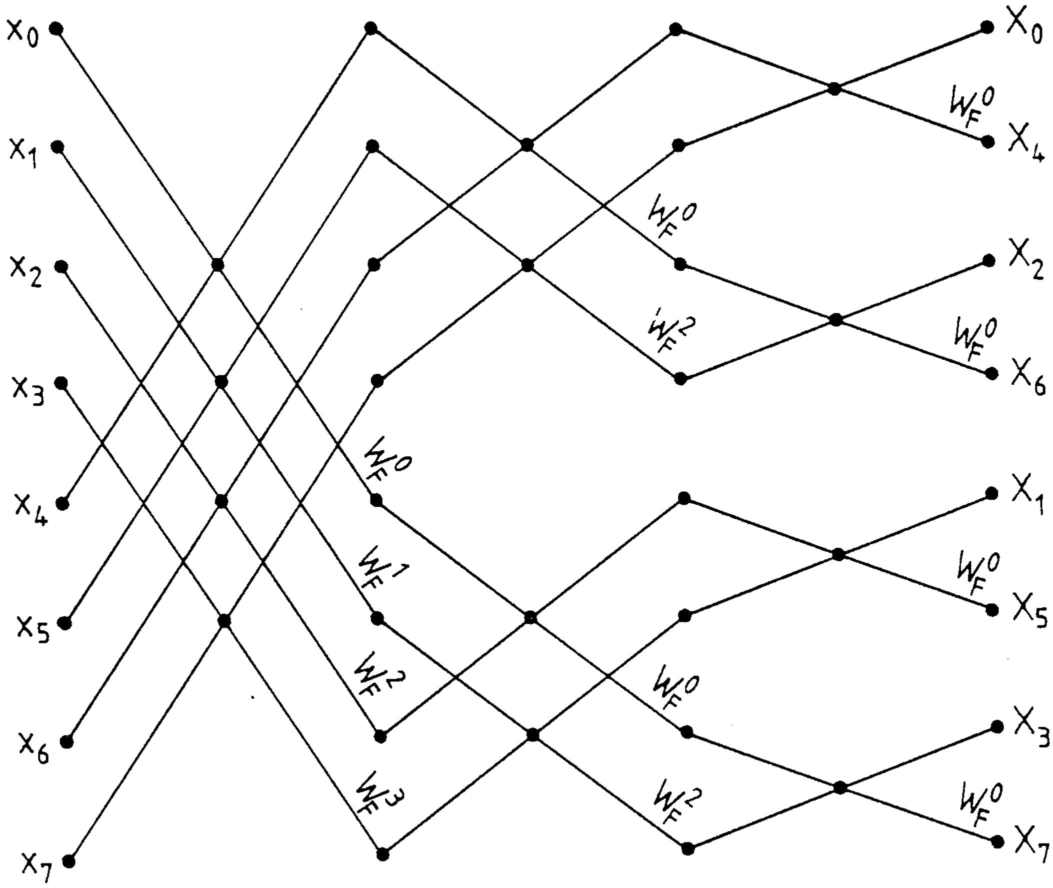


Fig 4.8 Complete eight point decimation in frequency FFT

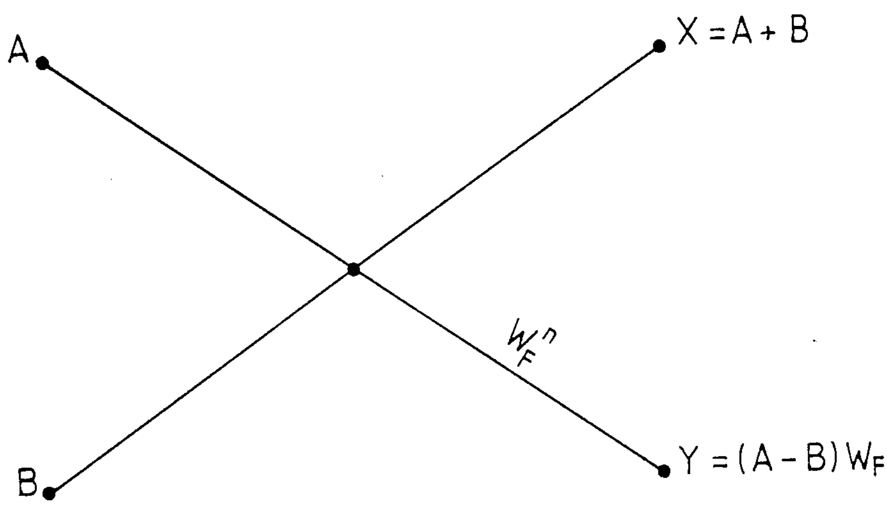


Fig 4.9 Butterfly for decimation in frequency

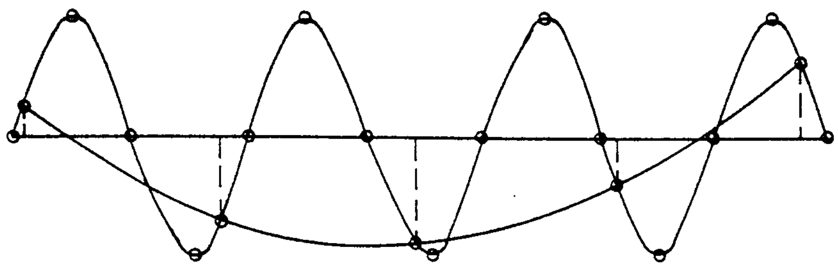


Fig 4.10 A high frequency impersonating a low frequency.

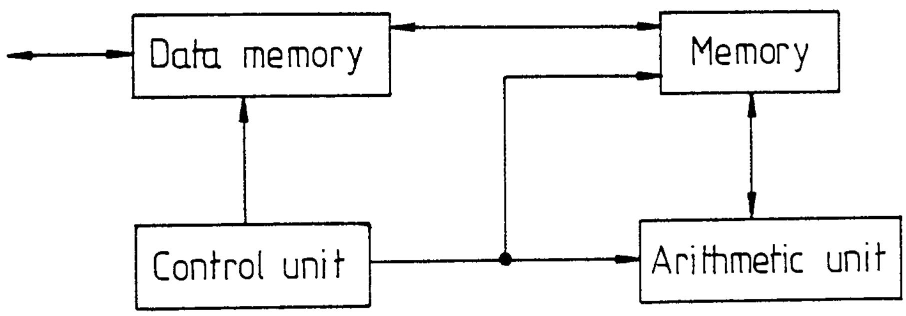


Fig 4.11 The sequential FFT processor.

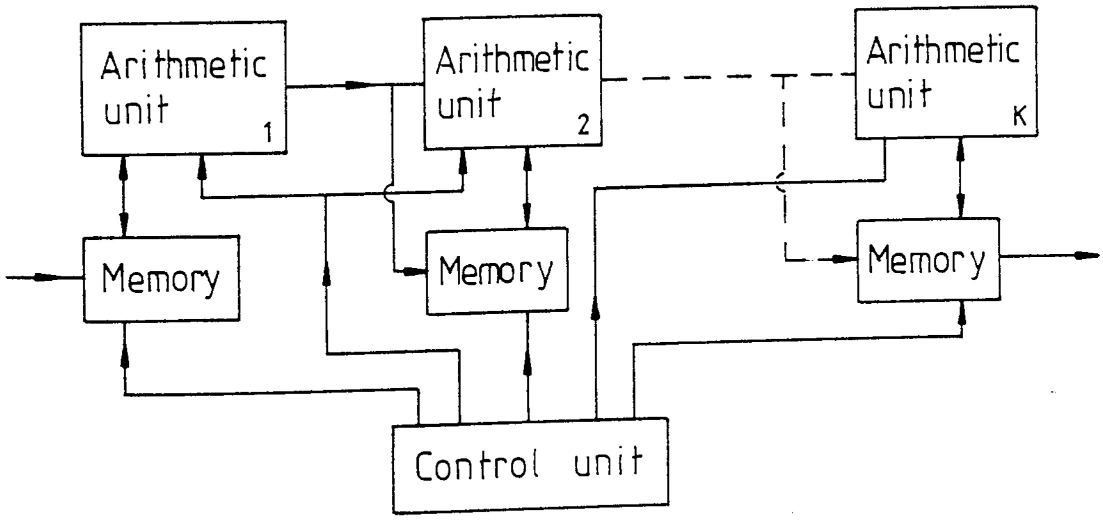


Fig 4.12 The cascaded FFT processor.

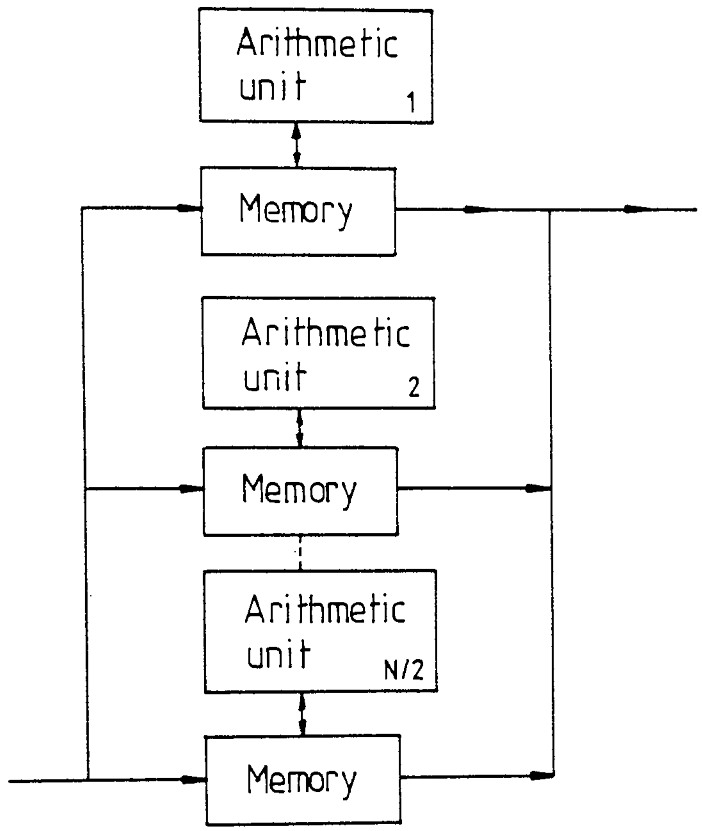


Fig 4.13 The parallel iterative FFT processor.

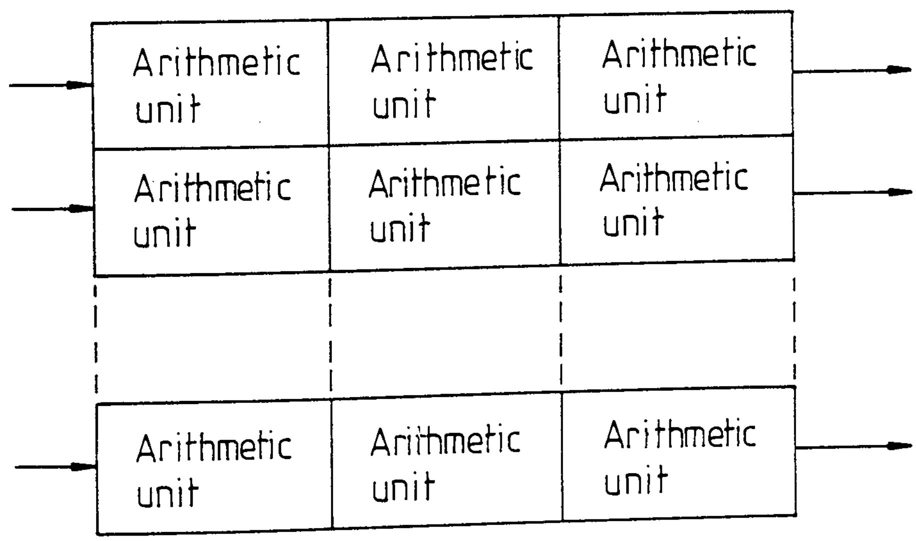


Fig 4.14 The pipeline FFT processor

CHAPTER FIVE

IMPLEMENTATION OF THE FH-MFSK EXPERIMENTAL SYSTEM BASED ON MICROPROCESSOR

IMPLEMENTATION OF THE FH-MFSK EXPERIMENTAL SYSTEM BASED ON A MICROPROCESSOR

5.1 INTRODUCTION

With the advancement of semi-conductor technology the industry now is able to produce LSI chips at lower cost. Microprocessors are a revolution in electronics and they appear in almost everything from domestic appliances to automobile control systems.

In this research, the microprocessor was used to implement the FH-MFSK experimental system without recourse to actual, and often expensive, hardware. Software is not only less expensive, it also offers several other very important advantages over hardware. It is more flexible, it is easier to implement and more reliable. Together, the hardware and software may be utilised to create a more versatile experimental system. Furthermore, by using a versatile Intel 8085 microprocessor system[27, 68], many tasks can be tackled and the development cost can be shared by using the microprocessor for several applications.

In this chapter the information presented in the previous two chapters is used as a foundation for the implementation of an FH-MFSK system. Figure 5.1 illustrates the general layout of the experimental system which is supervised by the Intel 8085. FH-MFSK signals are generated using a set of data samples for a sinusoidal waveform stored in memory. These signals will be transmitted through a channel characterized by A.W.G.N. The noise samples are obtained from a gaussian noise generator and digitized using an analogue to digital (A/D) converter. The received signals are then analyzed using a FFT and the transmitted digital message is determined from the analyzed spectrum using a majority logic decoder. An experimental error-rate model for the FH-MFSK detection process was implemented utilising the experimental system. This error-rate model is fleixble enough to use it to investigate the effects of different parameters on the performance of the FH-MFSK system. A time delay model was also implemented which could be operated in conjunction with the error-rate model.

The software required for this system may be considered as a number of groups modules, each group performing a certain function for generation of FH-MFSK signals or for spectral analysis using FFT or for decoding. The module of groups are linked together to form a complete program. Software was written using the Intel 8085 instruction set [27,28,50].

5.2 FH-MFSK TRANSMITTER

The FH-MFSK transmitter can be considered as a generator operating in the following way. Every T seconds K message bits are loaded into a shift register. Thus, by using an address generator at the transmitter, many users can be multiplexed in order to share simultaneously the same spectral band. The address generator generates a set of Einarsson finite-field addresses along L chips for each user, as mentioned in section 3.5.3.2. The address set can be obtained according to equation (3.11) for user 1. This set of addresses was converted into binary format as shown in Table 3.1.

These binary addresses are converted into hexadecimal format in order to store in the random-access-memory (RAM) of the microprocessor system. The transmitted sequence is formed according to equation (3.1) by adding a K bit message to a K bit address word, during τ seconds, for L times during T seconds ($T=L\tau$). Then during each T sec., L frequencies will be generated sequentially by the tone generator according to the transmitted sequence.

The strategy employed in generating tones of FH-MFSK signals by the use of a microprocessor will depend on the relationship between the rate at which FH-MFSK samples need to be generated, the instruction cycle time and the number of instructions which are required per sample.

From the above it can be concluded that a fast instruction-time microprocessor is required to generate FH-MFSK.

There are two techniques for generating tones in FH-MFSK experimental systems using microprocessors based on the look-up table method. One of the methods is to store in the memory of the processor sets of data which correspond to samples from 32 sinusoidal waveforms. Using this method, each set can be read independently by the microprocessor at the output, depending on the frequency corresponding to the transmitted data. The second method considered is to store only one set of data samples from a sinusoidal waveform. Then, by varying the speed at which these samples are read at the output, the desired modulating frequencies can be produced. Thus, by reading the samples at a faster rate, a higher frequency is produced and vice-versa.

Comparing the above two methods, it is obvious that the first one needs a large memory size together with a simple program to handle the process. The second method needs less memory size but the software is more complex than in the first method and requires a longer time to execute the program. Only low frequencies can therefore be generated. For this reason the first method was selected for use in the system.

To obtain the samples needed to produce the 32 frequencies required, a full cycle of a cosine wave was accurately constructed. Samples to be generated can be represented as a function of the sampling interval as follows:-

$$X(nT_s) = A \cos(2\pi f_m nT_s) \tag{5.1}$$

where

f_m is the m'th generated frequency

T_s is the sampling frequency interval

n is an integer number.

Each cosine wave was sampled at sampling frequency, (f_s), such that

$$f_s = 2\times 32\times f_o \tag{5.2}$$

where

f_o is the fundamental frequency.

The selection of the sampling frequency is a compromise between providing sufficient samples per cycle to reduce sampling distortion to acceptable limits and to provide sufficient time for the microprocessor to execute its instructions within its acceptable clock limit. Each sample was represented by eight binary digits. The sample amplitude was measured, scaled appropriately into a range between 63 and -63 decimal, converted into hexadecimal format and stored in the RAM of the microprocessor system. Figures 5.2 and 5.3 illustrate the flow diagrams of the FH-MFSK transmitter.

5.3 FH-MFSK RECEIVER

FH-MFSK recieved data is coming in at the same rate as the transmitted data, but in analogue form. The FH-MFSK receiver operation is the exact opposite operation to that performed by the transmitter. At the FH-MFSK receiver, spectral analysis, detection, modulo-2^K subtraction of the address used in the transmitter and a majority logic decoder decision are performed every T seconds. Implementation of an FH-MFSK receiver is discussed in this section.

The processing of the received FH-MFSK signals can be implemented by using a FFT as mentioned in section 4.4. A FFT signal flow graph for $N=2^6$ is given in Fig. 5.4. The experimental system was designed for 32 frequencies. General properties of the signal flow graph are summarised below in order to provide a framework for developing a FFT program:-

(1) It includes sixty-four samples.

(2) It includes six passes.

(3) Number of groups per pass is equal to 2^{OP} where OP is the rank of pass (0,1,2,3,4,5).

(4) Number of butterflies per group per pass is equal to

$$\frac{32}{\text{Number of groups per pass}}.$$

Implementation of the FFT in this work was based on the sequential processor and DIF form mentioned in section 4.5.2, since the sequential processor is a simpler processor and suitable for a non real-time system. Also the DIF does not require re-ordering of the input samples. All $(\frac{N}{2}\log_2 N)$ butterflies in the signal flow graph are given for one set of input samples and are executed sequentially.

The mathematical expression of the FFT algorithm involves the rotation factors W_F. These rotation factors can either by calculated directly in the same FFT program or can be more quickly obtained from a look-up table which provides the W_F values. The look-up table method was selected since it is quicker and W_F values can be calculated once for each pass and stored in the RAM of the microprocessor system.

In order to simplify the implementation of the FFT program, the multiplications are performed by integer arithmetic. Firstly, the values of W_F are multiplied by 64, and then the final product is divided by 64. However, this method degrades the accuracy of the result. FFT flow diagrams are shown in Figs. 5.5, 5.6, 5.7 and 5.8. In these diagrams the following nomenclature is used:

GP = Number of groups per pass in FFT flow process

KG = Number of samples per group/2 in FFT flow process

BT = Number of butterflies per group in FFT flow process.

The received signal decision for each τ seconds after spectral analysis by the FFT is based upon a comparison of the amplitudes of the FFT output for each frequency against a threshold level. A detector is used to carry out this operation.

A received spectrum matrix can be formed after spectral analysis by the FFT and detection for L received words. The received spectrum matrix is a frequency-time (F-T) matrix containing Q rows and L columns. Each row represents one frequency. A level value can be set for each row which is equal to the level required to generate one frequency from Q available frequencies in the FH-MFSK transmitter. Thus the F-T matrix can be considered also as a level matrix, since the row does contain information about frequency and the corresponding level for that frequency. Also, each column corresponds to one transmitted word during τ seconds ($L\tau=T$).

Modulo-2^K subtraction is used to subtract the address sequence from the level matrix for every τ according to equation (3.2). Thus, a detection matrix is formed by subtracting the address sequence from the level matrix. The address sequence in the receiver is generated by an address generator which is similar to that used in the transmitter. In the ideal case of operation, where the operation is not contaminated by either interference or noise, a single row formed belongs to one transmitted message (X_{mt}). Corrupted operation can cause a detection matrix to have either more than one row or no complete row. To allow for this possibility, a majority logic decision decoder is used as mentioned in section 3.3.

The detection matrix and majority logic decoder are implemented in the experimental system by providing a counter for each frequency to count the number of times the identical frequency (Xf_r) is received. Each one of these counters is equivalent to one row in the detection matrix. The received data message is associated with the counter containing the highest count. Signal received matrix, detection matrix and majority logic decoder flow diagrams are given in Figures 5.9 and 5.10.

5.4 GAUSSIAN NOISE DIGITIZATION

An A/D converter is used to digitize the gaussian noise. Gaussian noise is used for two purposes. The first is to generate A.W.G.N. which is added to the transmitted data signal. The second is to generate a random transmitted data sequence to test the FH-MFSK system. Gaussian noise was obtained from an HP3772A gaussian noise generator. The A/D converter circuit is microprocessor controlled and is identified to the processor by input and output instructions. Figure 5.11 is a flow diagram illustrating the insertion of A.W.G.N. and generation of the transmitted message. The A/D converter is used in conjunction with a Sample and Hold (S/H) circuit and an ISBC 80/30 board. The ISBC 80/30 board will be discussed in the next section. The Sample and Hold circuit is used to hold the noise signal steady at a particular value in order to convert it with minimum error.

The processor sends a start conversion message to the A/D converter to start the conversion process. When the conversion is complete, a binary word corresponding to this quantized level is loaded into the output latches and an end of conversion logic level appears. The output latches hold this data valid until a new conversion is completed and new data is loaded into the latches.

5.5 ISBC 80/30 SINGLE BOARD COMPUTER

The Intel ISBC 80/30 single board computer is used to implement the FH-MFSK system in this investigation[45]. This board contains an Intel 8085 processor; 16 K bytes of RAM memory; (8259A) Programmable Interrupt Controller (PIC); (8255) Programmable Parallel Interface (PPI); Two (8251) Programmable Interval Timers (PIT) and 24 parallel Input/Output lines.

The ISBC 80/30 board is interfaced with a keyboard and VDU by the board ISBC 517. The ISBC 80/30 board, ISBC 517 board, keyboard and VDU forms one work station.

Each work station is networked to an Intellec 231 microprocessor development system with floppy-disc printer and programmer. The Intellec microprocessor development system is used for assembling, linking and relocating user programs, as well as for saving programs and obtaining listing of programs.

The ISBC 80/30 contains on-board EPROM (2732) containing a monitor and editor to facilitate user software development at the work station.

5.6 THE ERROR-RATE MODEL

The error-rate (i.e. the probability of error as a function of signal-to-noise ratio) model is structured to simulate a FH-MFSK system utilizing the experimental system described earlier. This model is designed to compute the probability of error for a given input S/N when FH-MFSK signals are received with A.W.G.N. This error rate model has facilities for the investigation of the effects of the following parameters of the FH-MFSK system:-

1. Signal-to-noise ratio.
2. Number of message repetitions (L) (i.e. the number of times the same transmitted message is repeated with different address words).
3. Threshold levels.

Different input S/Ns are specified by first storing the same rms values of signal and the different rms values of noise, and secondly, dividing each using software to set a specific S/N. Having set the S/N, the user can operate the error rate model and compute the bit error rate P_b,(74) as follows:-

$$P_b = \frac{\text{Number of Errors}}{2 \times \text{Number of Transmitted Messages}} \qquad (5.2)$$

Figure 5.12 shows a flow diagram of the FH-MFSK error-rate model in which L orthogonal frequencies are sequentially generated, added to noise, spectrum analyzed and compared against the threshold level. The level and detection matrices are then generated and the received message is obtained from the majority logic decoder and errors detected.

Each FH-MFSK frequency is generated in accordance with each transmitted word formed by the addition of the message to one of the address words (L transmitted words for one message). The message is obtained from a gaussian noise generator by the sèlection of a given generated noise sample. New samples of white gaussian noise are added to each sample of FH-MFSK signal after the signal-to-noise ratio has been specified. Each output of the majority logic decoder results from the received and analyzed L transmitted word.

The received message is compared with the transmitted message and the output is fed to an error counter. After the required number of transmitted messages (1000 messages) for each value of signal-to-noise ratio has been processed the probability of error is readily calculated using equation (5.2). This error rate model is implemented for one user.

5.7 THE TIME DELAY MODEL

The time delay model is structured to investigate the failure of the window used in the FFT to start at the arrival time of the received FH-MFSK signals. This model is implemented by delaying the received signal samples a number of sample intervals (N_A), where N_A is the number of delaying samples.

Figure 5.13 shows a flow diagram of the time-delay model in which N_A locations in a shifting RAM area are set to zero value, the remaining samples of the received signal ($N-N_A$) are stored in the shifting RAM area following the N_A zero locations. Samples of received signals in the shifting RAM are transferred to the input FFT data RAM area in order to find the spectrum of the delayed received signal.

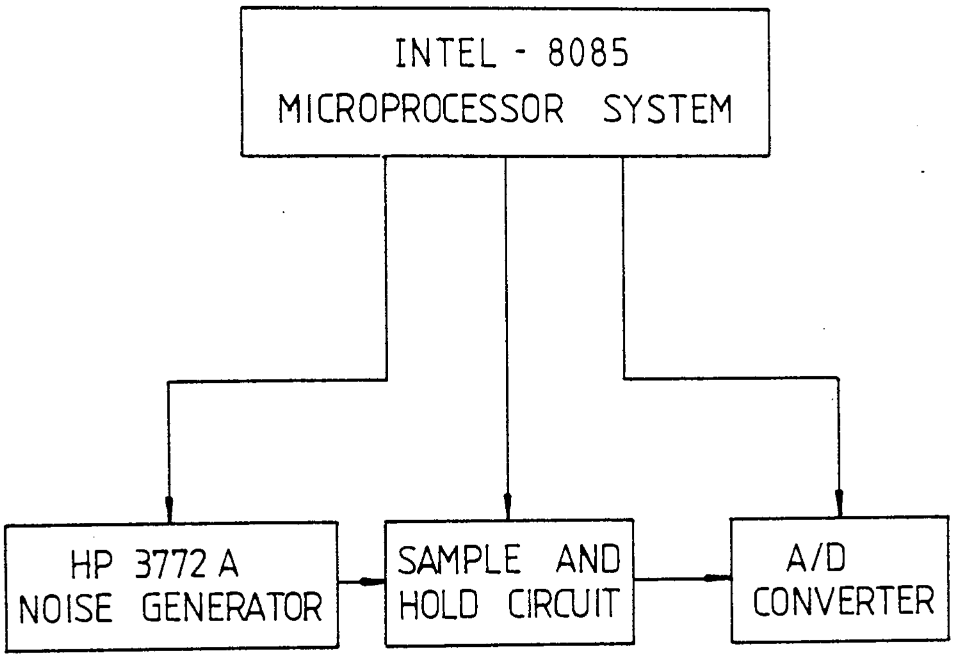


Fig 5.1 General layout of experimental system

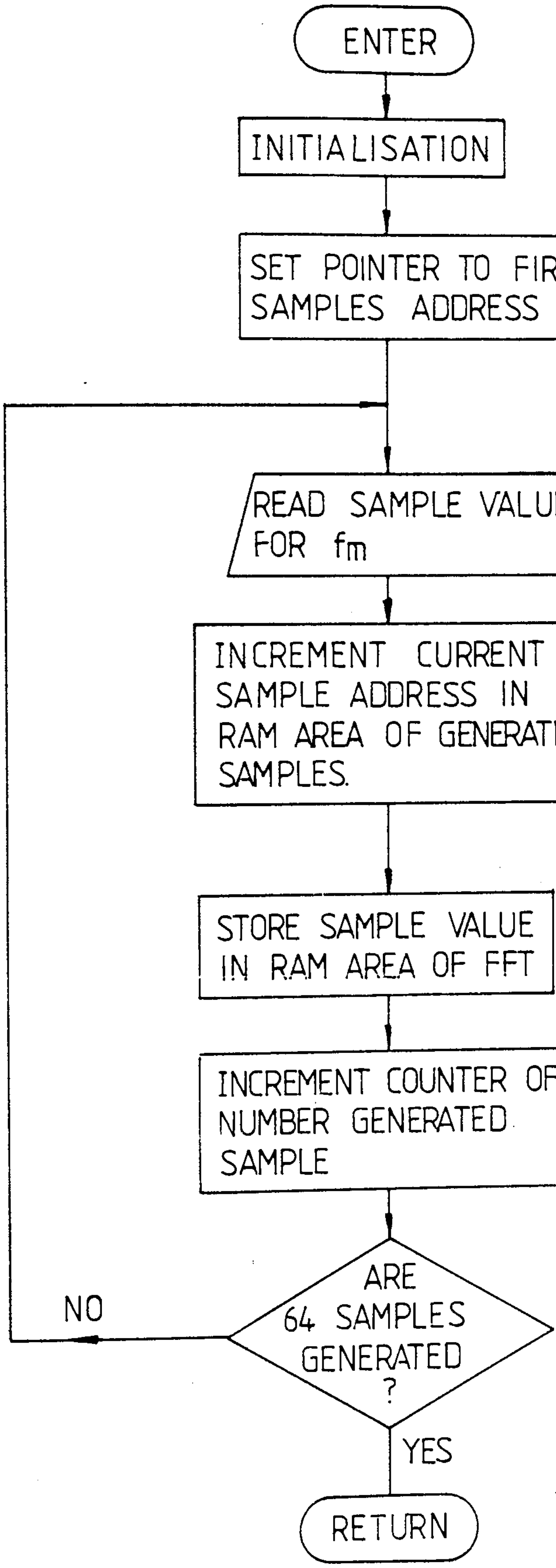


Fig 5.2 Tone generator flow diagram.

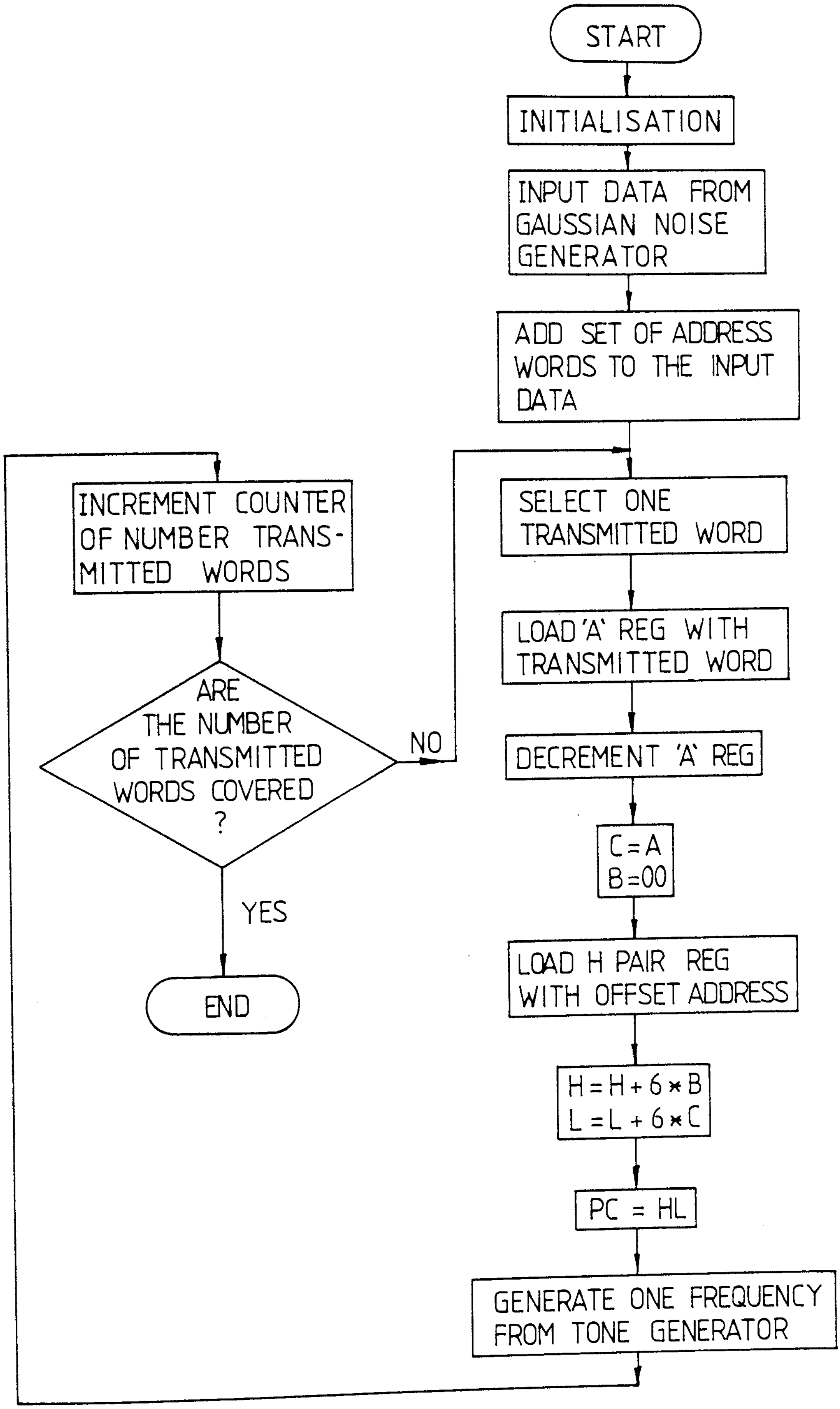


Fig 5.3 FH-MFSK transmitter flow diagram.

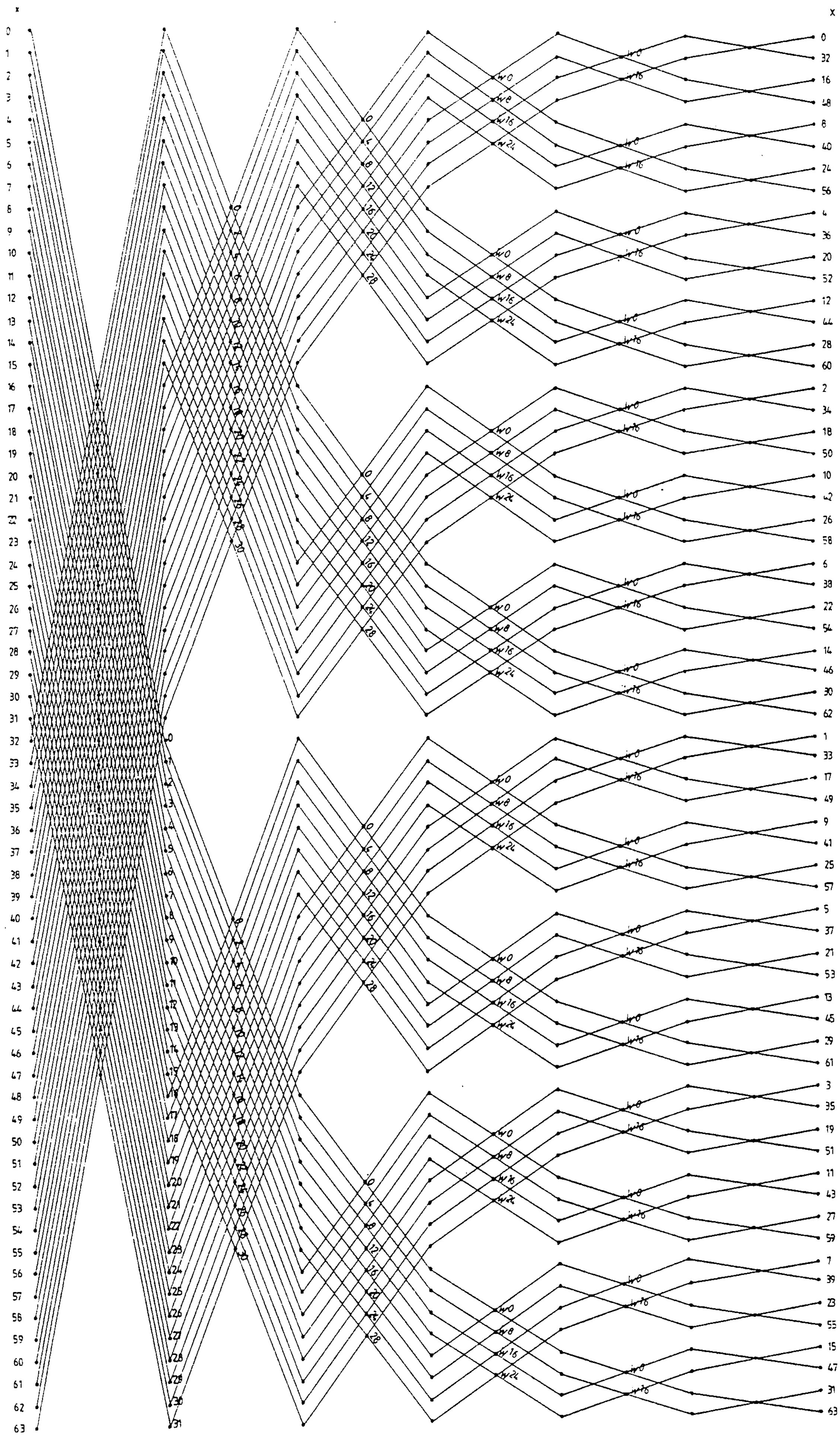


Fig 5.4 Complete 64 point, in-place, decimation in frequency FFT.

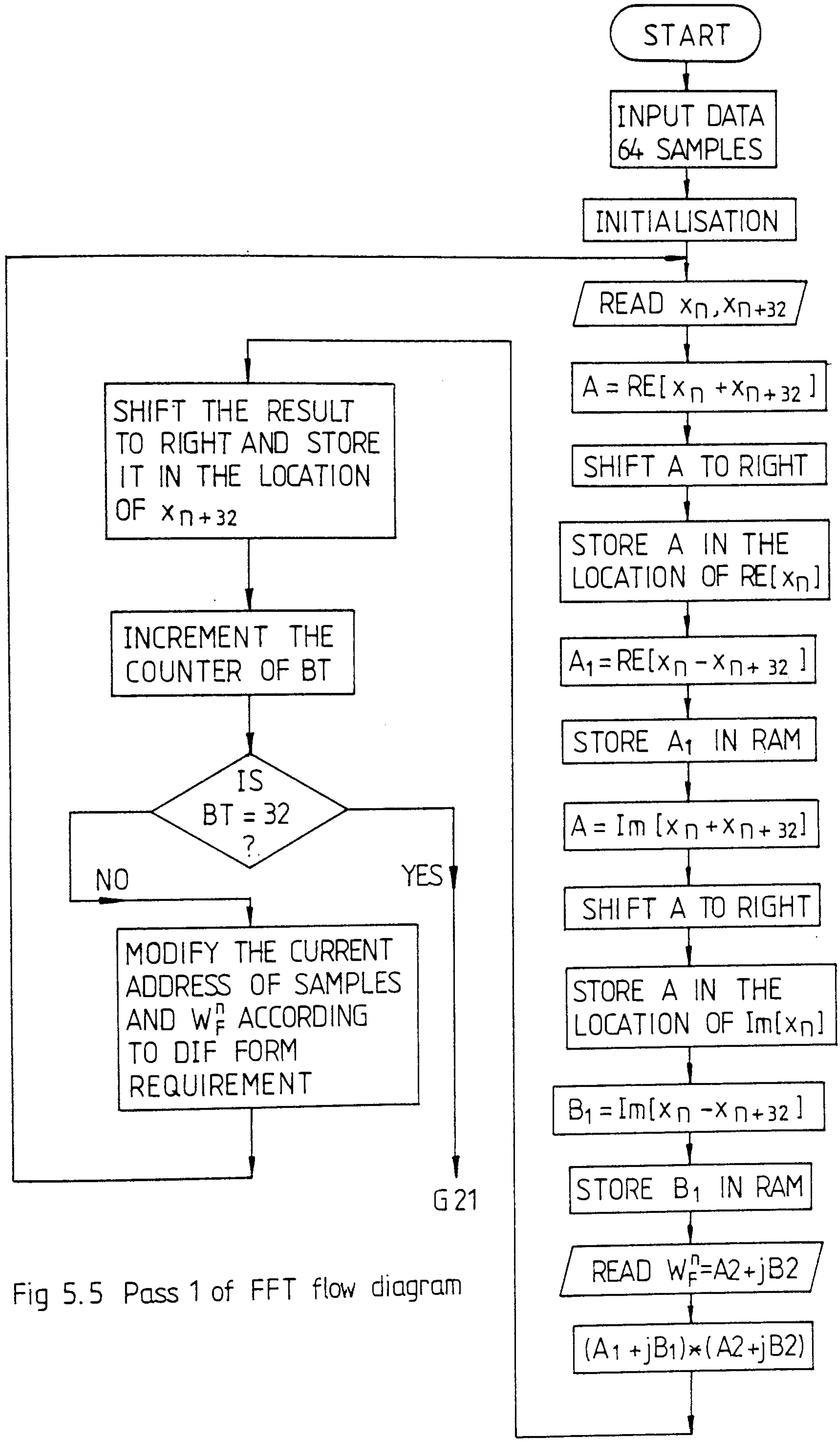


Fig 5.5 Pass 1 of FFT flow diagram

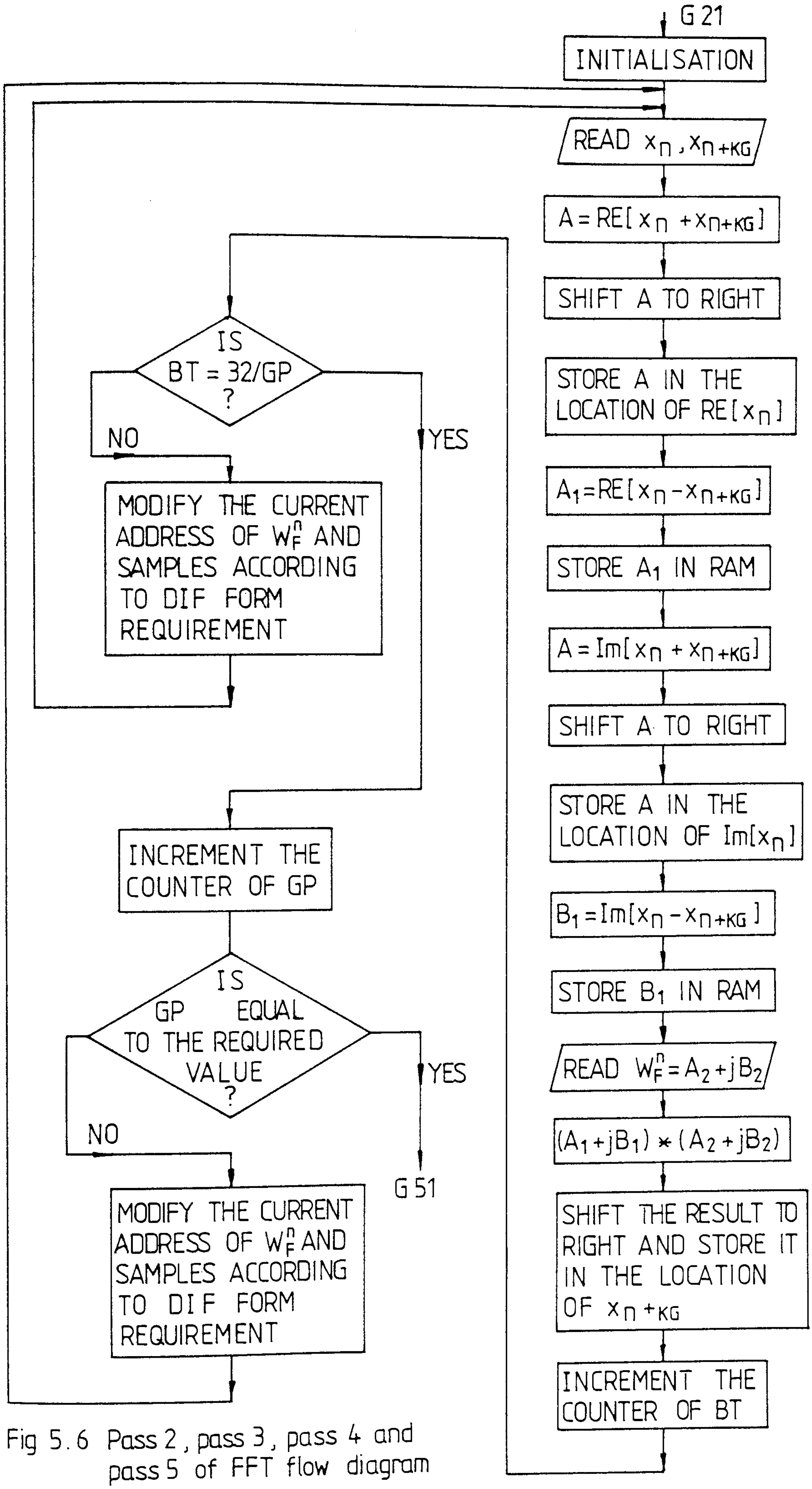


Fig 5.6 Pass 2, pass 3, pass 4 and pass 5 of FFT flow diagram

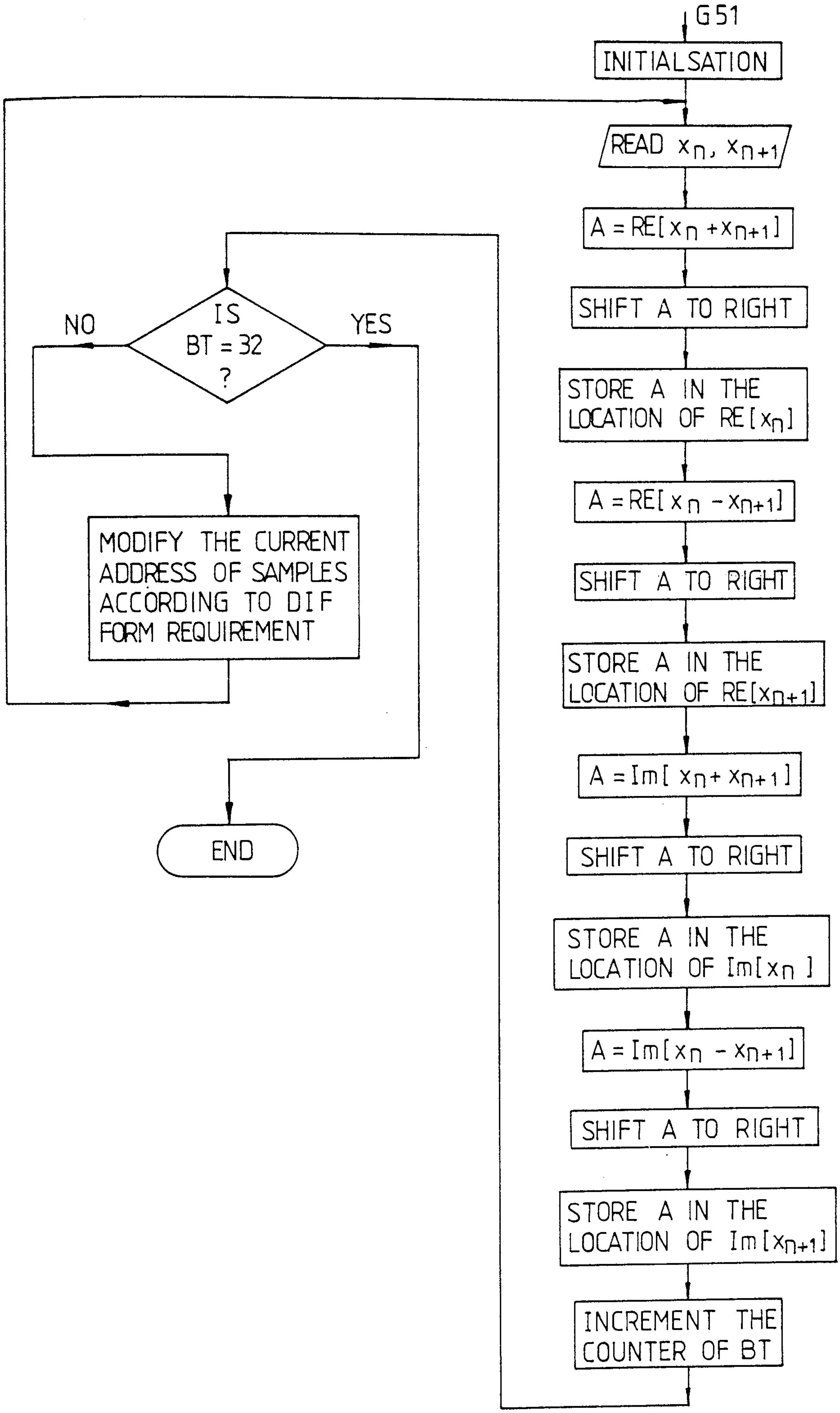


Fig 5.7 Pass 6 of FFT flow diagram.

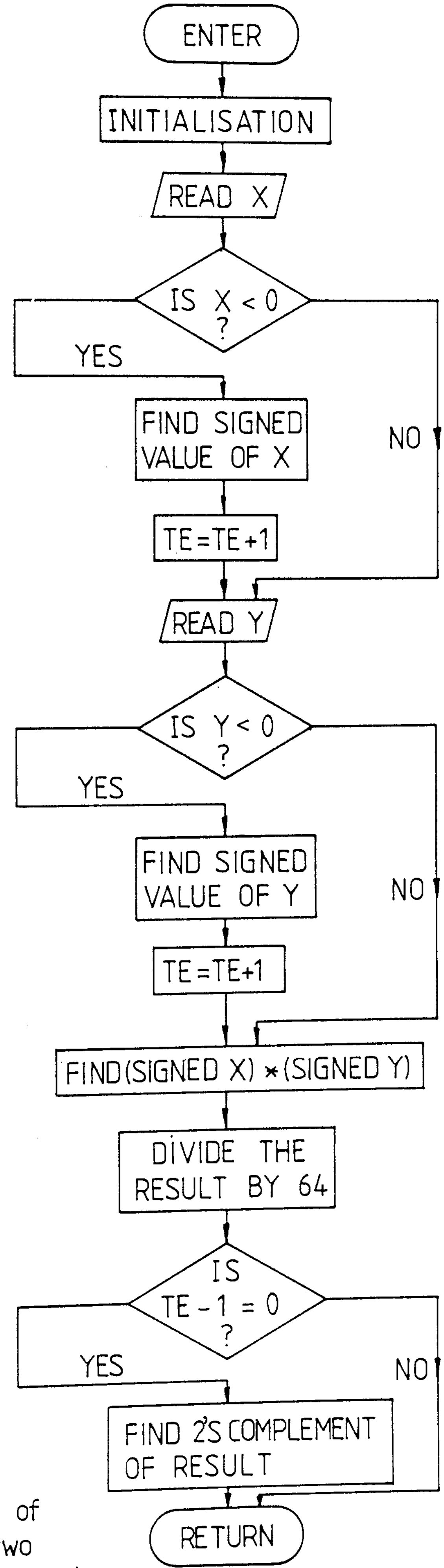


Fig 5.8 Flow diagram of multiplication two numbers (-ve or +ve) and divide the result by 64

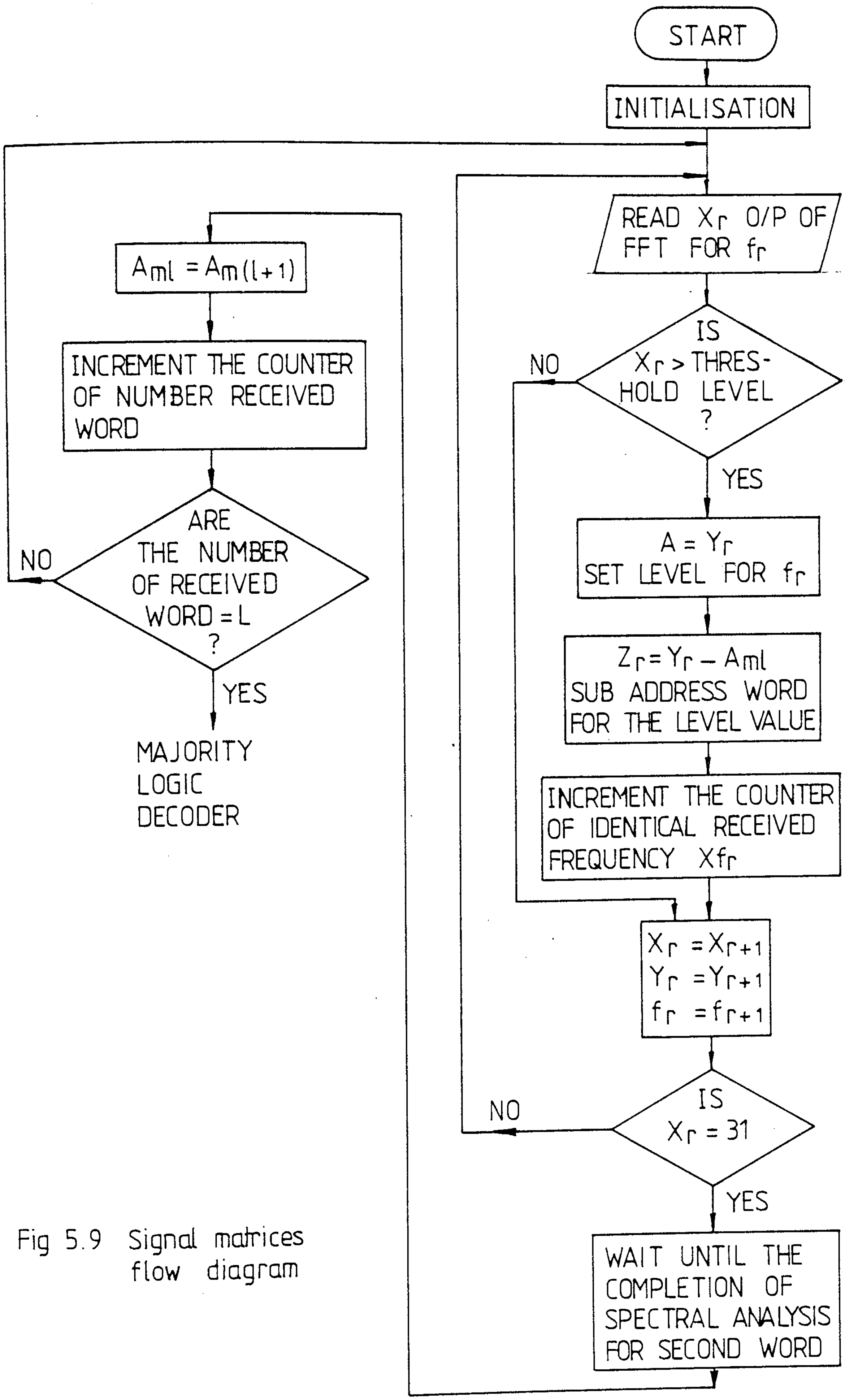


Fig 5.9 Signal matrices flow diagram

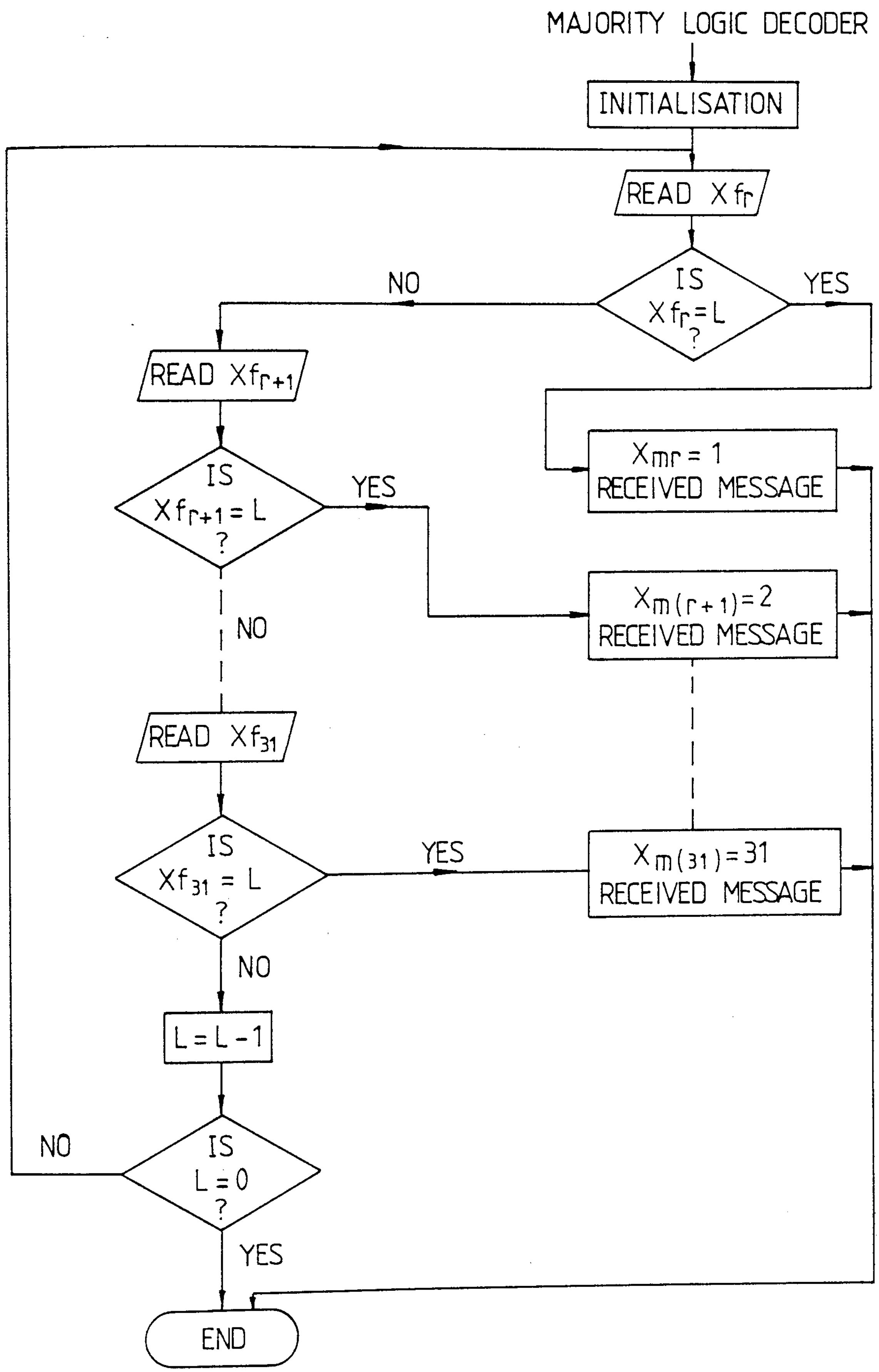


Fig 5.10 Majority logic decoder flow diagram

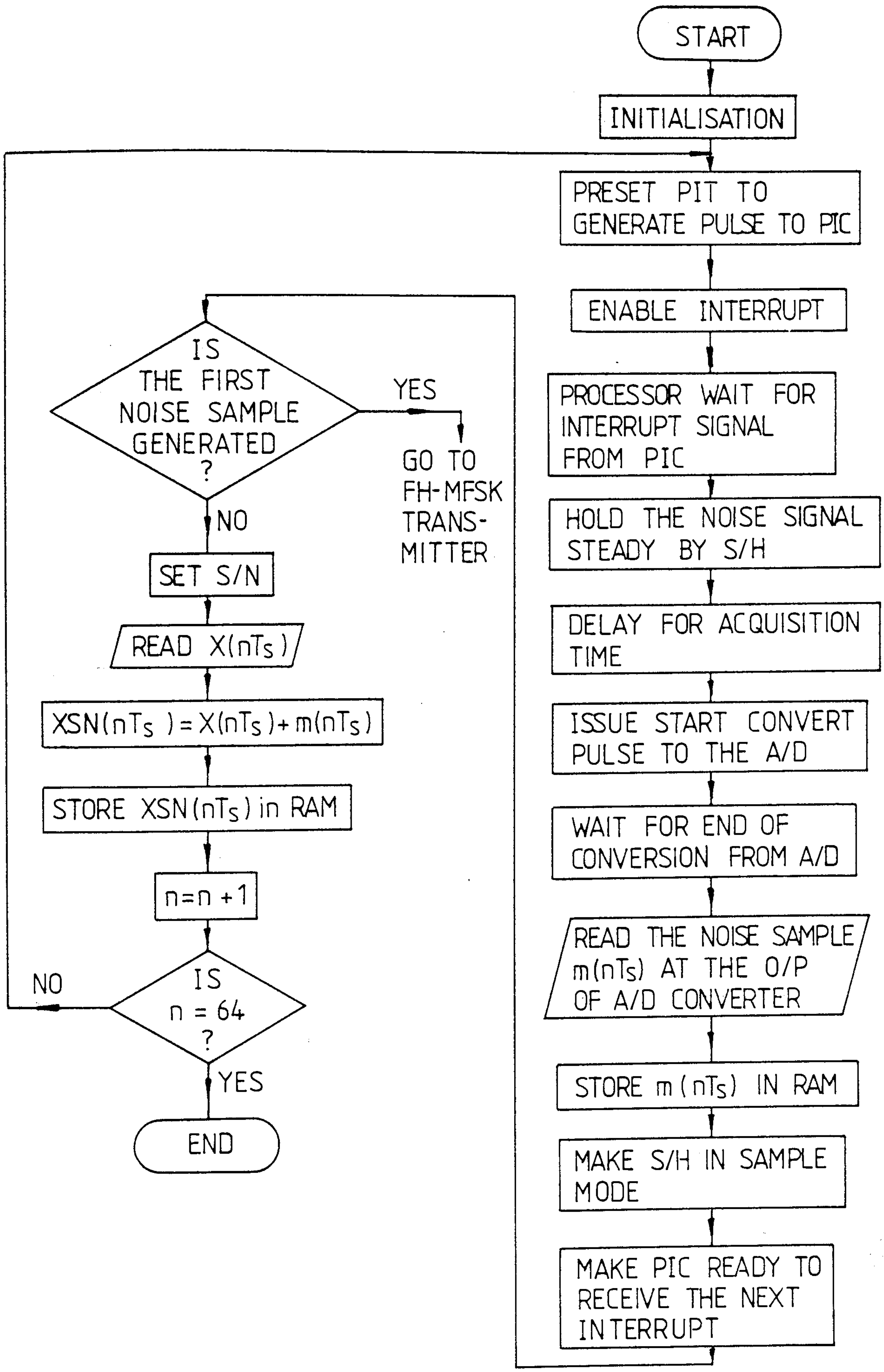


Fig 5.11 Inserting A.W.G.N. and generating transmitted message flow diagram.

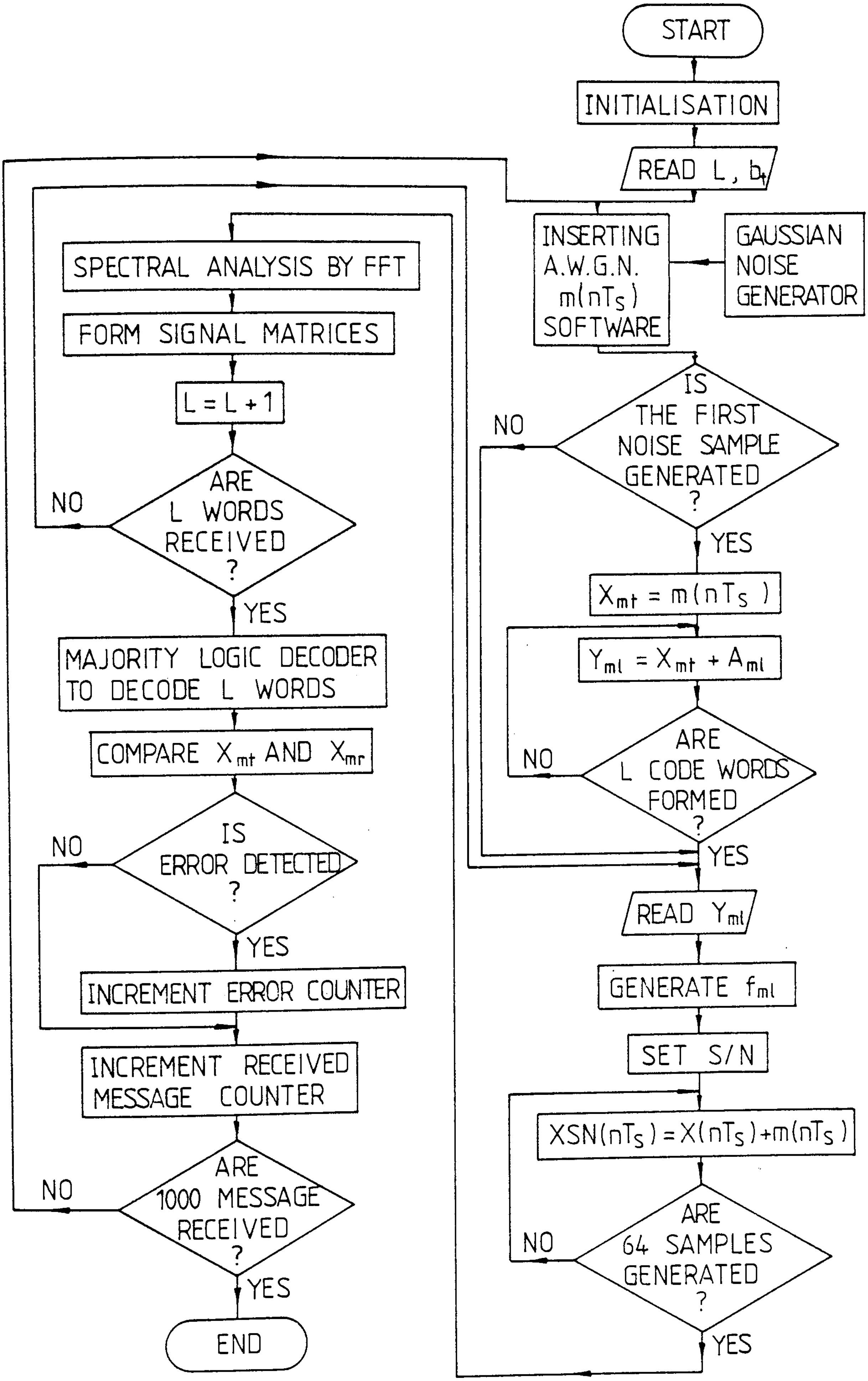


Fig 5.12 Error-rate model flow diagram.

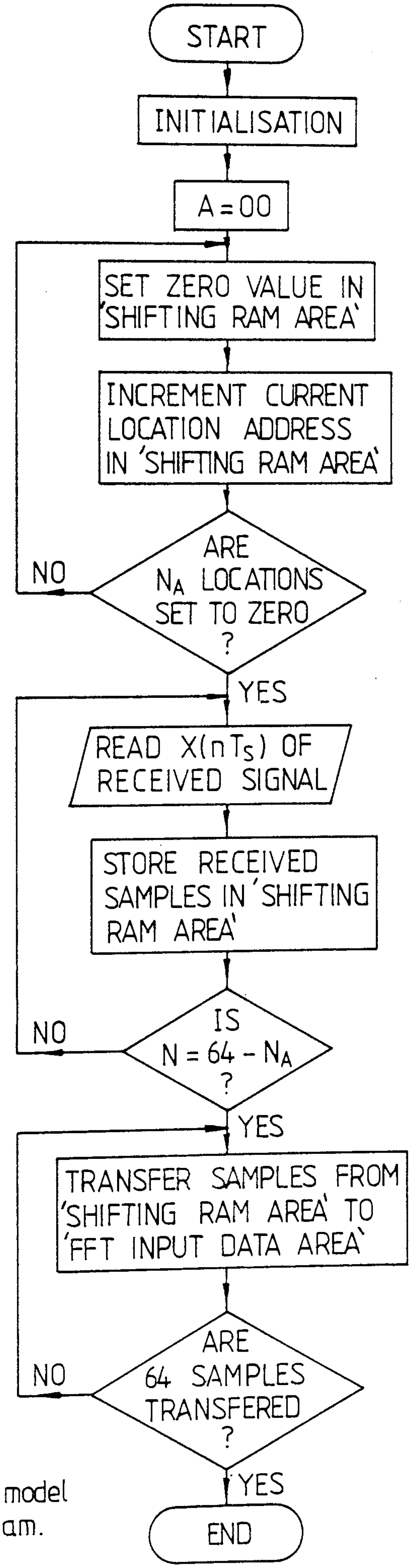


Fig 5.13 Time delay model flow diagram.

CHAPTER SIX

MEASUREMENTS AND RESULTS

MEASUREMENTS AND RESULTS

6.1 INTRODUCTION

The results of measurements obtained from the particular implementation of the FH-MFSK system described in the previous chapter are presented here. It is appropriate first to check the generation of the FH-MFSK signals by the transmitter in the noiseless case. Secondly, to examine experimentally and theoretically the performance of the FFT with different values of deleted and delayed samples. After the FH-MFSK transmitter and FFT have been tested, then they can be utilised to perform error-rate measurements, using the error-rate model developed in Chapter 5. All measurements were obtained by laboratory simulation of the channel, in terms of gaussian noise impairment. This had the advantage that the performance of the system could be assessed separately for gaussian noise. All experimental results established for the FH-MFSK system are compared with theoretical results and then discussed. The theoretical error-rate results were obtained using the analysis of the FH-MFSK system given in Chapter 3. The same analysis was also used to study the capability of the FH-MFSK system with multiusers.

Moreover, the error-rate results established for the FH-MFSK system are compared with published error-rate results for conventional digital systems. Also the performance of FH-MFSK and FH-DPSK (linear combining) spread spectrum systems are compared theoretically.

It should be stressed that all results presented in this chapter relate to the system as implemented in Chapter 5. The synchronization of the experimental system was achieved by a direct connection between the receiver and the transmitter. This departure from the practical situation was felt to be acceptable in order to simplify the investigation.

6.2 FH-MFSK TRANSMITTER

In order to test the FH-MFSK transmitter during transmission, the system shown in Fig. 6.1 was assembled. This makes use of the gaussian noise generator, the analogue to digital converter, the adder, address generator software and the tone generator software which were described earlier in Chapter 5. A gaussian noise generator was used to generate a sequence of samples. Each sample taken can be considered to be a message. The address sequence is generated by the address generator. Each address sequence contains L address words. The message was modulated by the address sequence and the transmitted sequence (L transmitted words) was formed. Each transmitted word caused the tone generator to generate a single frequency. The output samples for each frequency were displayed on the VDU of the laboratory work station. The display of the samples was clear, very accurate and coincided with the required frequency samples to be transmitted.

6.3 FFT MEASUREMENTS

Various measurements were made to study the FFT performance. These measurements are discussed in the following sections. Since all the input signals under investigation are used by the FH-MFSK system, then they should satisfy the following conditions:-

(1) All frequency components should be orthogonal.

(2) Maximum frequency component $\leqslant \frac{1}{2}f_s$, where f_s is the sampling frequency.

The above conditions are necessary for optimum spectral analysis by the FFT.

6.3.1 FFT in Noiseless Condition

To examine the performance of the FFT without any disturbance, the system shown in Fig. 6.2 was assembled. This uses the tone generator software and the FFT software, which were described earlier in Chapter 5. All input samples were generated and stored in RAM using the tone generator.

In the noiseless condition, the FFT programm was run with different input frequencies and amplitudes for each frequency response-measurement. The results of the frequency response measurement of the FFT for different input frequencies were found to be accurate within ±6%. The results also show that the real component of the output spectrum is very sharp for a single input frequency signal. However, the imaginary component is zero, since the input signal is a pure sinusoidal signal. Although this program gives reasonable accuracy, the speed of calculation is rather slow.

6.3.2 FFT in the Presence of A.W.G.N.

To examine the performance of the FFT with A.W.G.N., the system shown in Fig. 6.3 was assembled. This utilises the tone generator software, the gaussian noise generator, the analogue-to-digital converter, the adder and the FFT software which were described earlier in Chapter 5.

The gaussian noise generator was used to generate a sequence of noise samples. Measurements were carried out by adding the noise samples to the signal samples using an adder circuit. All signal samples were generated by the tone generator software. The FFT program was run for a certain frequency in the presence of A.W.G.N. with different values of signal-to-noise ratio for each run. Results were found to have a variable response depending on the signal-to-noise ratio. The characteristics of the frequency response of the FFT with A.W.G.N. can be summarised as follows:-

(1) The real frequency components of the FFT are reflected images about a vertical line at N/2(=32) while the imaginary components are reflected about the horizontal axis.

(2) The unwanted signal components appeared in the output of the FFT when the signal-to-noise ratio was equal to or less than 3.3 dB. This is due to the fact that the FFT is the optimal filter for detecting sinusoidal signal bursts in A.W.G.N.[41].

6.3.3 The Effect of Deleted and Delayed Samples on the FFT Performance

The received signal at the input of the FFT is defined by its signal-to-noise ratio, frequency bandwidth, number of samples and peaking instant. Peaking instant is investigated to study the effect of time delay between received signal and starting instant of FFT window. The effect of signal-to-noise ratio was discussed in the previous section. The frequency bandwidth of the received signals was chosen according to the design of the FH-MFSK system described in Chapter 3. Measurements were carried out using the systems of Fig. 6.2 and Fig. 6.4 to study the effects produced by deleted and delayed samples of the received signal on the FFT performance. The deleted samples test was carried out to study the effect of the number of samples on the FFT performance. The difference between the deleted samples and the delayed samples tests is that in the deleted samples test the input signal is synchronized with the window of the FFT. However, there is no synchronization between the input signal and the window of the FFT in the delayed samples test. The implementation of the two tests is therefore different as will be discussed in the next two sections.

6.3.3.1 Deleted Samples

The deleted samples test was implemented by inserting zeros in the location of the required number of deleted samples after the tone generator had generated the signal. Synchronization between the input signal and the FFT window was provided for this test by inserting zeros at the end of input signals. The FFT program was run for different values of deleted samples for a given frequency, then the test was repeated for different frequencies. The ratio of the output of the FFT to its input without deletion effects is plotted against number of samples. Figures 6.5a, 6.5b, 6.5c and 6.5d show the effect produced by deleting samples.

Experimental results are compared with theoretical results in Figs. 6.5a, 6.5b, 6.5c and 6.5d. The difference between them is seen to be small. The differences are due to the imperfections of the FH-MFSK generator and the FFT calculation. The theoretical results for this test were obtained using the Fourier transform of a sinusoidal signal in the presence of deletion effect, at $f=f_n$, as given by

$$X(f_n) = \frac{AT_o}{2} \left[1 - \frac{N_D}{64}\right] \tag{6.1}$$

where

$X(f_n)$ is the Fourier transform at $f=f_n$

A is the amplitude of the signal

T_o is the duration of the signal

N_D is the number of deleted samples.

A derivation of equation (6.1) can be found in Appendix B.

Results indicate that the deletion of input samples produces a reduction of the amplitude of the FFT output. This reduction is linear with the number of deleted samples because the deleted samples reduce the average value of the signal as shown in equation (6.1). The results indicate that the detection of the input frequency is not affected by deletion of samples. However, it may be expected from this test that the presence of deleted samples would produce higher error-rates in the FH-MFSK system. To overcome this effect the amplitude of the input signal must be controlled to compensate for the reduction in the amplitude of the FFT output produced by deleting samples.

The conclusion from this test is that it is not necessary to use the complete number of samples from each transmission in the analysis process. Thus the analysis can be speeded up and the design of the FFT simplified.

6.3.3.2 Delayed Samples

The system shown in Fig. 6.4 was used to examine the performance of the FFT with delayed samples. This utilises the tone generator software, the time delay software and the FFT software which were described earlier in Chapter 5. The FFT program was run for different values of delayed samples for a given frequency, then the test was repeated for different frequencies. The ratio of the absolute value of the FFT output to its input without delayed samples was plotted against the number of delayed samples. Figures 6.6a, 6.6b, 6.6c and 6.6d show the effect produced by delayed samples.

Experimental results are compared with theoretical results in figs. 6.6a, 6.6b, 6.6c and 6.6d. The difference between them are small and are due to the imperfections of the FH-MFSK generator and the FFT calculation. The theoretical results for this test were obtained using the Fourier transform of a sinusoidal signal in the presence of the delayed effect, at $f=f_n$, as given by:

$$X(f_n) = \frac{AT_o}{2} \left[1 - \frac{N_A}{64}\right] e^{-j\omega_n \delta} \tag{6.2}$$

and

$$\delta = \frac{N_A}{T_o} \tag{6.3}$$

where N_A is the number of delayed samples.
A derivation of equation (6.2) can be found in Appendix C.

Theoretical and experimental results show two important characteristics of the delayed samples tests. These are:-

(1) The FFT output contains real and imaginary components which are functions of the number of delayed samples and for this reason the absolute value is calculated to simplify the implementation of the FH-MFSK system in the presence of delayed samples. This simplification is used to overcome the problem of time delay.

(2) Results of the delayed samples test are similar to those obtained from the deleted samples test.

Delayed samples may produce higher error-rates in an FH-MFSK system. In order to overcome the effect of time delay, the absolute value of the FFT output may be used as mentioned above and the amplitude of the input signal can be controlled.

These results may be used to find numbers of deleted and delayed samples that the FFT can tolerate.

6.4 ERROR-RATE MEASUREMENTS

After the tone generator and the FFT had been tested and their characteristics with different conditions had been found, they were incorporated with the detectors, the majority decision decoder, the error-rate model and the time delay model described in Chapter 5, to find the error-rate of FH-MFSK system.

In this test the message was transmitted in one direction once to simplify the implementation.

Error-rate measurements for the FH-MFSK system were carried out in the presence of A.W.G.N. by using the system of Fig. 6.7. The r.m.s. values of FH-MFSK signals are stored in RAM as mentioned in Chapter 5. Different signal-to-noise ratios are set by dividing the added noise from the gaussian noise generator. The probability of error was determined for different signal-to-noise ratios by comparing the recovered message with the transmitted message and counting the errors due to noise. White gaussian noise added to the FH-MFSK system produced two different types of errors, deletion and false, as explained in Chapter 3.

Extensive runs of the FH-MFSK system error-rate model were performed to compute the final estimate of the probability of error for each input signal-to-noise ratio with different values of message repetitions (L) and threshold levels.

The performance of the FH-MFSK system was determined by counting the errors detected for each 1000 transmitted messages for each signal-to-noise ratio. However, when error counts were repeated they gave different values for the same input signal-to-noise ratio due to the nature of the A.W.G.N. These were therefore averaged in order to get the final estimation for error-rate.

All experimental results of error-rate were compared with theoretical results of error-rate calculation. Theoretical error-rates were calculated using equation (3.29). Also the 95% confidence limits for the average of the experimental results of error-rate are calculated. The confidence limits can be found by considering each test run for a given signal-to-noise ratio as an experiment of nt independent trials, each of which can be a success, with probability p_B or a failure with probability $q_B = 1-p_B$. Thus, each run is distributed according to the Binomial distribution. The lower and upper 95% confidence limits for the average of the experimental error-rate,[2,12], are given by

$$\pm 1.96 \sqrt{\frac{p_B q_B}{nt}} \tag{6.4}$$

where nt = number of trials

= number of transmitted messages

Results are plotted in two groups to simplify the study of results. The first group of results include the following:-

(1) Experimental and theoretical results of error-rate for different values of message repetitions (L) for a

given threshold level (0.3938V). These are shown in Figs. 6.8 and 6.9 respectively.

(2) Experimental and theoretical results and confidence limits of the experimental error-rates for a given value of L and threshold level, are shown in Figs. 6.10a, 6.10b and 6.10c.

(3) Experimental and theoretical results of bit error-probability plotted against L for a given threshold level and signal-to-noise ratio are shown in Figs. 6.11a, 6.11b and 6.11c.

The second group of results include the following:-

(1) Experimental and theoretical results of error-rate for different values of threshold levels and for a given message repetition (L=12), are shown in Fig. 6.12 and Fig. 6.13 respectively.

(2) Experimental and theoretical results and confidence limits of the experimental error-rates for a given value of L and threshold level, are shown in Figs 6.14a and 6.14b.

(3) Experimental and theoretical results of bit error-probability plotted against threshold level for a given message repetition and signal-to-noise ratio are shown in Figs 6.15a, 6.15b and 6.15c.

Receiver operating curves of noncoherent on-off keying (O.O.K.) in A.W.G.N. are shown in Fig. 6.16.

The results of error-rate measurements show the following characteristics of the FH-MFSK system. These are:-

(1) Signal-to-noise ratios greater than 9.32 dB produced no error in the received message for 1000 transmitted messages.

(2) Error-rate decreases smoothly as the number of transmitted message repetitions increases. This is due to the fact that the majority logic decoder decreases the error-rate when message repetition is used. This majority logic decoder improves the chance of correctly decoding the transmitted message on the basis of a majority decision of identical received words.

(3) Error-rate decreases as the threshold level increases. This is because the probability of error is dependent on the threshold level as mentioned in Chapter 3. False and deletion probability decreases smoothly as the threshold level increases, but the threshold level must be less than the amplitude of the actual signal.

For the range of errors investigated, the FH-MFSK system performance may be considered acceptable. This will be confirmed when the measured and calculated performances are compared with those for conventional systems later in this chapter.

6.4.1 Error-rate Calculations in the Presence of A.W.G.N. with Deleted or Delayed Samples

Using the same models as before for deleted or delayed samples tests, error-rate calculations could have been conducted in the presence of A.W.G.N. with deleted or delayed samples. However, these tests were not implemented for the following reasons:-

(1) Long tests would have been required, the time required for each run (transmission of 1000 messages) being 13.3 hours.

(2) Additional time would be required for the delayed samples test due to the need to calculate the absolute value of the FFT output.

(3) Time restriction in the availability of the microprocessor facilities.

Error-rates for the FH-MFSK system with deleted or delayed samples may be found with less effort using the previous results of error-rate and then modifying by the results of the FFT. This is due to the fact that the effect of deleted and delayed samples on the FFT output is equivalent to the reduction of the input signal amplitude as mentioned in section 6.3.3. Thus from the FFT tests the amount of the reduction in the signal level in the presence of deleted or delayed samples can be calculated. The expected signal level can be found by subtracting the amount of reduction in the level from the actual level of the signal. Thus, expected signal-to-noise ratio can be found by dividing the expected signal level by the noise level. By repeating the above procedure for all values of signal-to-noise ratios, their expected signal-to-noise ratios can be found. Error-rates can be found from Figs 6.8 and 6.9, by reading the amount of error for each value of the expected signal-to-noise ratio.

Figures 6.17 and 6.18 show the error-rate results in the presence of A.W.G.N. with deleted or delayed samples. Results indicate that the FH-MFSK system resists the deleted and delayed samples with error proportional with number of deleted or delayed samples. For signal-to-noise ratio above 9.3 dB with 8 samples delayed or deleted, the probability of bit error is found to be 0.5×10^{-3}. With 16 samples delayed or deleted, the probability of bit error is found to be 2.5×10^{-3}. However, the effect of deleted or delayed samples may be compensated by controlling the level of the input signal as mentioned earlier.

6.5 FH-MFSK SYSTEM WITH MULTI-USERS

In order to assess the application of FH-MFSK to multi-user systems, it is necessary to compute their bit error-probability under this condition. However, multi-users generate additional errors due to interference between the system users as mentioned in Chapter 3. Because of the complexity of the implementation of the experimental system with multi-users, only theoretical bit error-probability calculations have been carried out. It is even more difficult when based on the microprocessor system due to the fact that long program running times are necessary.

Mathematical analysis of the FH-MFSK system in the presence of A.W.G.N. with multi-users was derived in Chapter 3. Equation (3.29) is used to find and plot bit error-probability as functions of the number of users, signal-to-noise ratio, number of message repetitions and threshold levels.

Figure 6.19 shows bit error probability as functions of the number of users with different values of message repetitions.

Figures 6.20a, 6.20b and 6.20c show bit error-probability as functions of the number of users with different values of L, in the presence of A.W.G.N. for signal-to-noise ratio = 13.4 dB, with threshold levels of 0.2363V, 0.31504V and 0.3939V.

Figures 6.21a and 6.21b show bit error-probability as functions of the number of users with different values of L, in the presence of A.W.G.N. for signal-to-noise ratio 9.32 dB with threshold levels 0.31504V and 0.3939V.

Figures 6.22a and 6.22b are similar to Figs 6.21a and 6.21b except the signal-to-noise ratio is = 7.4 dB.

Figure 6.23 shows the maximum number of users as functions of repetition messages and signal-to-noise ratio's with bit error-probability less than 10^{-3}.

Results show the following characteristics of the FH-MFSK system with multi-users:-

(1) Errors increase sharply when number of users increases.

(2) Errors can be decreased by increasing the number of message repetitions. Message repetition greater than 12 produced bit error-probability less than 10^{-3} with 27 users. This is due to the use of the majority logic decoder as mentioned in section 6.4.

(3) A.W.G.N. degrades the number of users because it generates additional errors as mentioned in section 3.6.4.3.

(4) Errors decrease as the threshold level increases for different values of signal-to-noise ratio and message repetition, as mentioned in section 6.4.

(5) Signal-to-noise ratios greater than 13.4 dB with message repetition = 15 and threshold level = 0.393V produce

error less than 10^{-3} in the received message for 32 users. However, signal-to-noise ratios less than 13.4 dB may be used for 32 users but the value of message repetitions must be increased. The disadvantage of a large value of message repetitions is the requirement for a fast synthesizer and a fast spectral analysis processor. For this reason a message repetition = 15 is chosen.

6.6 DISCUSSION

The good performance of the FH-MFSK system does confirm that this system is of practical interest for data transmission. Error-rate measurements were carried out using a combined software and hardware experimental system, which requires less instrumentation compared with an all hardware experimental system. The obtained experimental performance is slightly inferior to the theoretical performance, since the theoretical performance is calculated as an upper bound of errors. The precision of the error measurements is limited because of the limited time available for testing. To increase the measurement precision would have required an increase in the number of transmitted messages used for each test.

The FH-MFSK transmitter was implemented based on the Intel 8085 microprocessor using the table look-up method. It can generate up to 32 frequencies and is suitable for testing the experimental system. The result which is obtained from testing the FH-MFSK transmitter is quite acceptable since no error was noticed during its operation. Also it may be used as a real time modulator for low data transmission rates.

However, FH-MFSK signals are orthogonal signals and bandlimited, thus they satisfy conditions for ideal spectral analysis by FFT. The FFT has improved the established performance in A.W.G.N., since it is an optimal filter for detecting a sinusoidal signal in the presence of A.W.G.N. as mentioned in section 6.3.2.

The majority logic decoder has improved the performance of the system due to improving the decoding of received messages as mentioned in section 6.3.2.

The FH-MFSK system resists the deleted and delayed samples in the presence of A.W.G.N.. This is clear from the results discussed earlier in this chapter. However, to obtain this improvement in performance, the added complexity in the design has to be accepted.

The number of users of an FH-MFSK system sharing the same bandwidth are evaluated theoretically as mentioned in section 6.5. It was found that 32 users could share the bandwidth with a signal-to-noise ratio = 13.4 dB and L = 15. This result demonstrates the capability of this system for multi-users. The limitation on the number of users is fixed by the minimum cross-correlation required between the addresses of the users, as mentioned in Chapter 3.

A comparison technique was established by Hansen[40] and used by Robin[67] and Jibrail[46] to compare the performance of various data transmission systems under random noise conditions. This technique may be employed to compare the established experimental results of the FH-MFSK system with several other data transmission systems. The technique employs a parameter (ρ), which was defined as[67]

$$\rho = \frac{\text{Signal energy per bit}}{\text{Noise power per unit bandwidth}} \qquad (6.5)$$

$$= \left(\frac{S}{N}\right) \left(\frac{Bo}{C}\right) \text{ dB} \qquad (6.6)$$

where

$\frac{S}{N}$ is the signal-to-noise ratio

Bo is the bandwidth where noise is measured

C is the information rate

For the FH-MFSK system, the noise was measured in a bandwidth equal to that occupied by the FH-MFSK signals. Comparison is carried out for the following systems

(a) 32 tone piccolo

(b) F.S.K. ±50 Hz shift

(c) F.S.K. ±425 Hz shift

(d) F.S.K. chirp, Bo/C = 50

(e) F.S.K. by pulse compression, Bo/C = 4

(f) D.P.S.K. by pulse compression, Bo/C = 4

(g) FH-MFSK

The performance of the binary systems (d, e and f) above are given in the form of bit error-rate (35,46), whereas the performance of the systems (a - c) are given in character error-rate (67). The error-rate of the binary systems above were converted (46) to make them equivalent to the rest of the systems involved in the comparison, so that they would be compared on the basis of character error-rate (i.e. equivalent to an alphabet of 32 characters). Figure 6.24 shows the system (a-g) plotted on ρ vs. Bo/C. The better system is the one which has the lowest ρ and larger Bo/C for the same character error-rate, in order to get high jamming margin as mentioned in the section 2.3.

Unfortunately, no experimental results were found for M-ary systems which could be used for comparison with the present system in the presence of A.W.G.N.. Thus, comparison is carried out for the theoretical results of the following systems (74), using the same technique as before.

(g) FH-MFSK

(h) Coherent detection of MPSK

(i) Differentially coherent MPSK

(k) Coherent detection of MFSK

(ℓ) Noncoherently MFSK

Figure 6.25 shows the systems (g - ℓ) plotted on ρ vs. Bo/C for the same character error-rate. The better system is again the one which has the lowest ρ and larger Bo/C .

The spectrum efficiency may be used to compare the FH-MFSK and FH-DPSK spread spectrum performance as mentioned in section 3.4. The number of users that can be accommodated simultaneously in a given bandwidth for a certain bit error-probability can be evaluated by the spectrum efficiency [44,60]. Comparison is carried out theoretically for the following systems[34], since no experimental results were found for other spread spectrum systems which could be used for comparison in the presence of A.W.G.N.

(1) FH-MFSK, with $\frac{R}{W} = 0.0016$

(2) FH-DPSK (linear combining), with $\frac{R}{W} = 0.0016$

(3) FH-MFSK, with $\frac{R}{W} = 0.0104$.

Figure 6.26 shows the systems (1 - 3) plotted on spectrum efficiency vs. bit error-probability so that at the same bit error-rate, the better system is the one which has the larger spectrum efficiency.

In the light of the results obtained, it may be concluded that the FH-MFSK system is of practical interest. Moreover, the performance obtained highlights the candidacy of this system for consideration for use in data transmission in the presence of A.W.G.N..

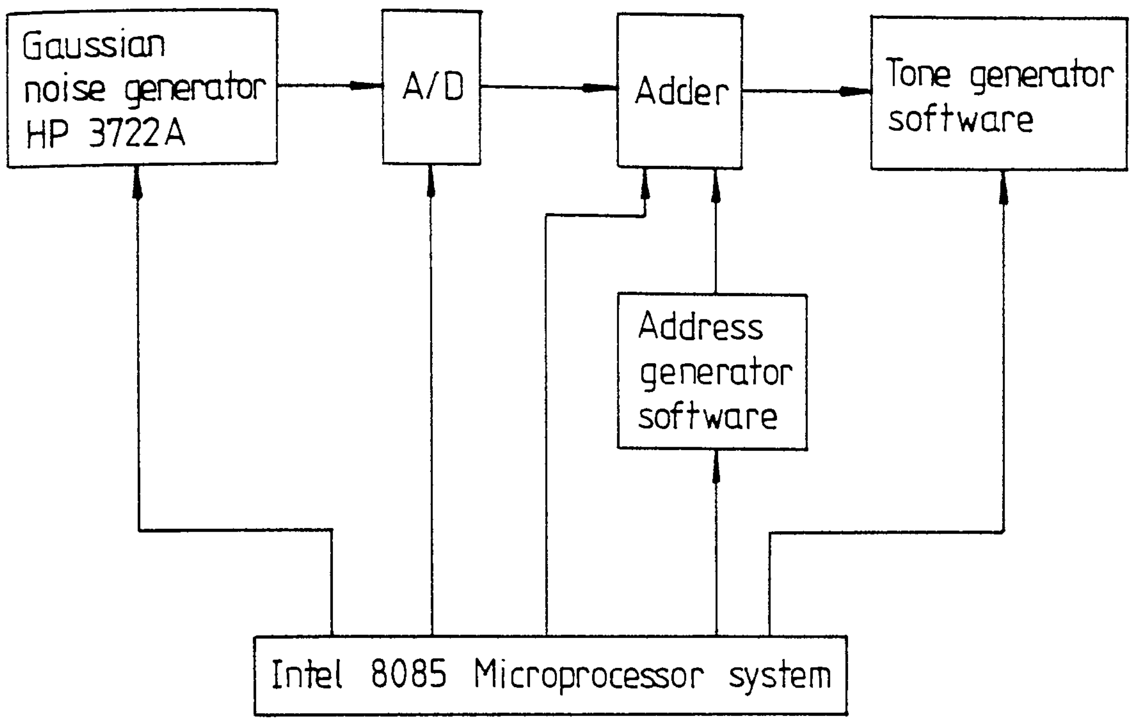


Fig 6.1 Experimental system for testing the FH-MFSK transmitter.

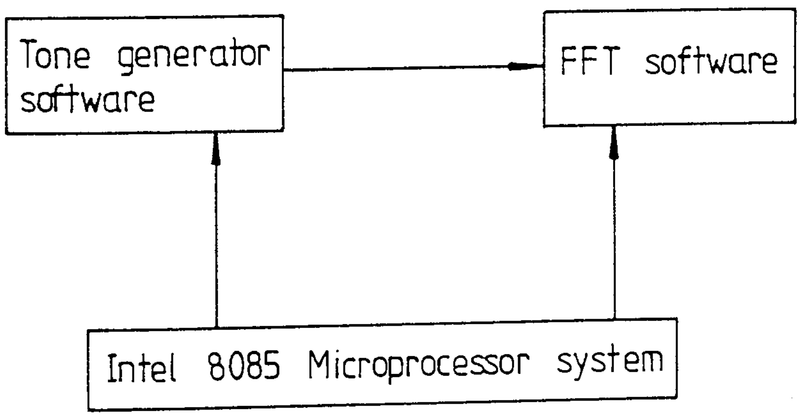


Fig 6.2 Experimental system for testing the FFT in noiseless.

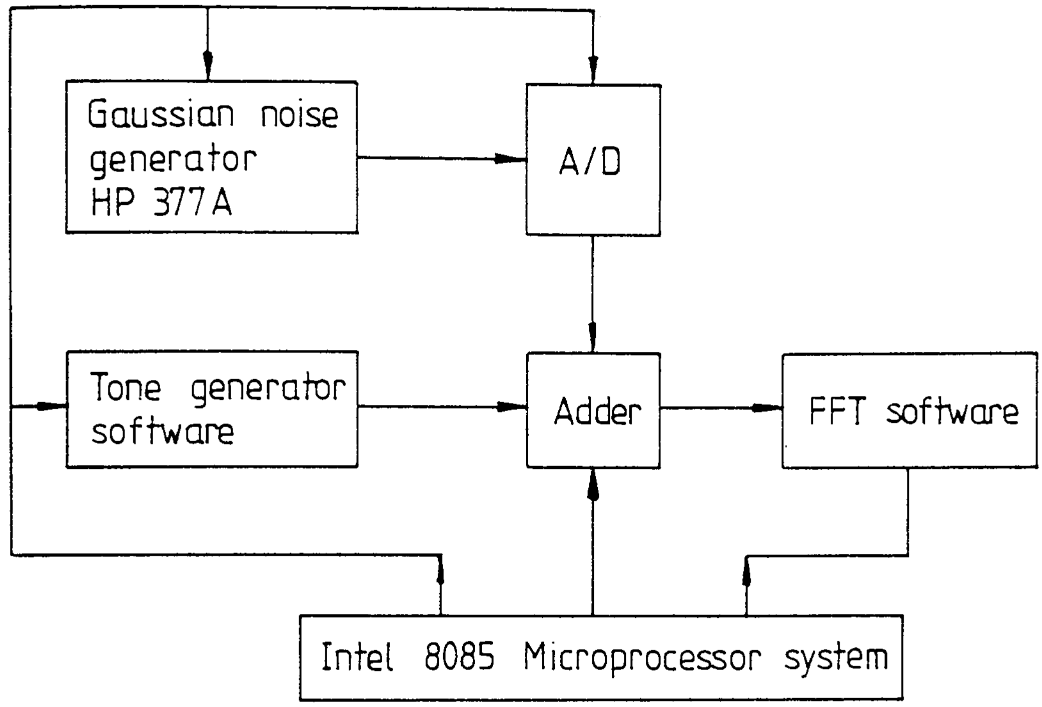


Fig 6 . 3 Experimental system for testing FFT in the presence of A.W.G.N.

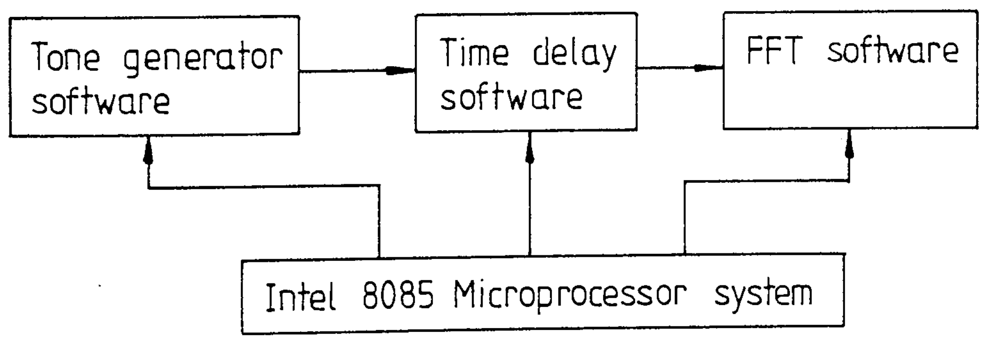


Fig 6 . 4 Experimental system for testing FFT with time delay.

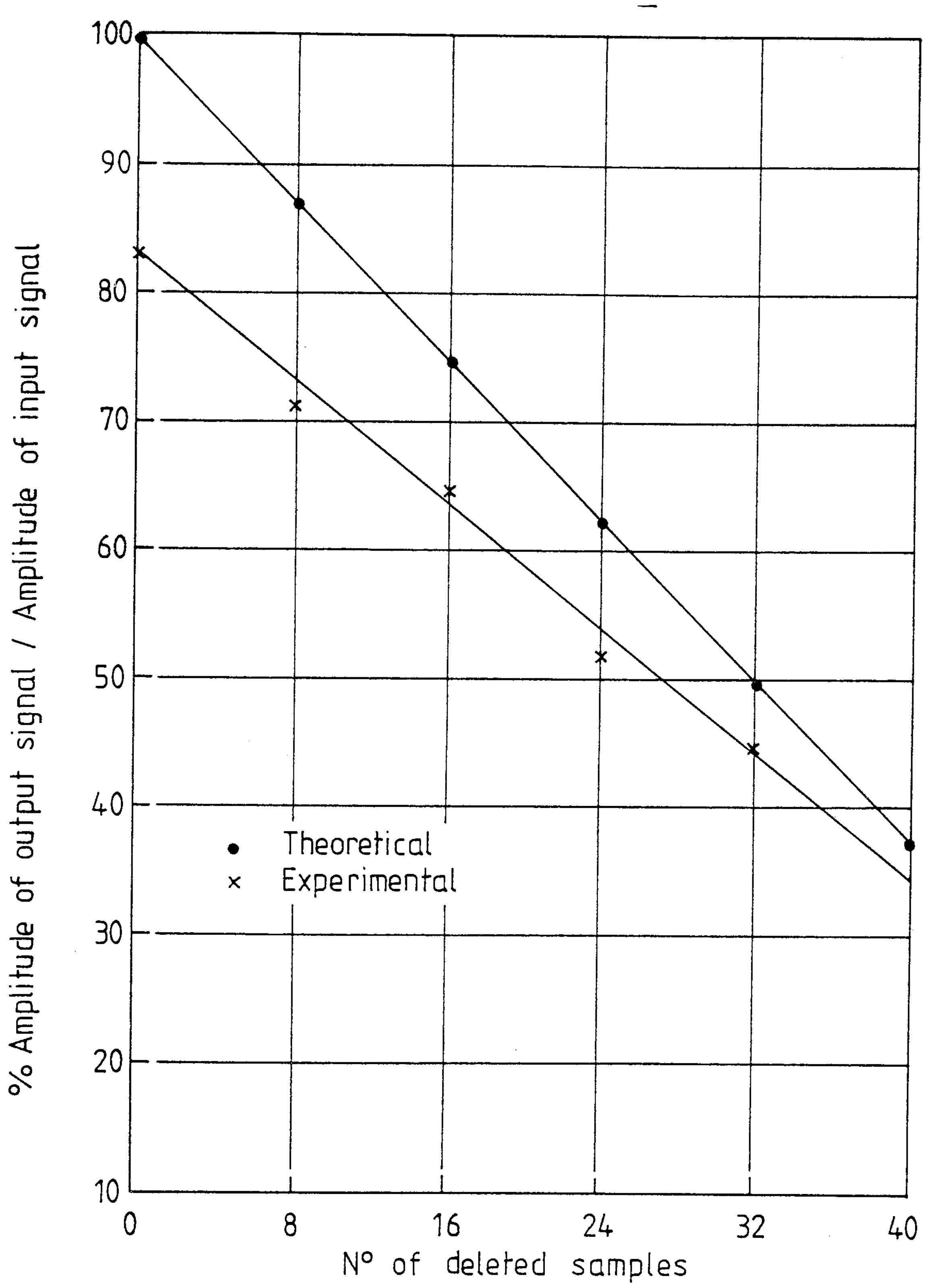


Fig 6.5a Effect of deleted samples on FH-MFSK signal after FFT for f_1

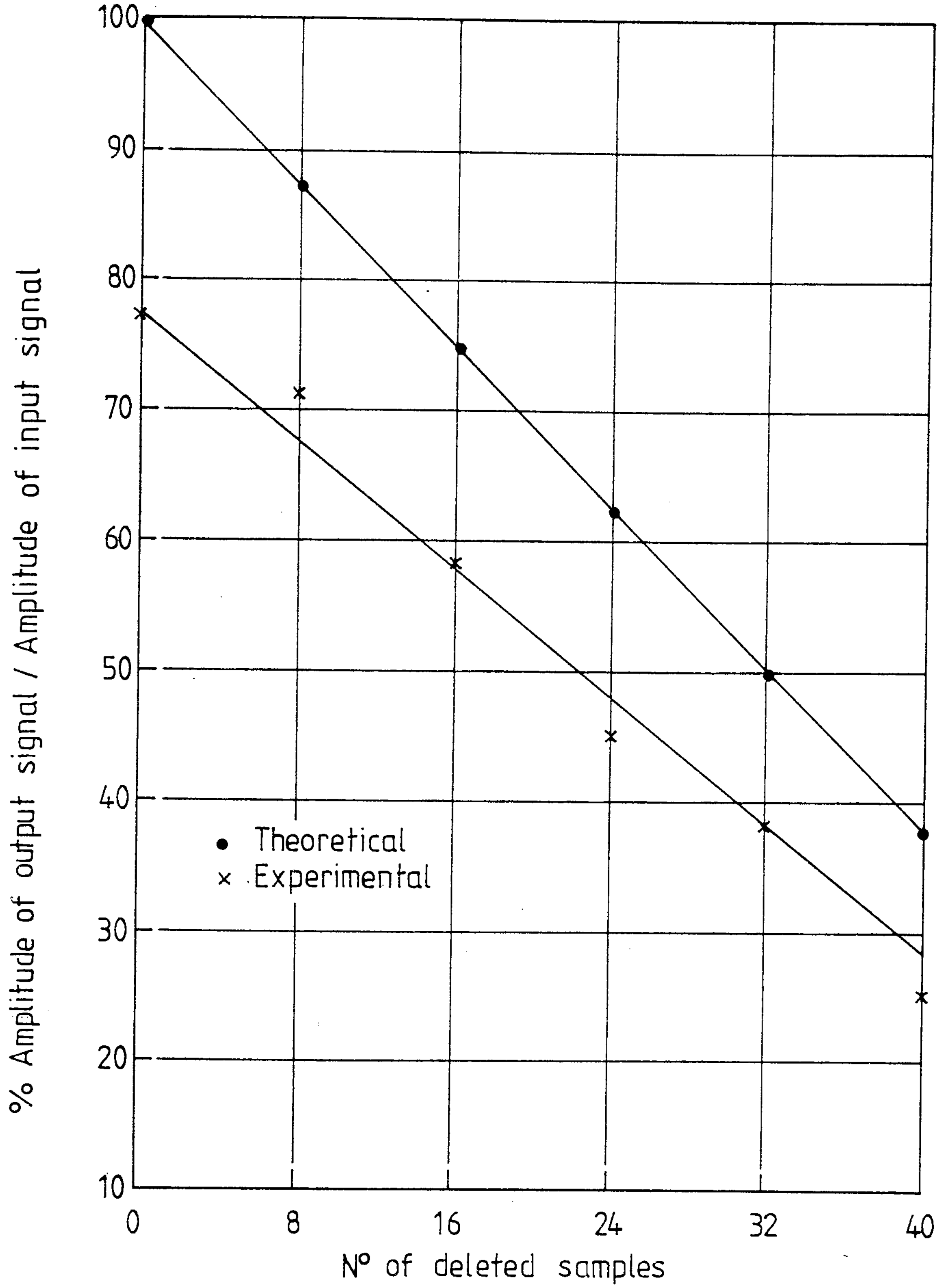


Fig 6.5b Effect of deleted samples on FH - MFSK signal after FFT for f_5

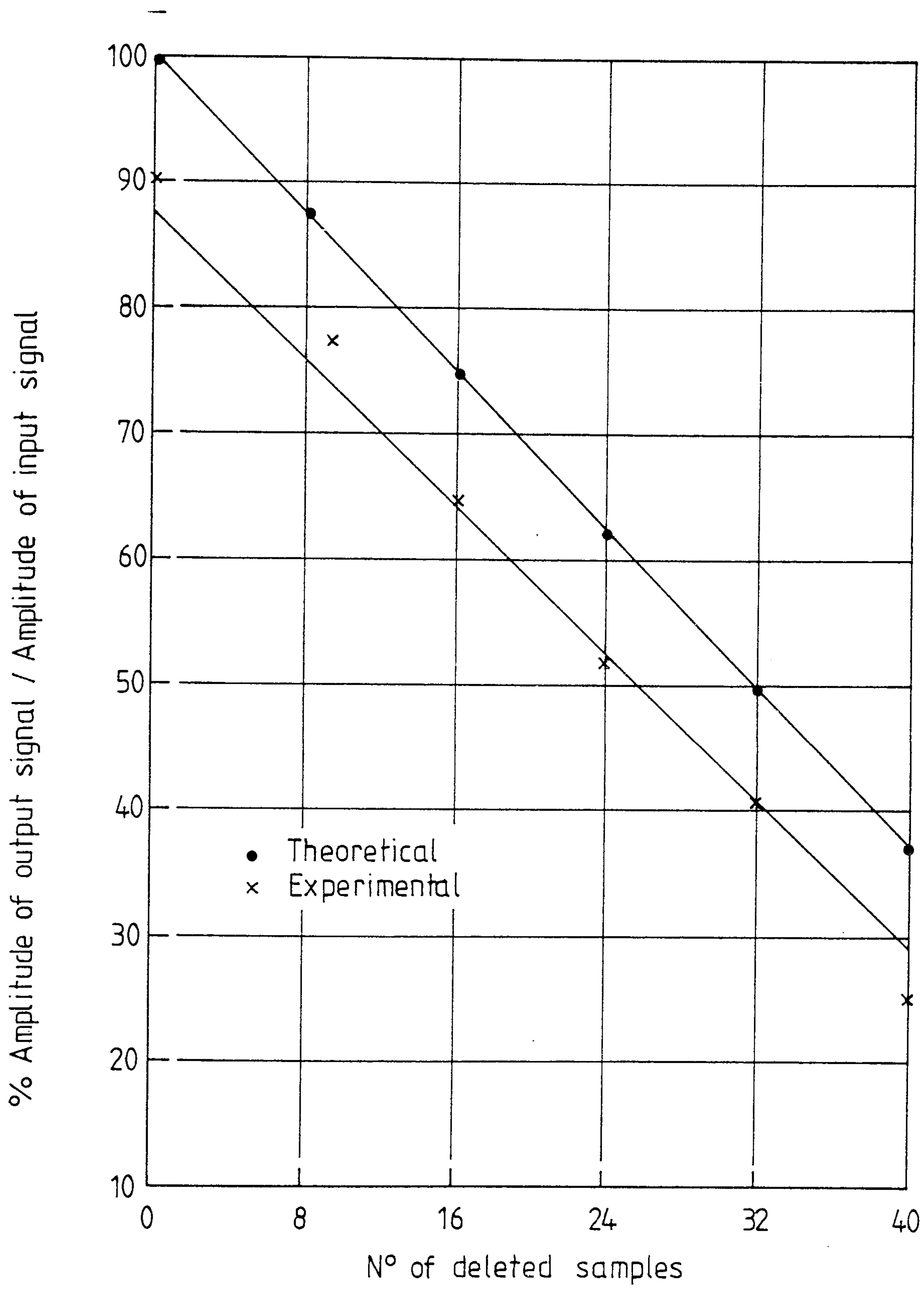


Fig 6.5c Effect of deleted samples on FH-MFSK signal after FFT for f_8

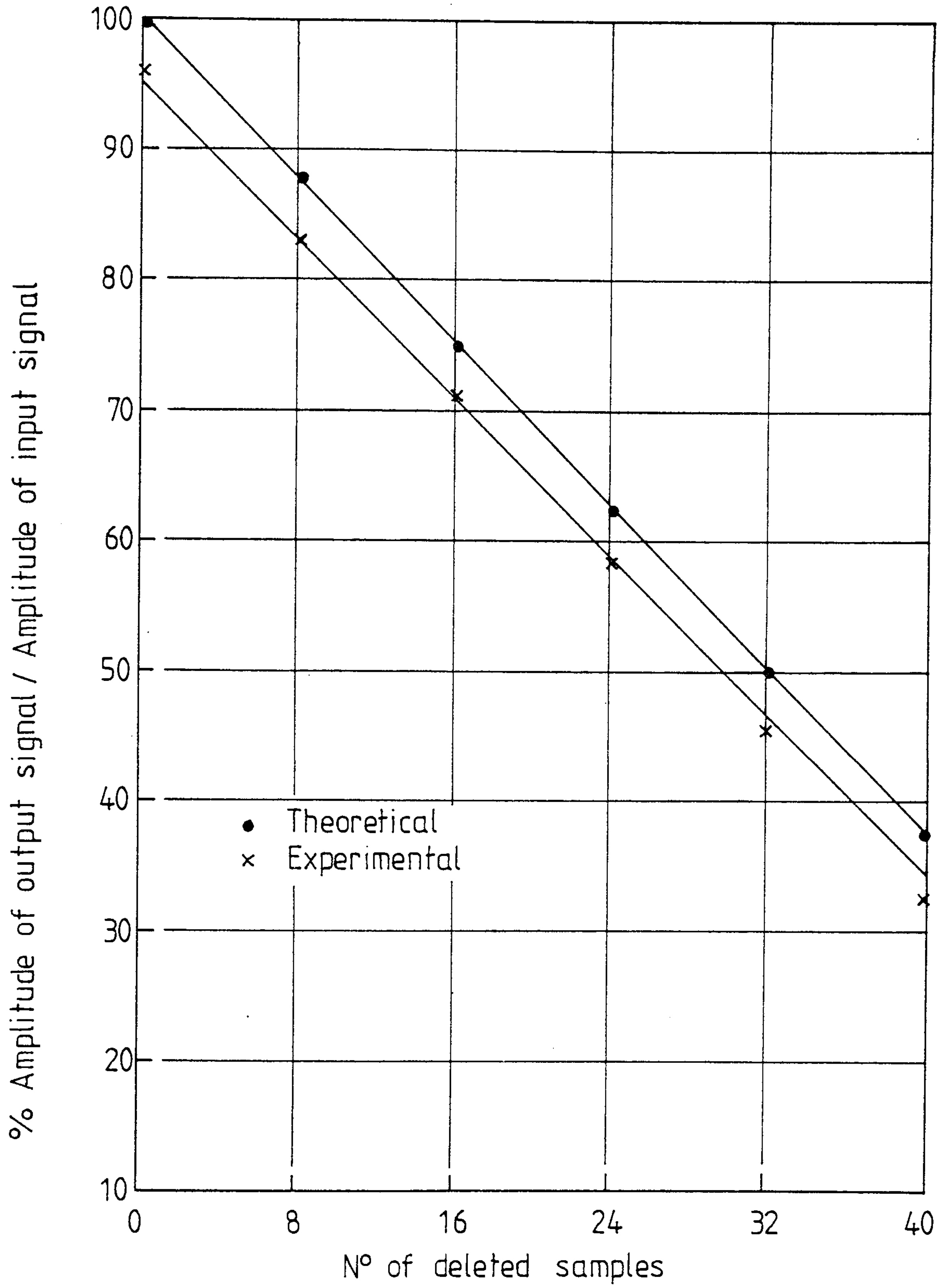


Fig 6.5d Effect of deleted samples on FH-MFSK signal after FFT for f_{16}

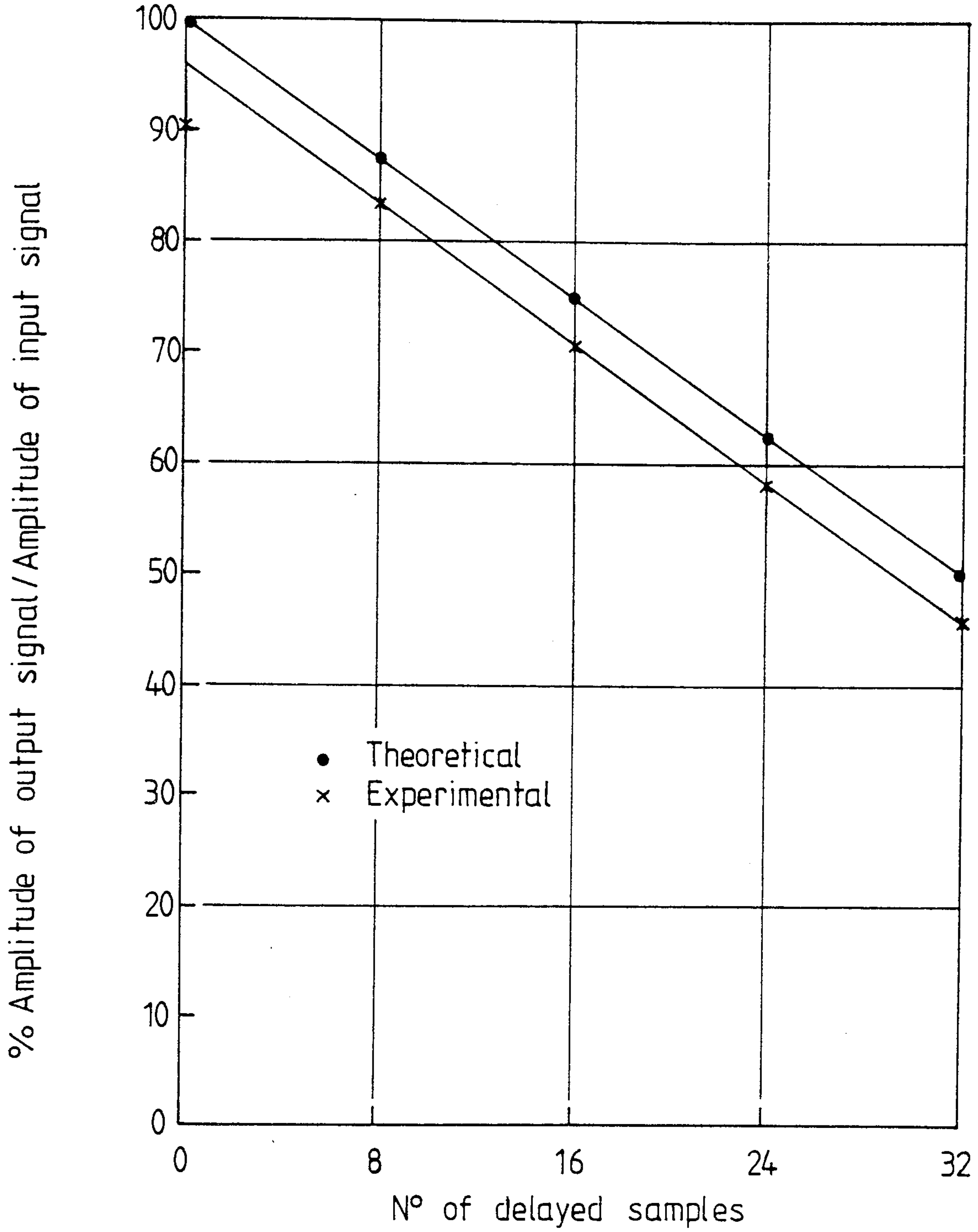


Fig 6.6a Effect of delayed samples on FH-MFSK signal after FFT for f_1

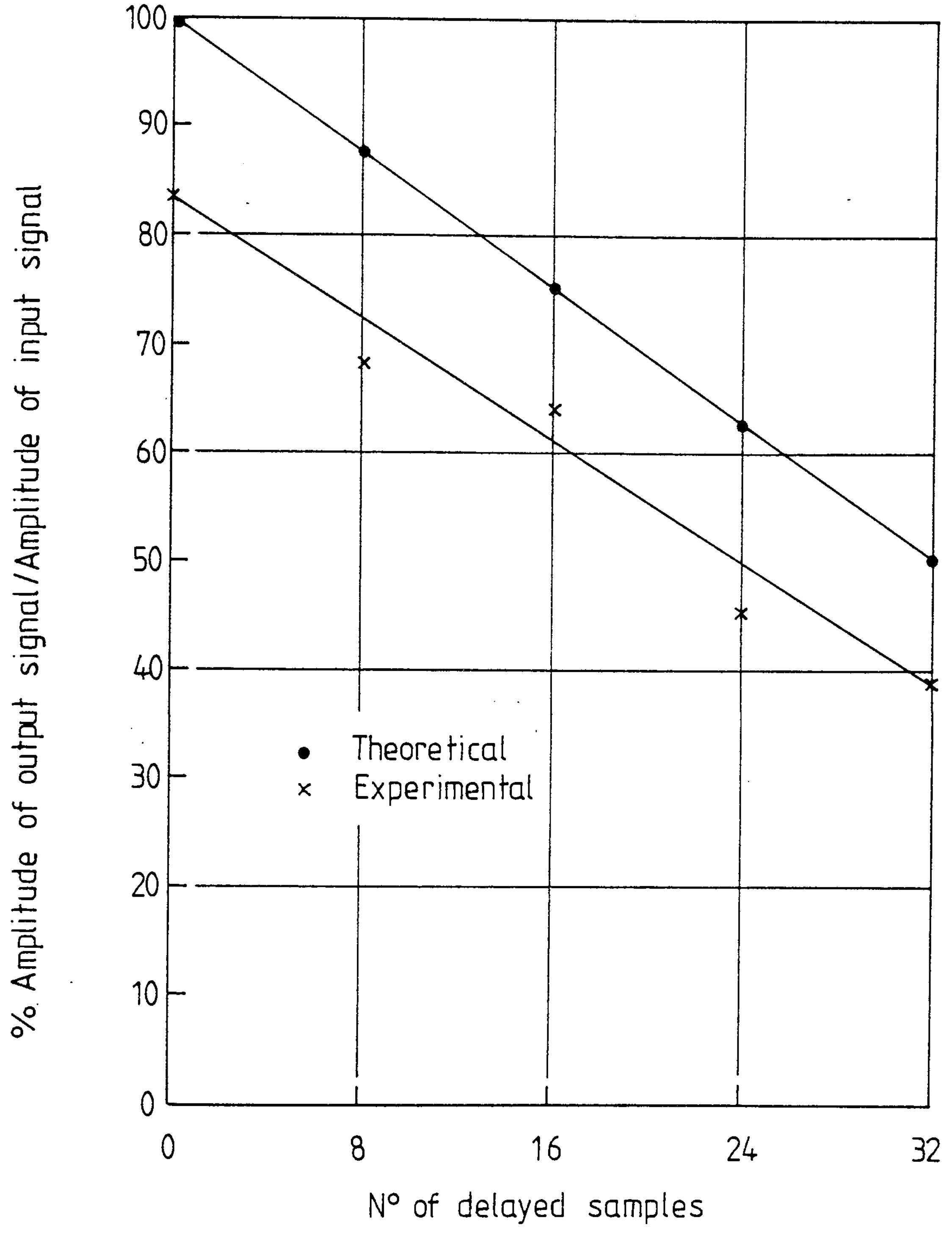


Fig 6.6b Effect of delayed samples on FH-MFSK signal after FFT for f_5

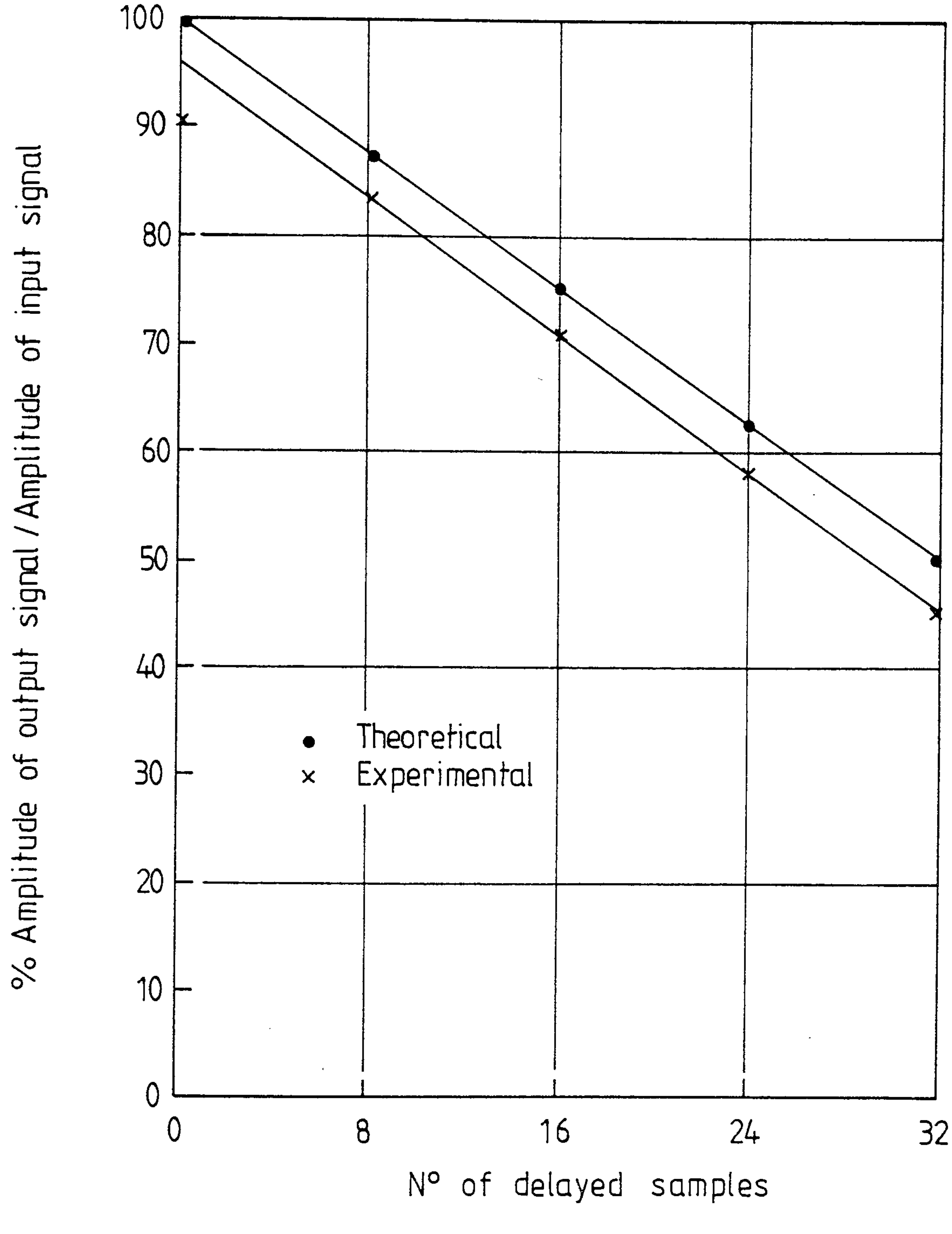


Fig 6.6c Effect of delayed samples on FH-MFSK signal after FFT for f_8

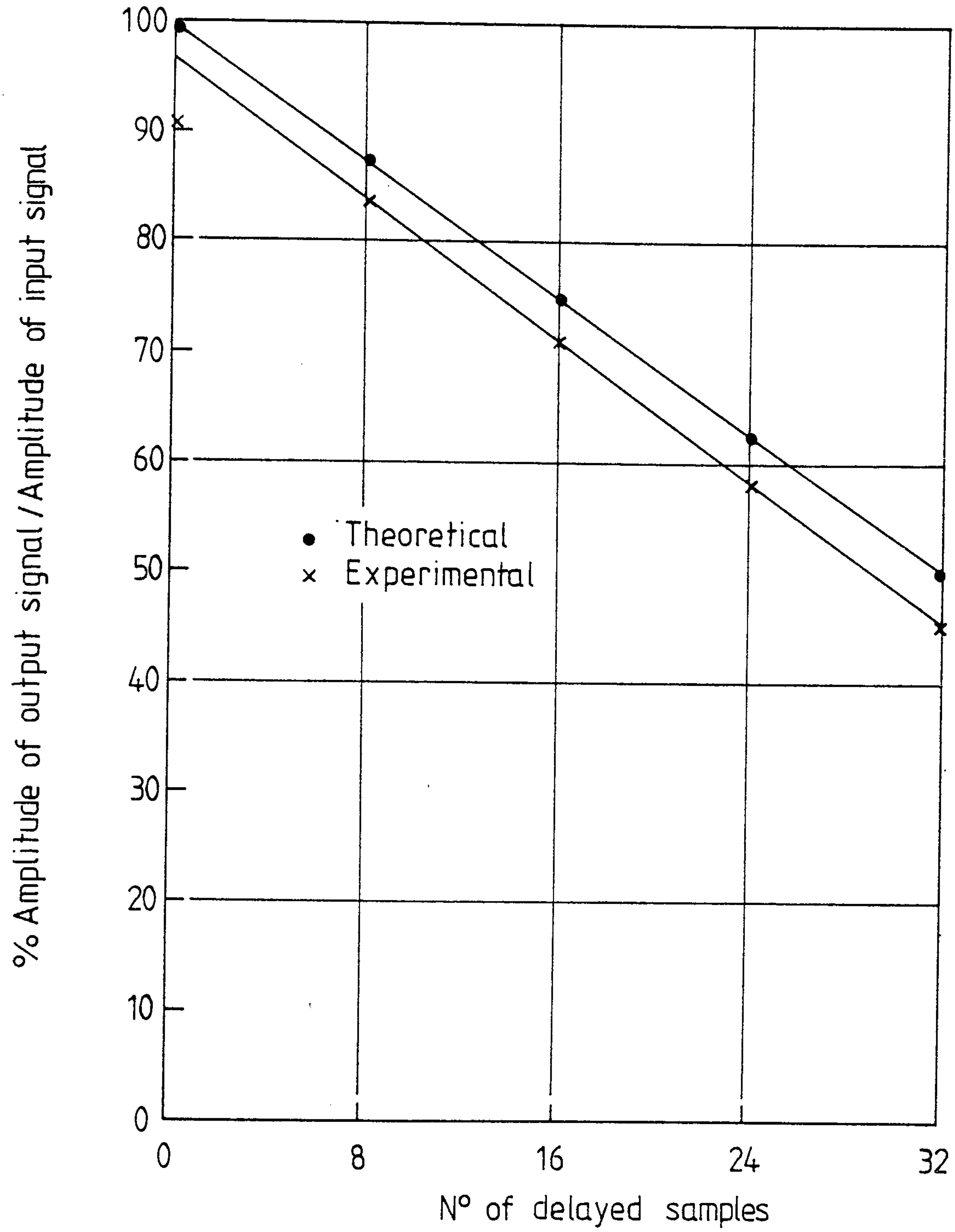


Fig 6.6d Effect of delayed samples on FH-MFSK signal after FFT for f_{16}

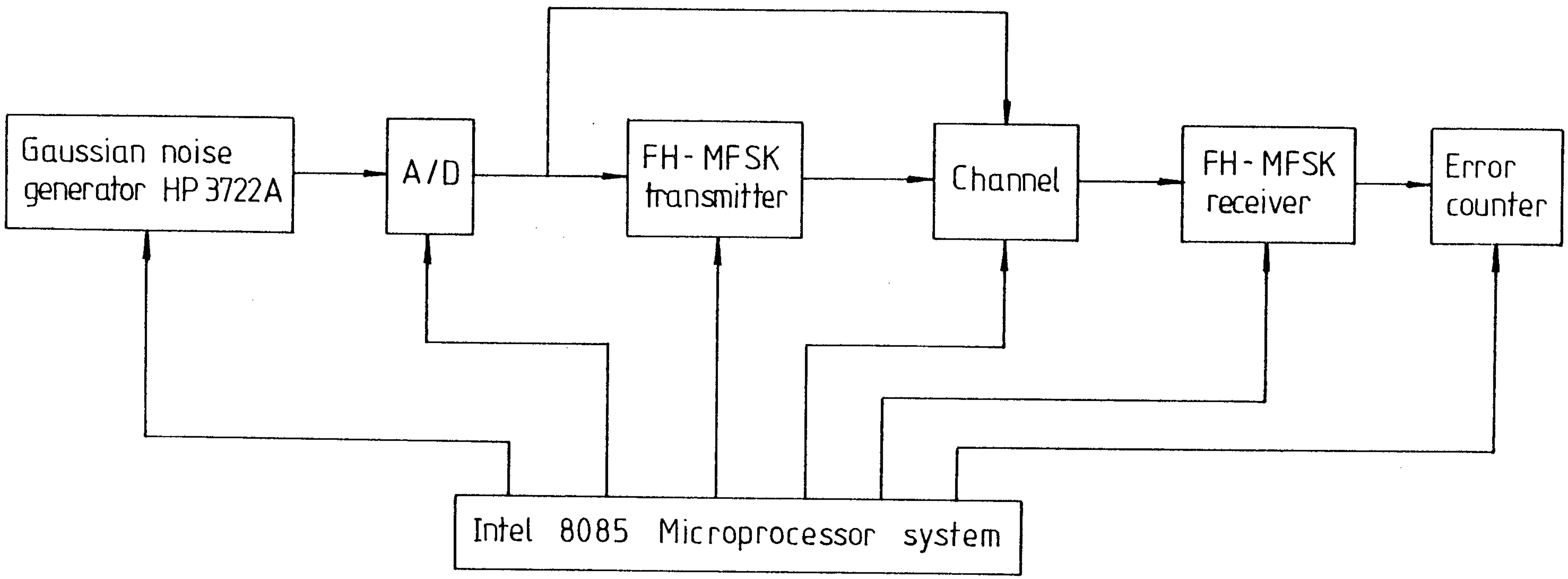


Fig 6.7 Experimental system for the error-rate model.

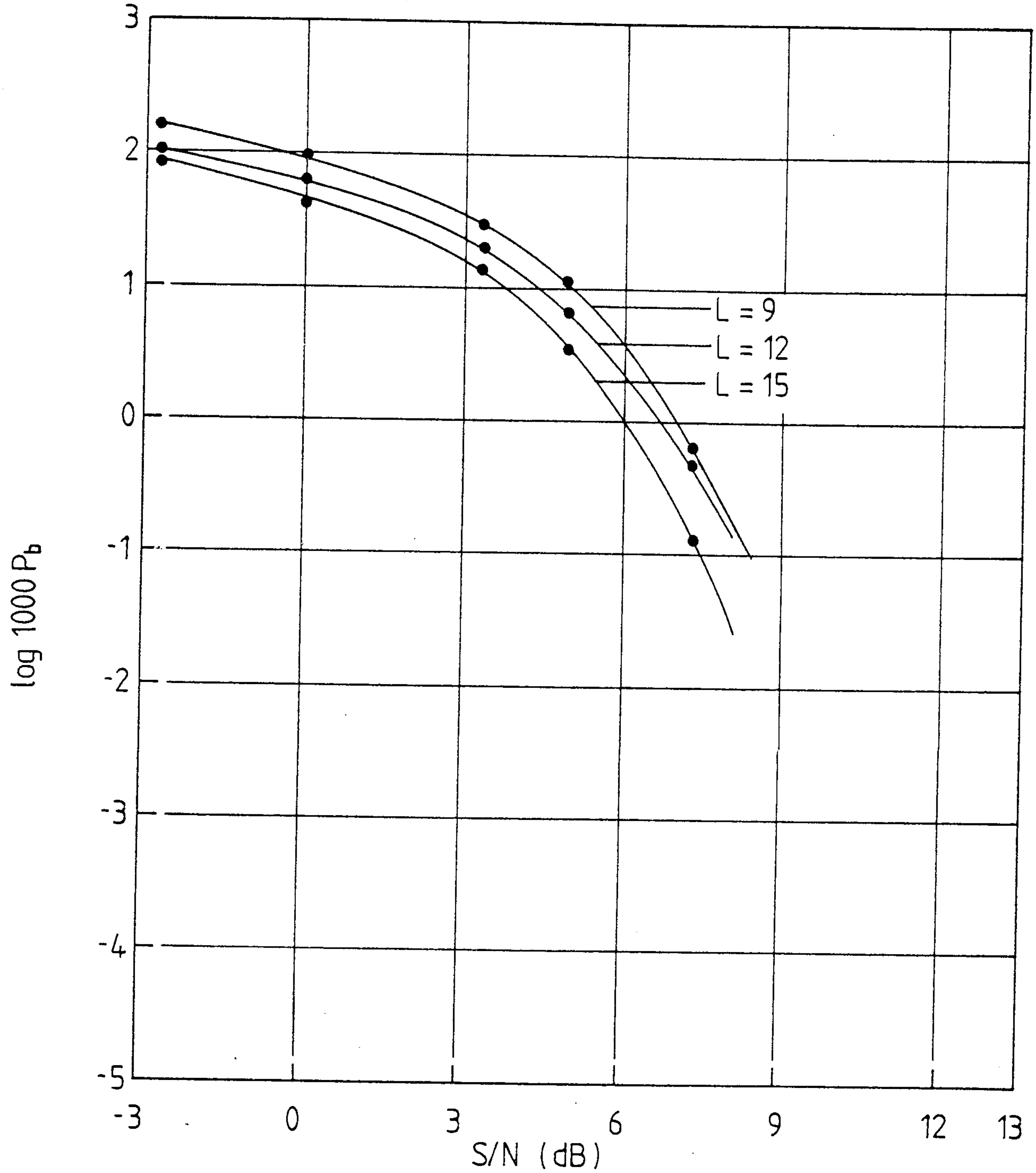


Fig 6.8 Experimental error-rate of FH-MFSK in the presence of A.W.G.N. with threshold level = 0·3938 V

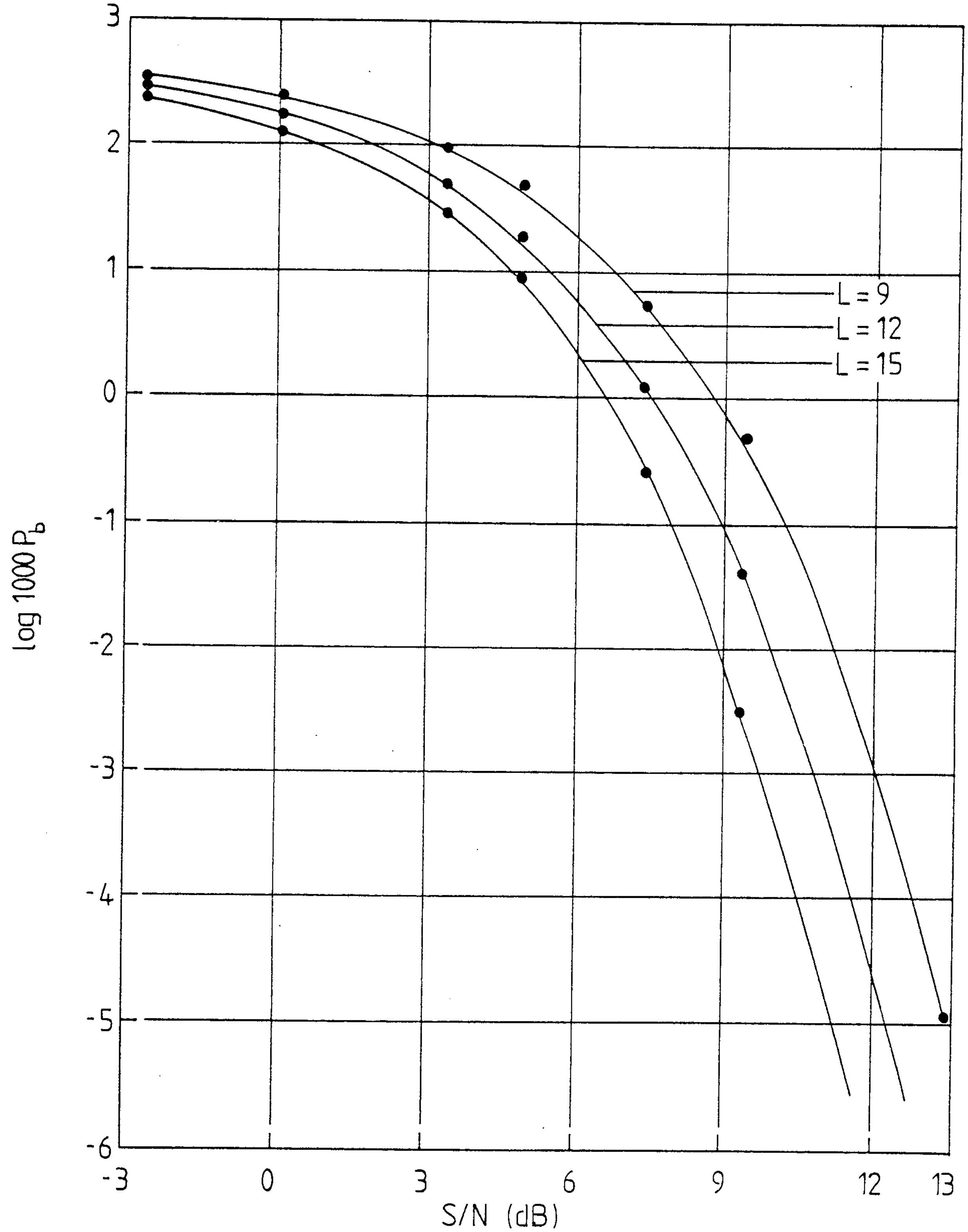


Fig 6.9 Theoretical error-rate of FH-MFSK in the presence of A.W.G.N. with threshold level = 0·3938V

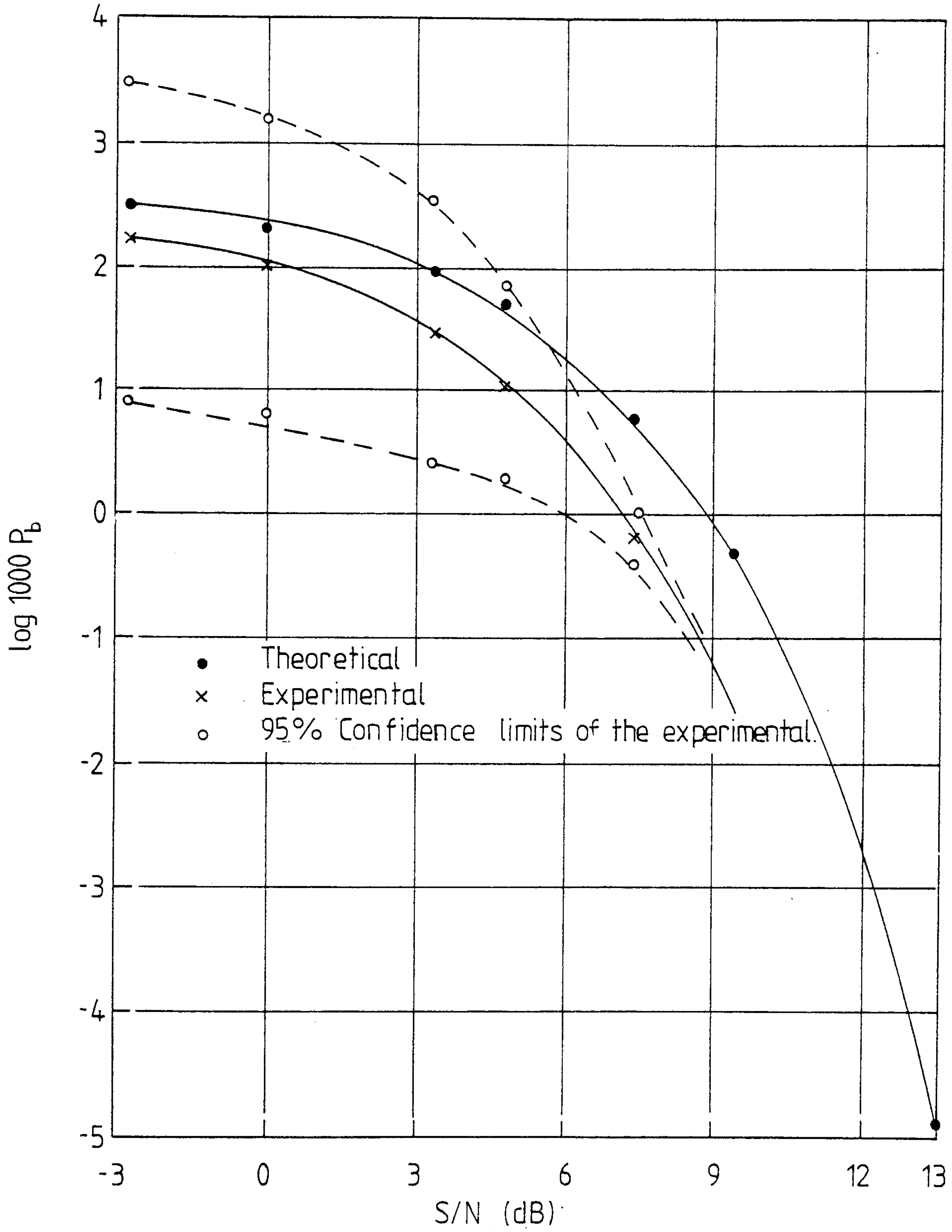


Fig 6.10a Error-rate of FH-MFSK in the presence of A.W.G.N. with threshold level = 0·3938 V , L = 9

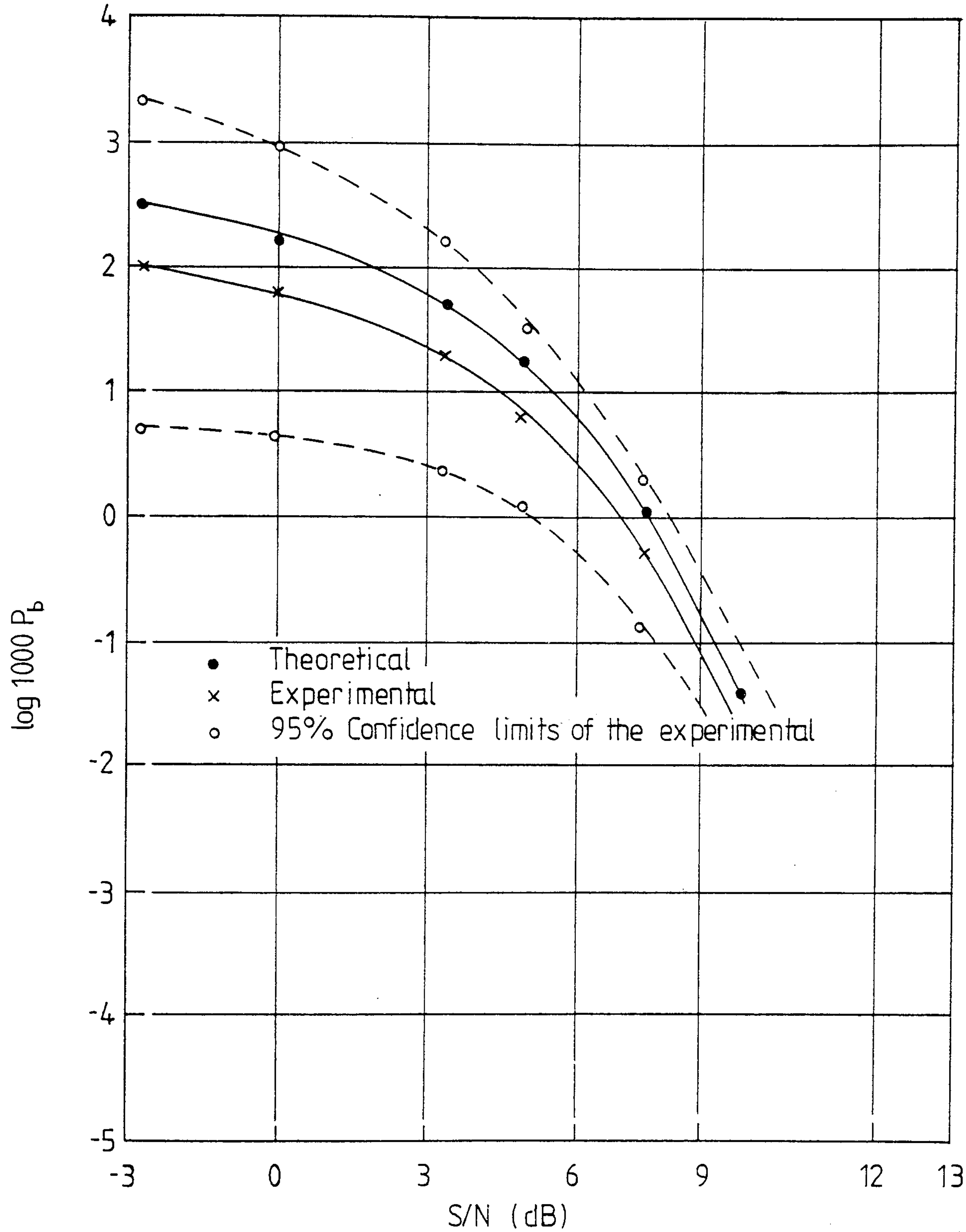


Fig 6.10b Error-rate of FH-MFSK in the presence of A.W.G.N. with threshold level = 0·3938V, L=12

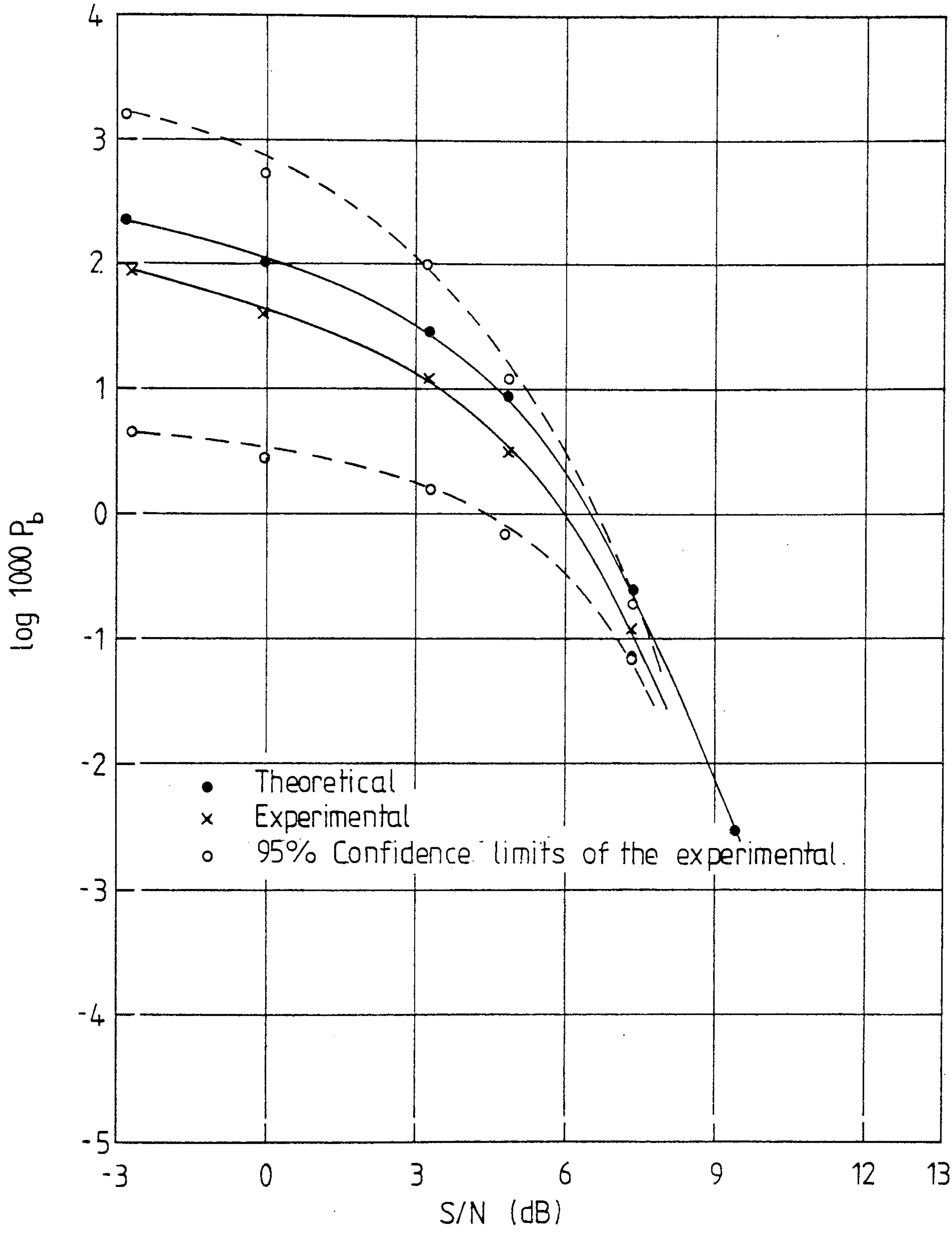


Fig 6.10c. Error-rate of FH-MFSK in the presence of A.W.G.N. with threshold level = 0·3938 V, L = 15

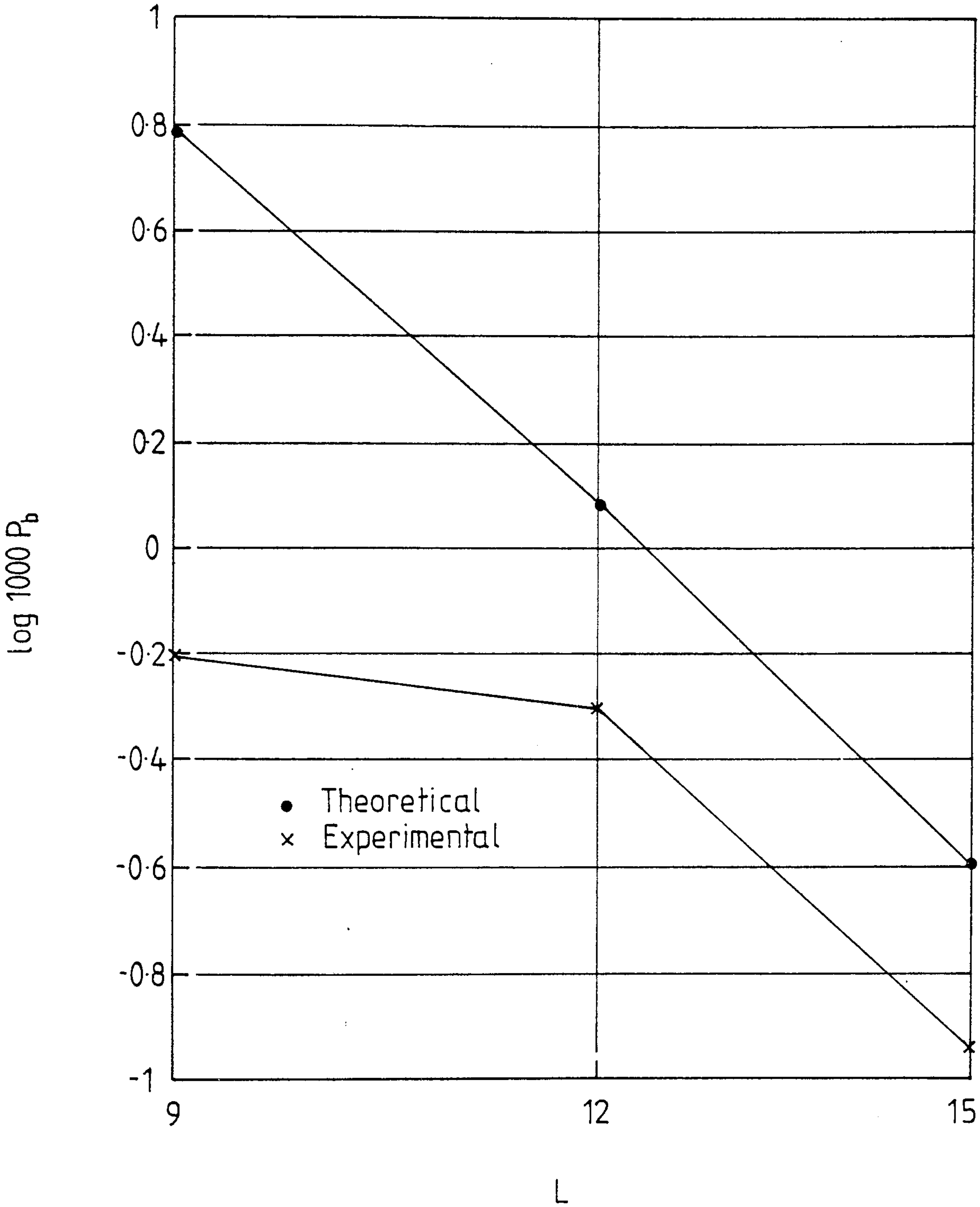


Fig 6.11a Bit error-probability as functions of the number of message repetitions for a FH-MFSK system in the presence of A.W.G.N. for S/N = 7·4 dB and threshold level = 0·3938V

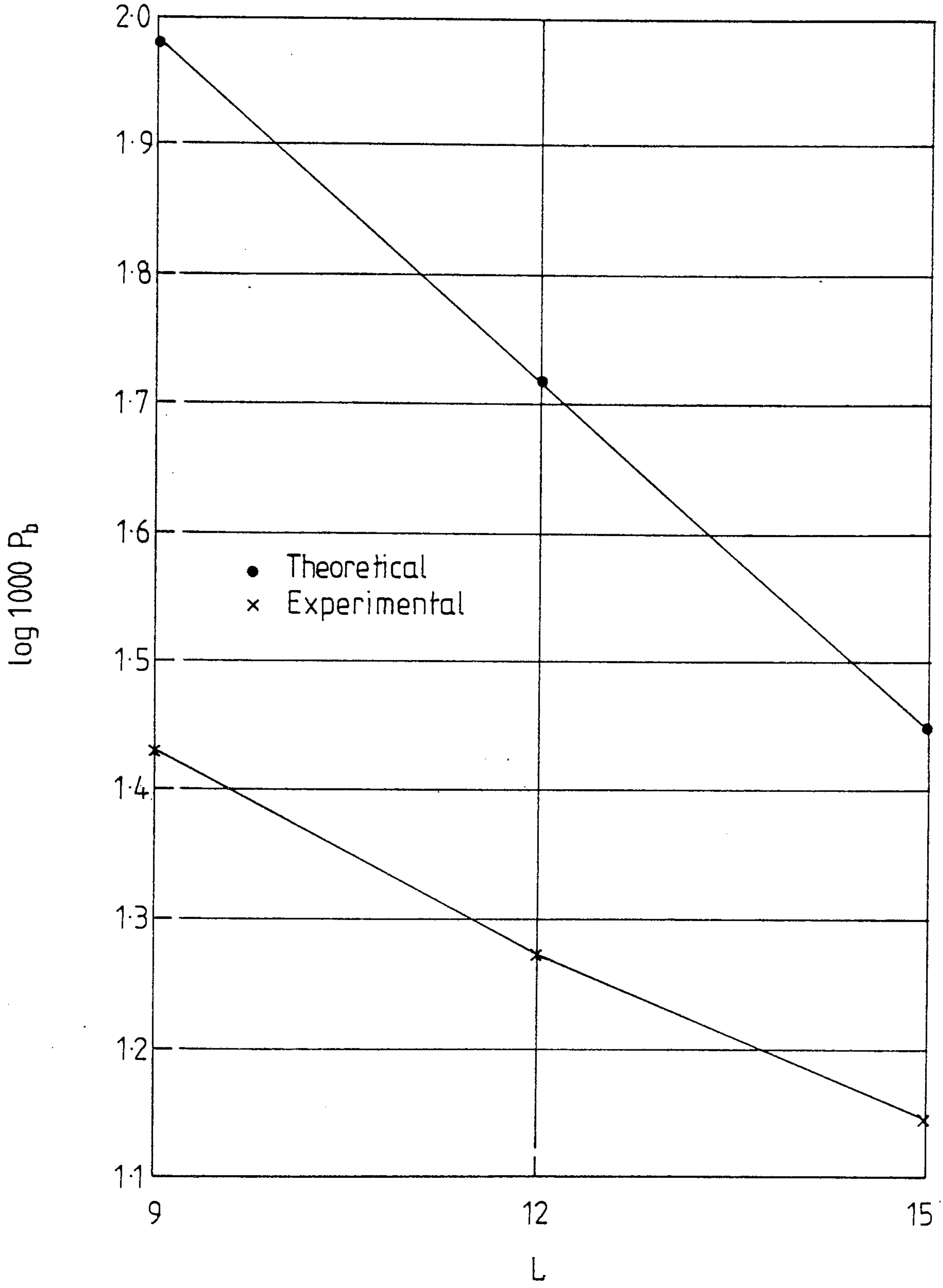


Fig 6.11b Bit error-probability as functions of the number of message repetitions for a FH-MFSK system in the presence of A.W.G.N. for S / N = 3·3 dB and threshold level = 0·3938V

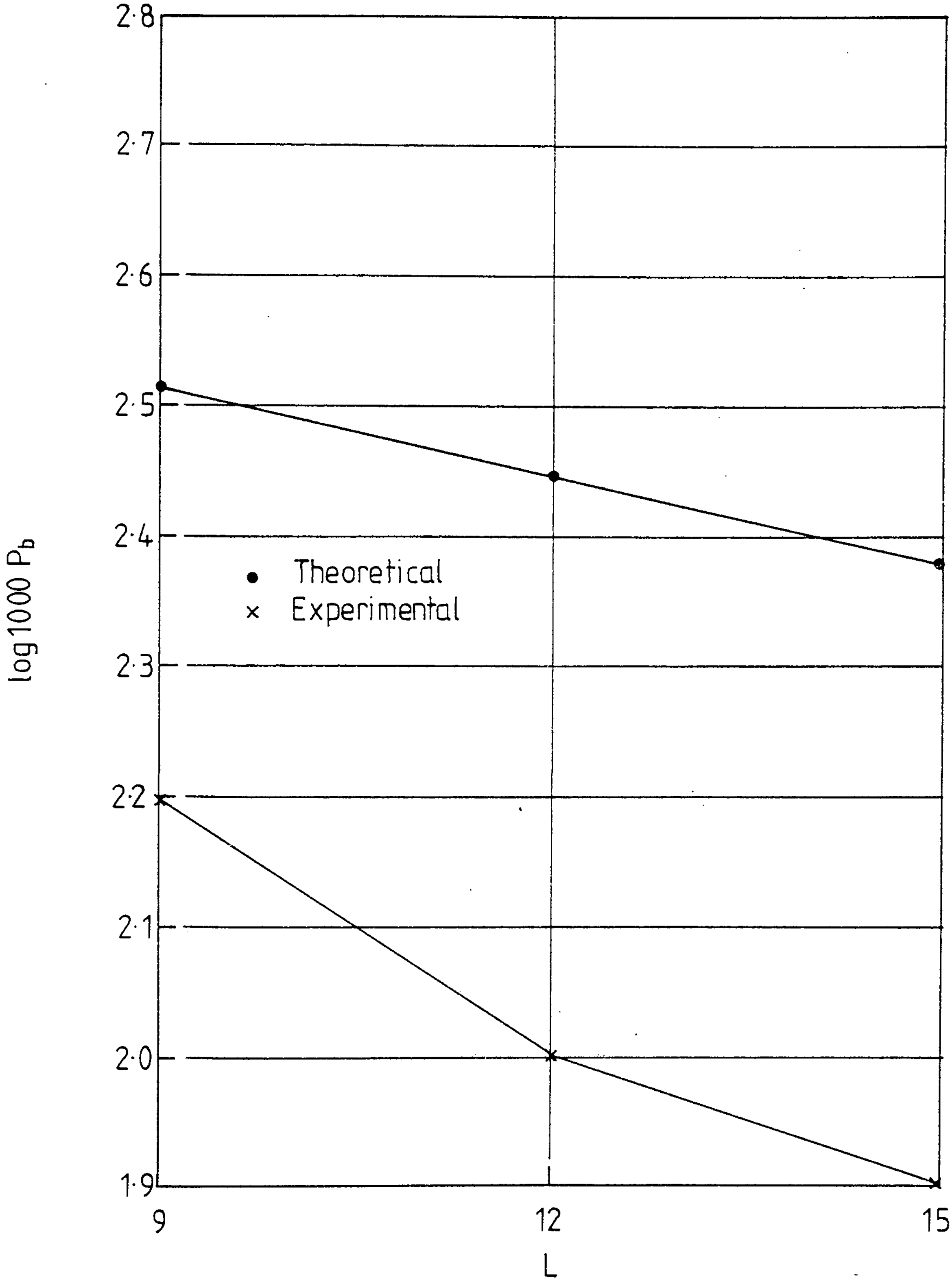


Fig 6.11c Bit error-probability as functions of the number of message repetitions for a FH-MFSK system in the presence of A.W.G.N. for S/N = - 2·72 dB and threshold level = 0·3938 V

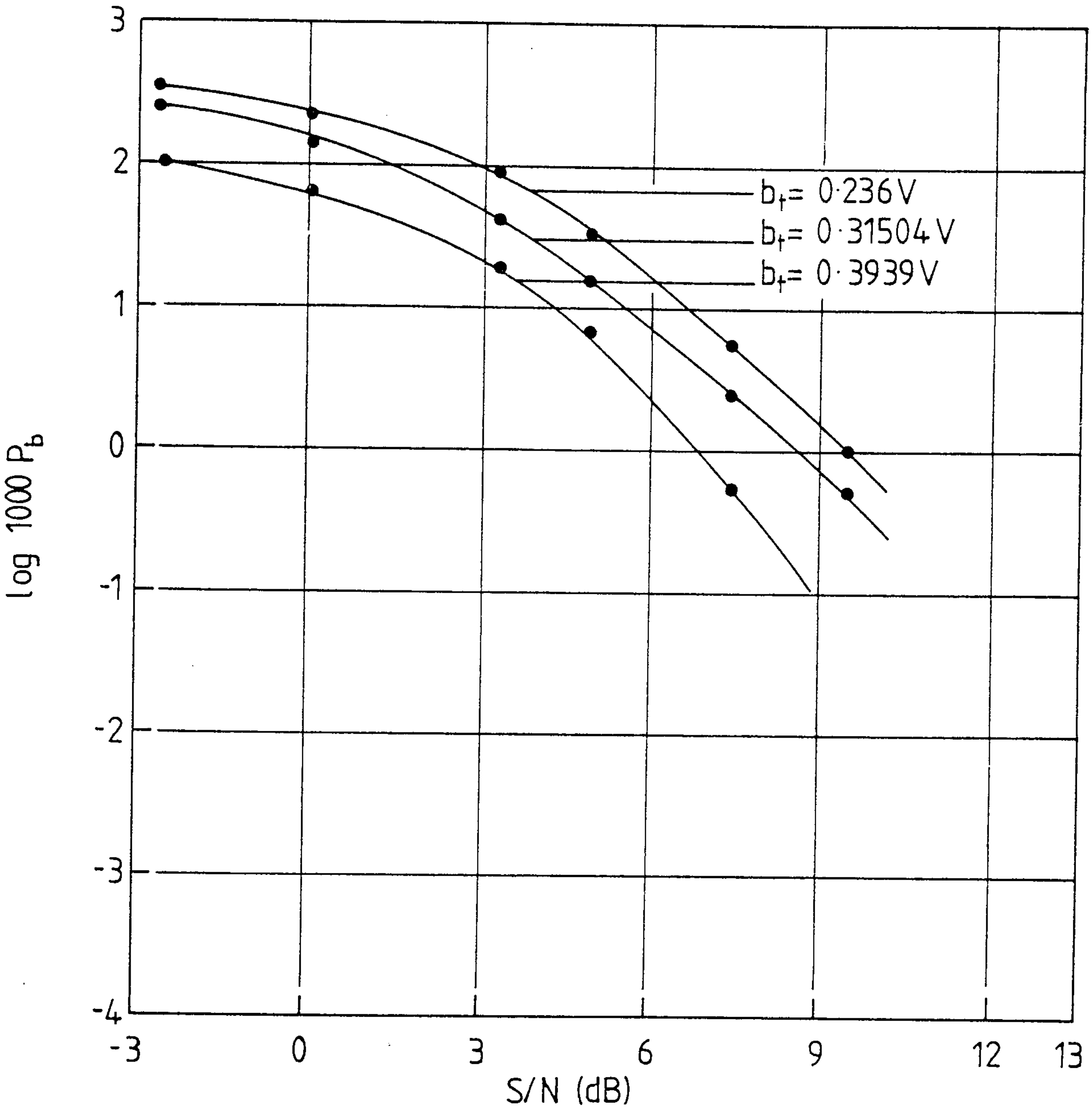


Fig 6.12 Experimental error-rate of FH-MFSK in the presence of A.W.G.N. with L=12

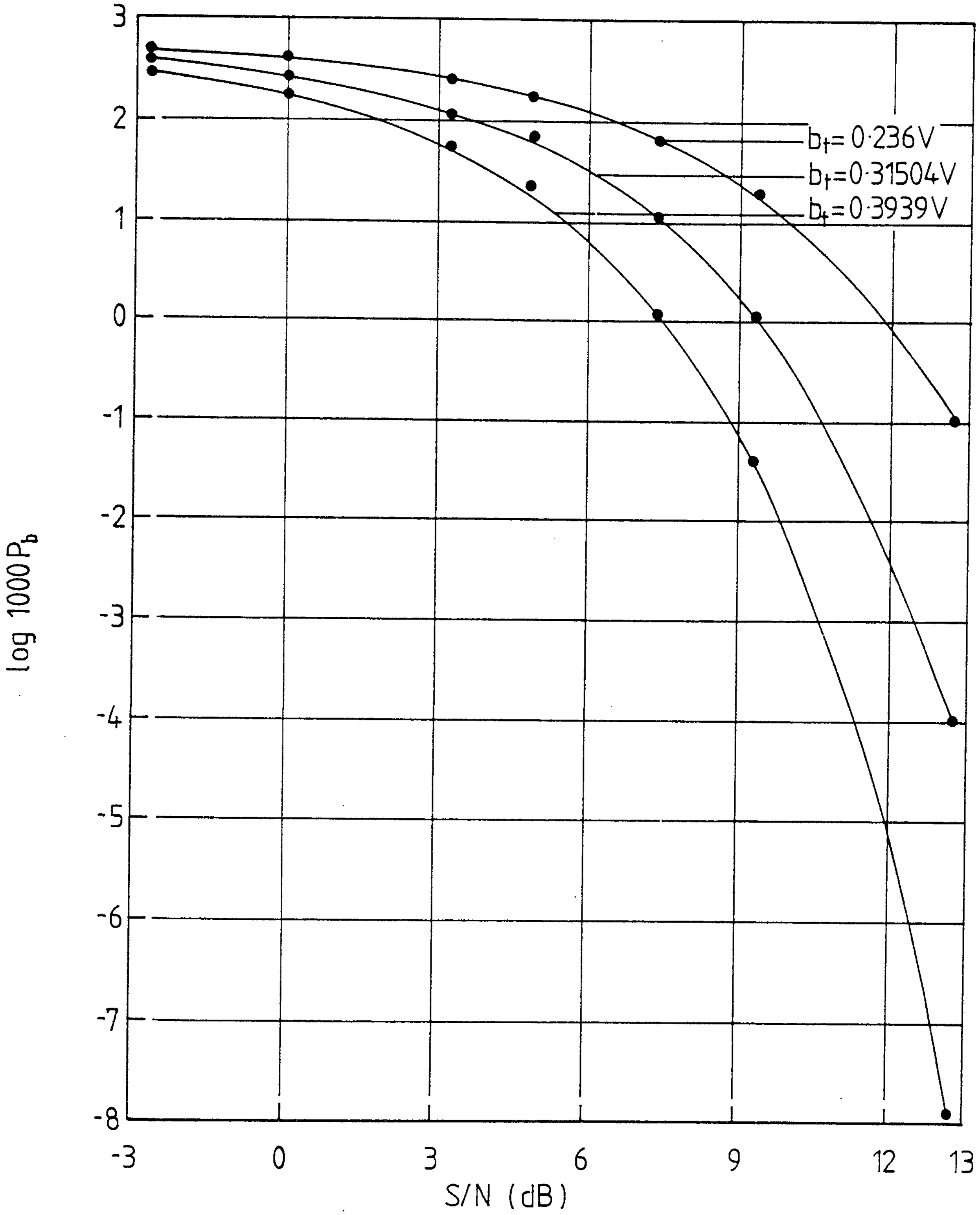


Fig 6.13 Theoretical error-rate of FH-MFSK in the presence of A.W.G.N. with L = 12

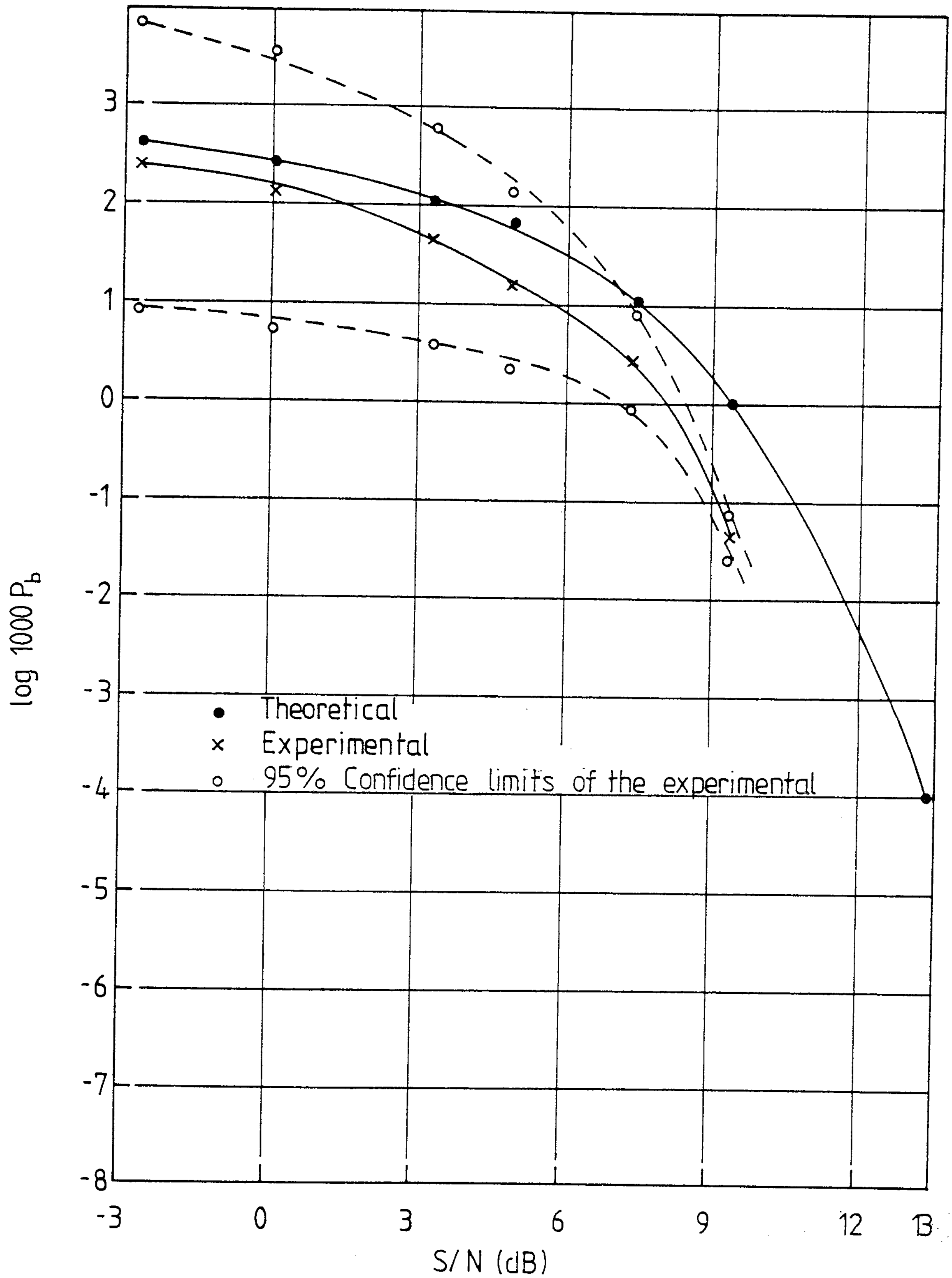


Fig 6.14a Error-rate of FH-MFSK signal in the presence of A.W.G.N. with threshold level = 0·31504V, L=12

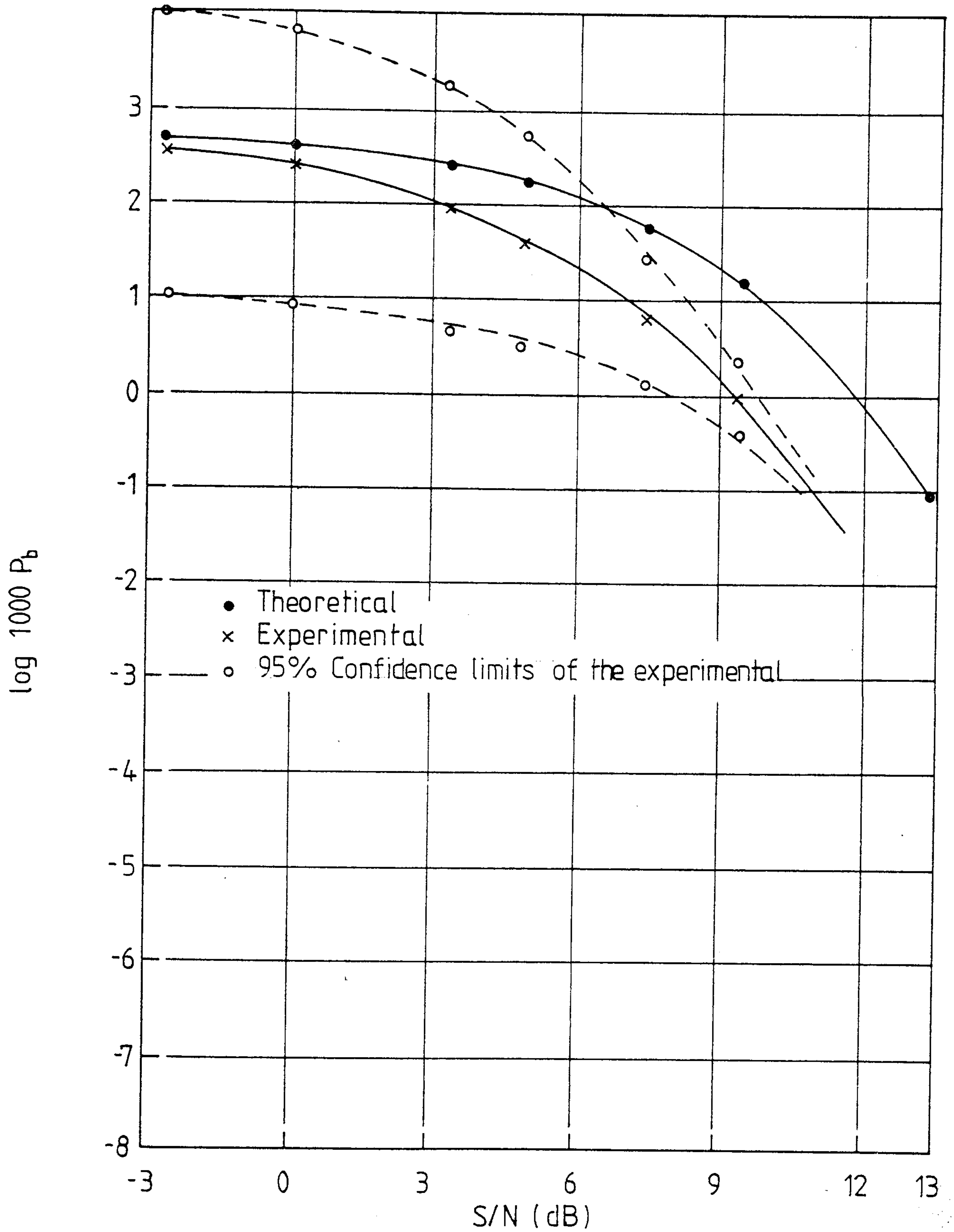


Fig 6.14b Error-rate of FH-MFSK signal in the presence of A.W.G.N. with threshold level = 0·236V, L=12

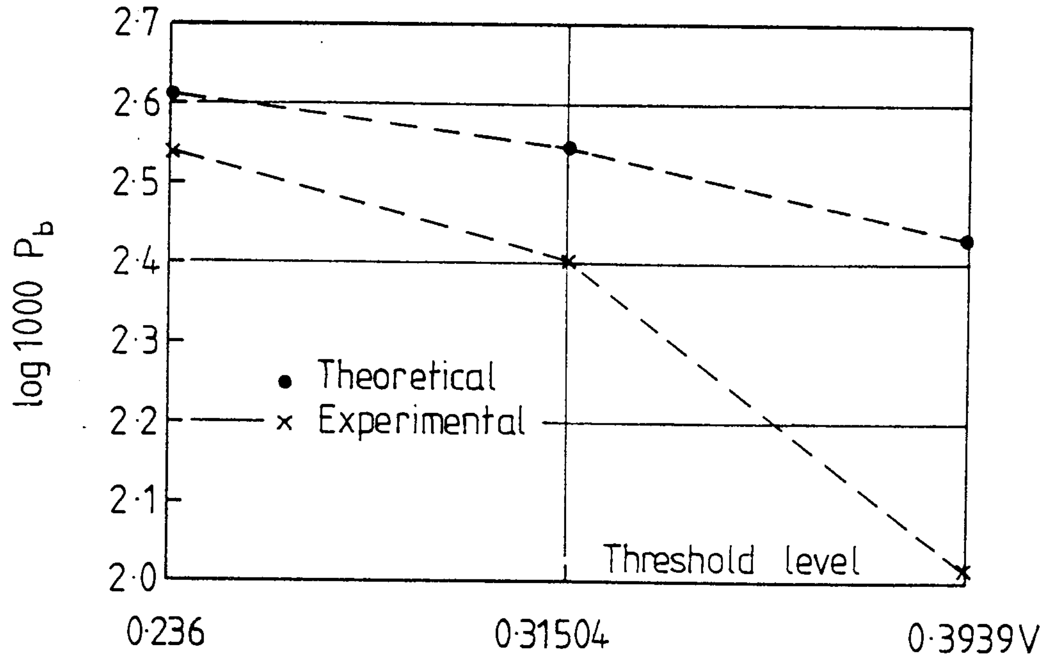


Fig 6.15a Bit error-probability as functions of the threshold levels for FH-MFSK system in the presence of A.W.G.N. for S/N = −2·72 dB , L = 12

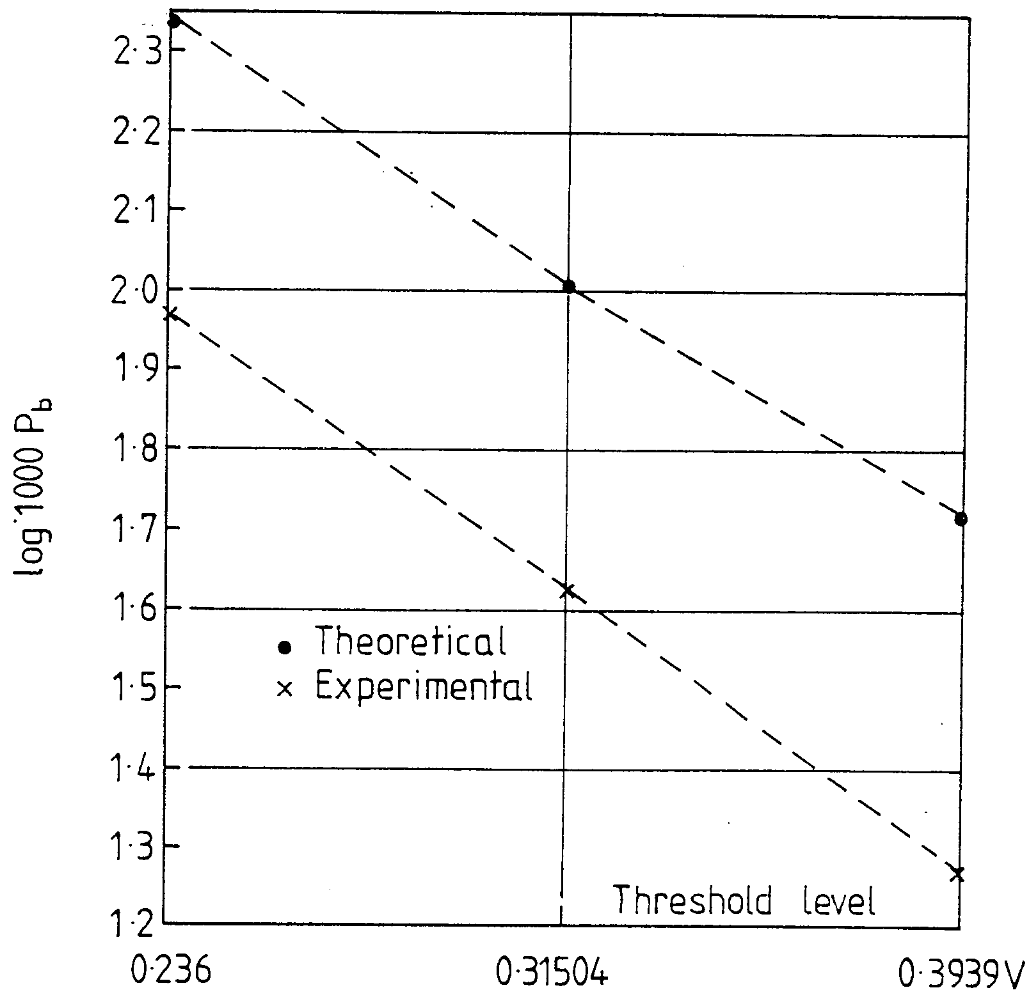


Fig 6.15b Bit error-probability as functions of the threshold levels for FH-MFSK system in the presence of A.W.G.N. for S/N = 3·3 dB, L = 12

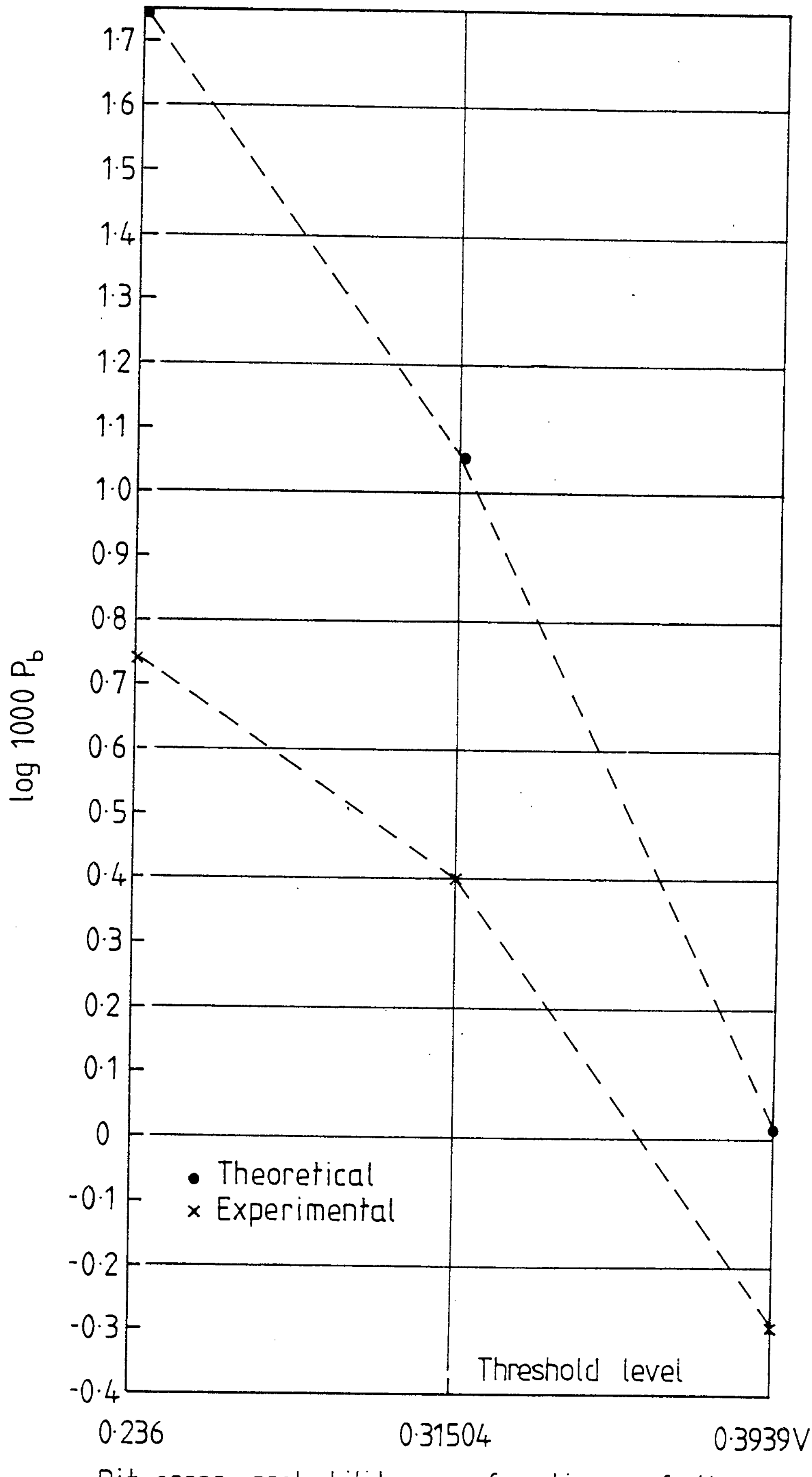


Fig 6.15c Bit error-probability as functions of the threshold levels for FH-MFSK system in the presence of A.W.G.N. for S/N = 7·4 dB, L =12

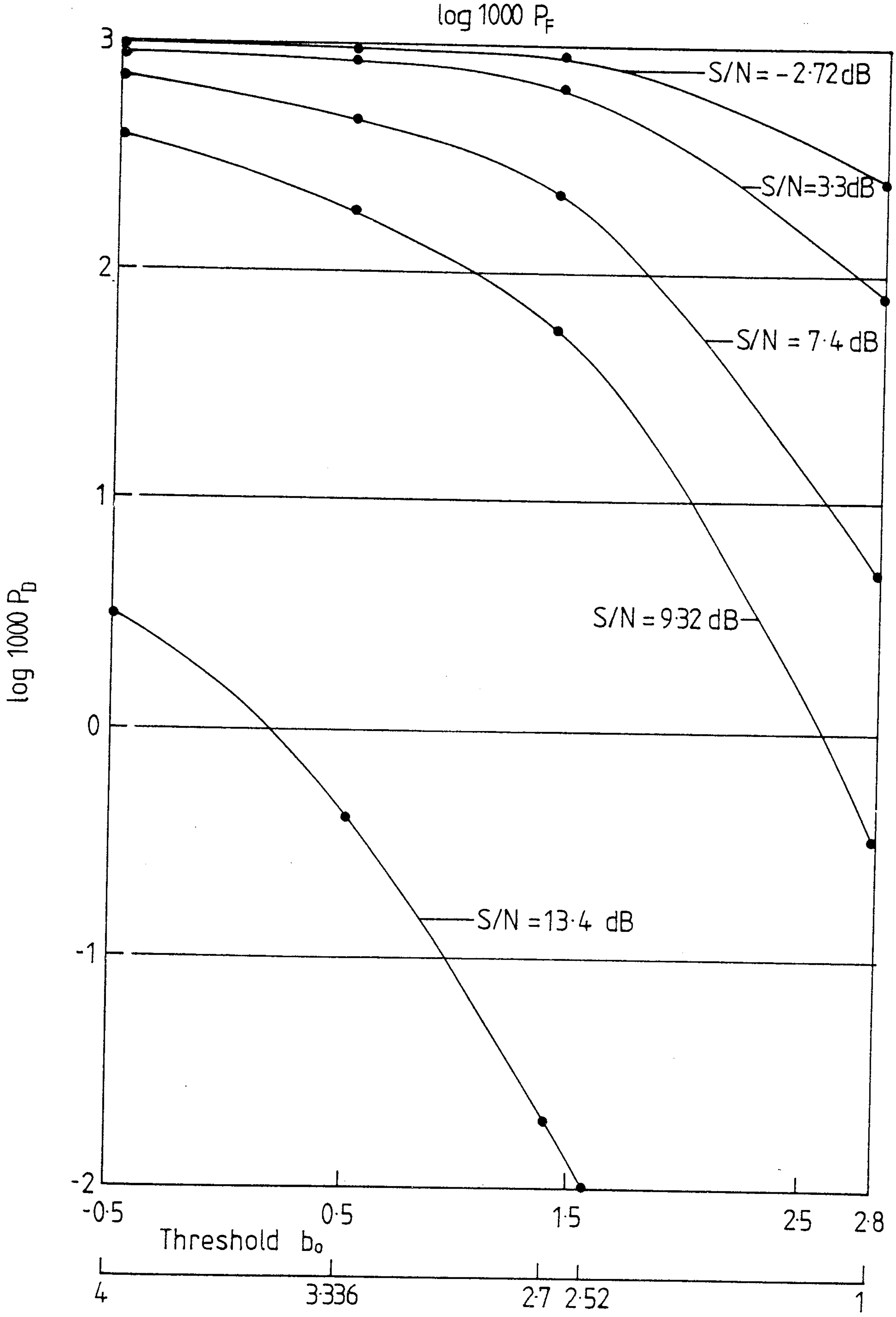


Fig 6.16 Receiver operating theoretical curves for non-coherent O.O.K.

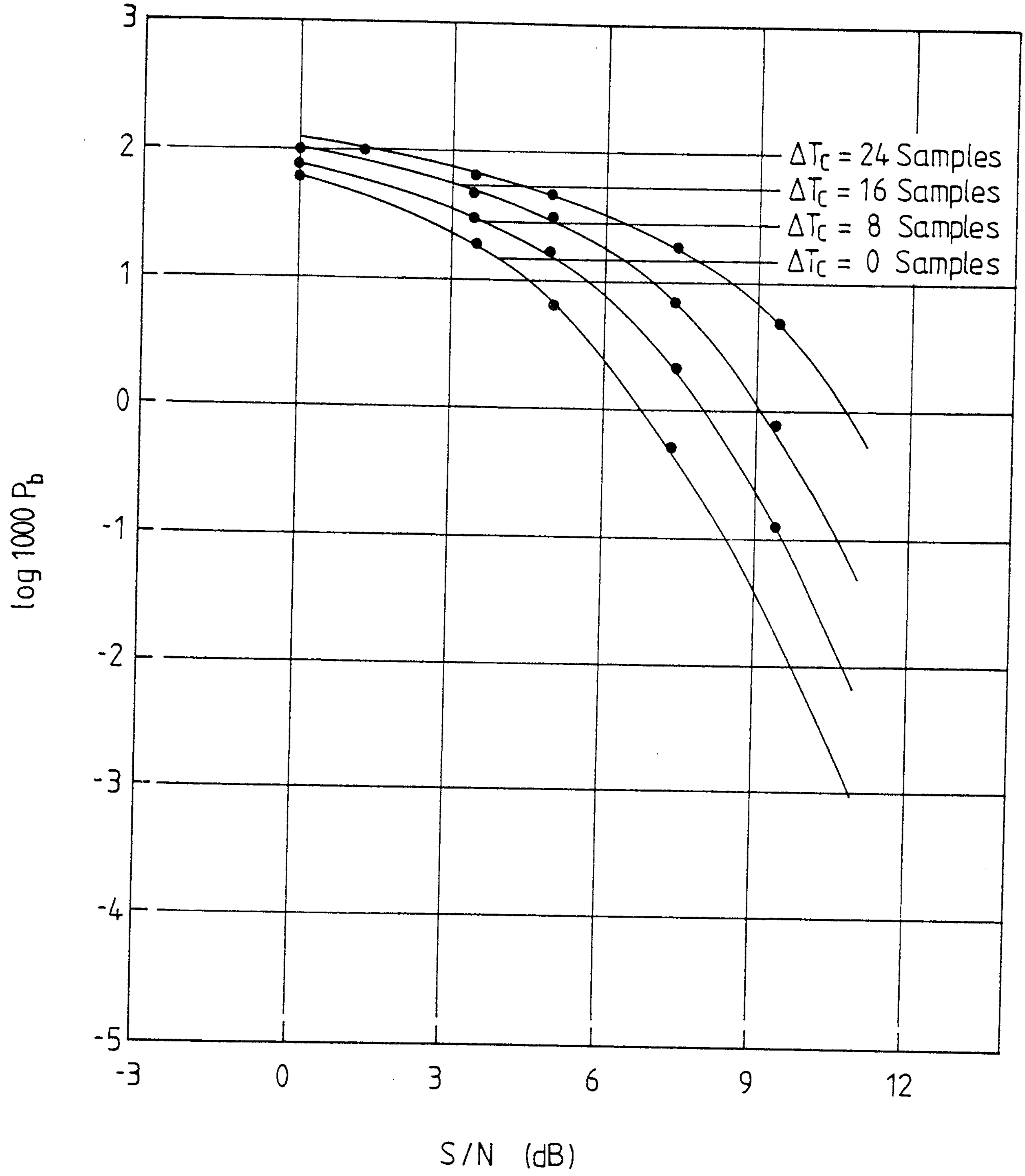


Fig 6.17 Experimental error-rate of FH-MFSK in the presence of A.W.G.N. with deleted or delayed samples for L=12 and threshold level = 0·3938 V.

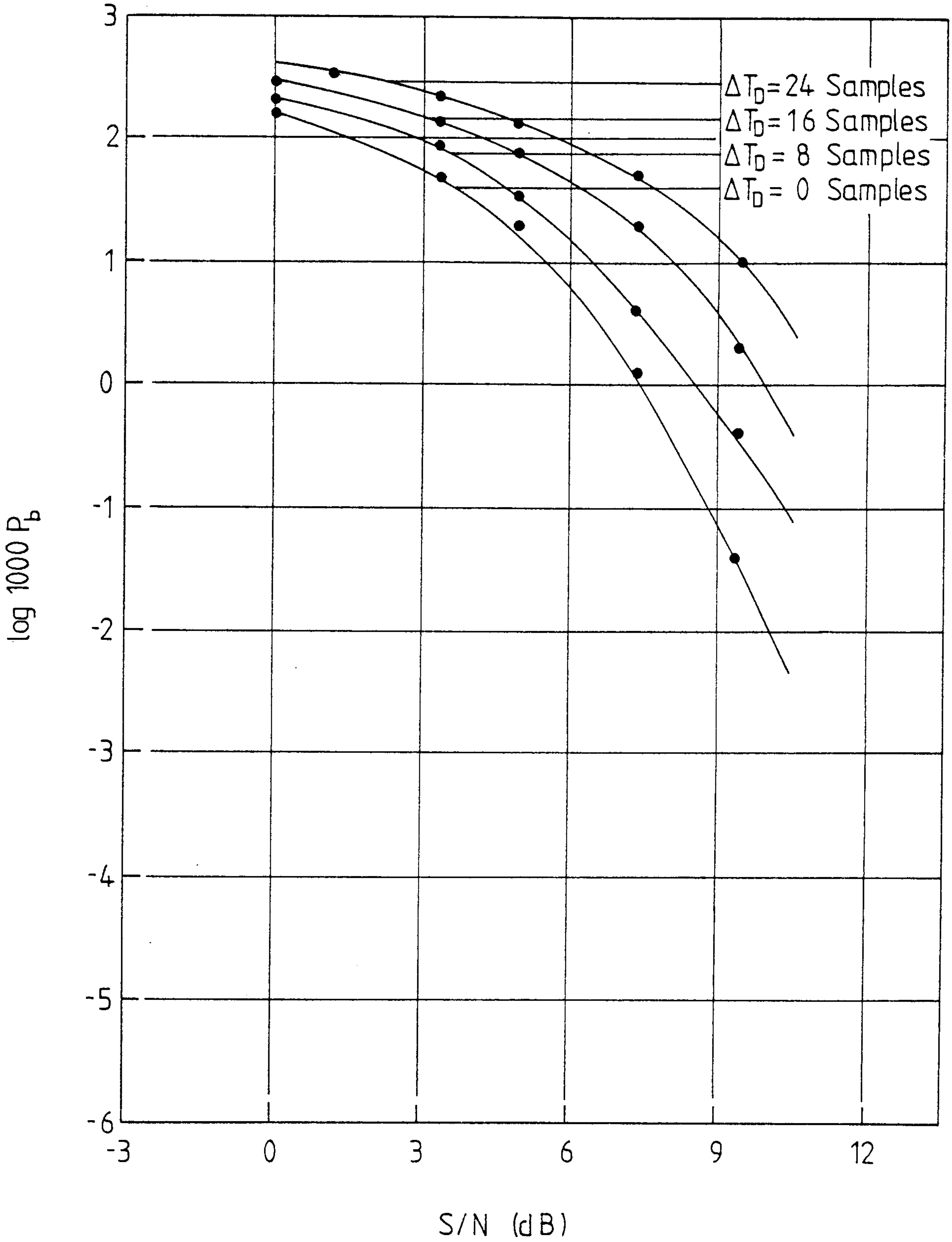


Fig 6.18 Theoretical error-rate of FH-MFSK in the presence of A.W.G.N. with deleted or delayed samples for L=12 and threshold level = 0·3938V.

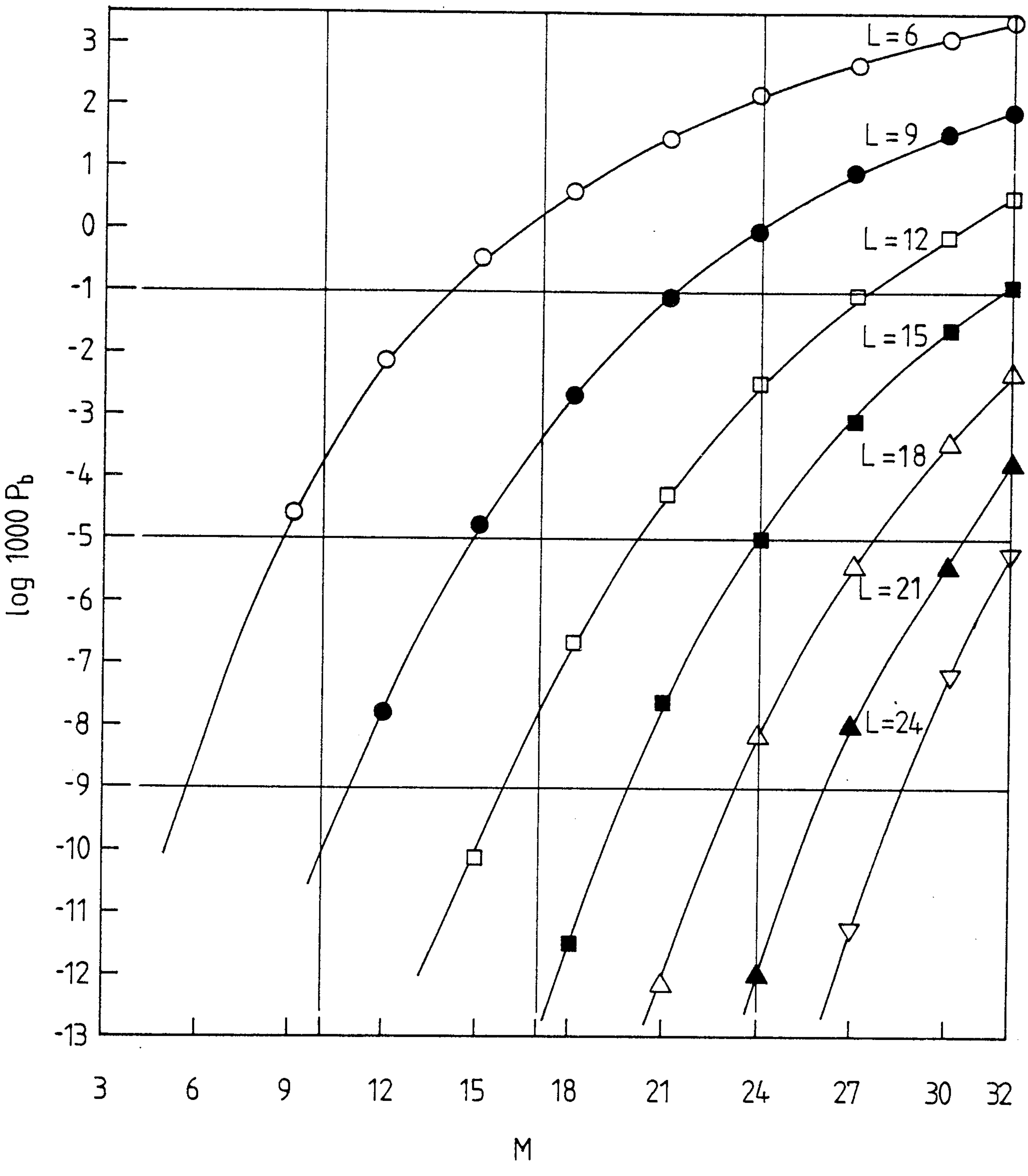


Fig 6.19 Bit error-probability as functions of the number of users for a FH-MFSK system with $Q=32$, $S/N=\infty$

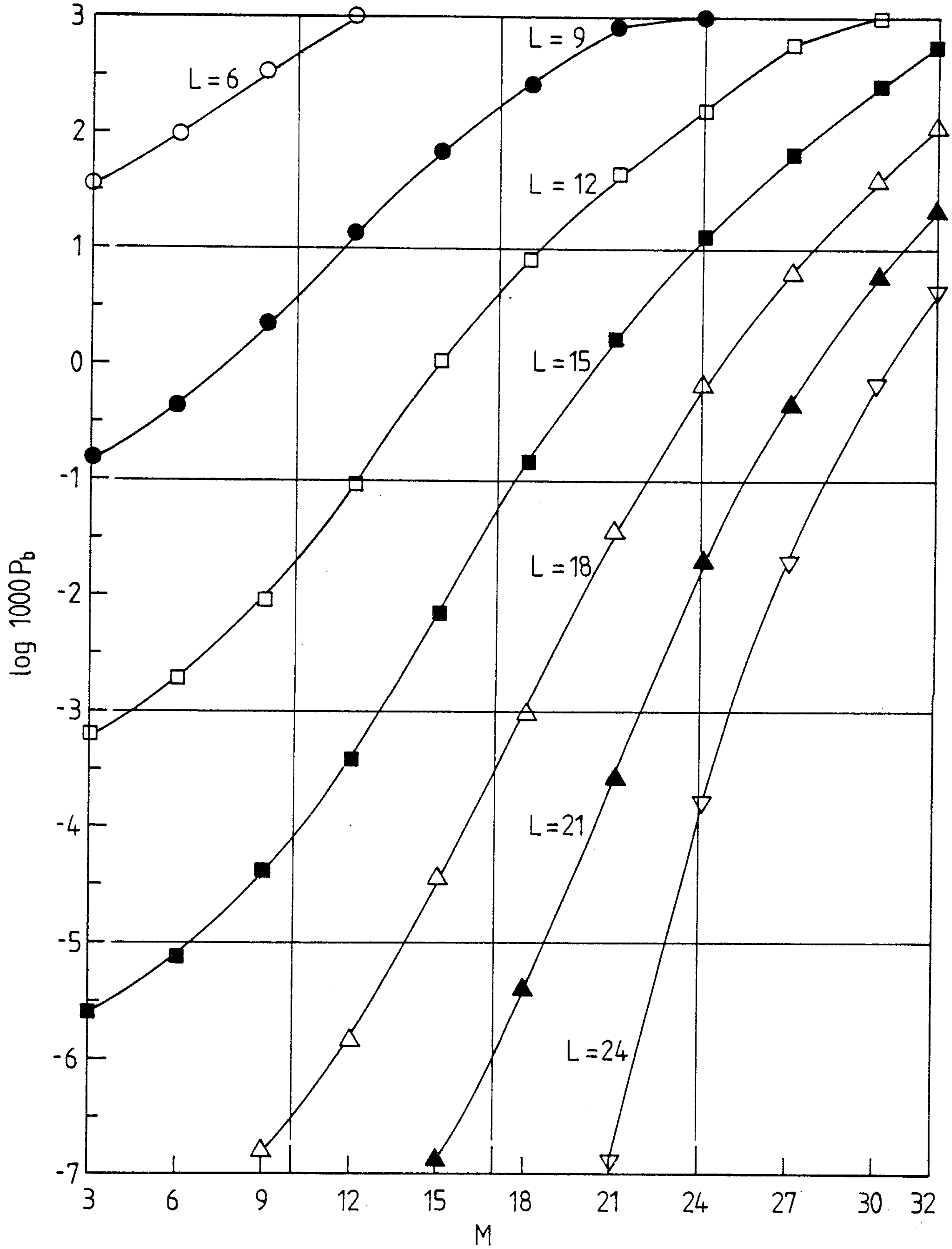


Fig 6.20a. Bit error-probability as functions of the number of users for a FH-MFSK system with Q=32, in the presence of A.W.G.N. for S/N=13·4 dB, threshold level 0·2363V.

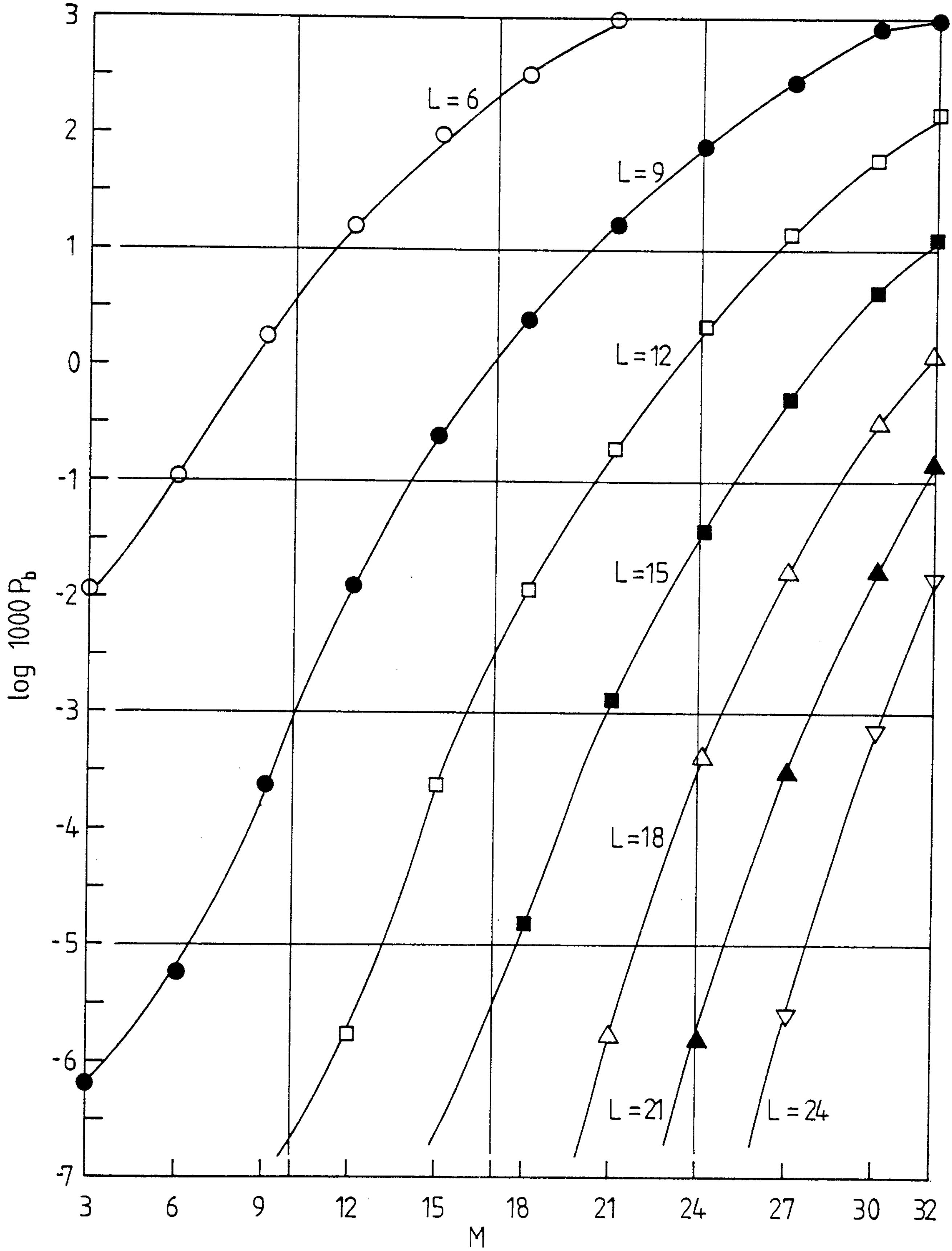


Fig 6.20b Bit error-probability as functions of the number of users for a FH-MFSK system with Q = 32, in the presence of A.W.G.N. for S/N = 13·4 dB and threshold level = 0·31504 V

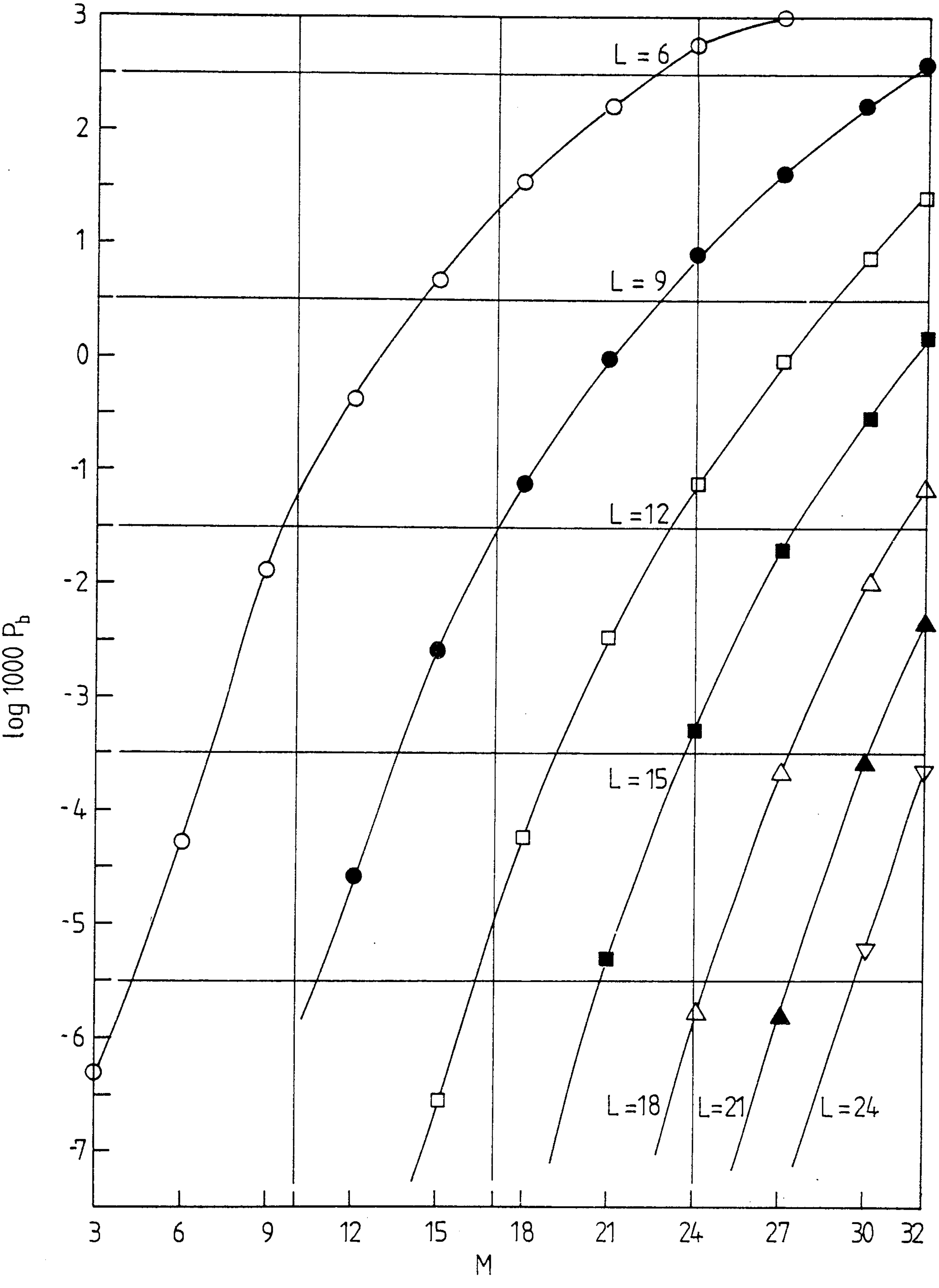


Fig 6.20c Bit error-probability as functions of the number of users for a FH-MFSK system with Q=32, in the presence of A.W.G.N. for S/N = 13·4 dB, threshold level = 0·3939V

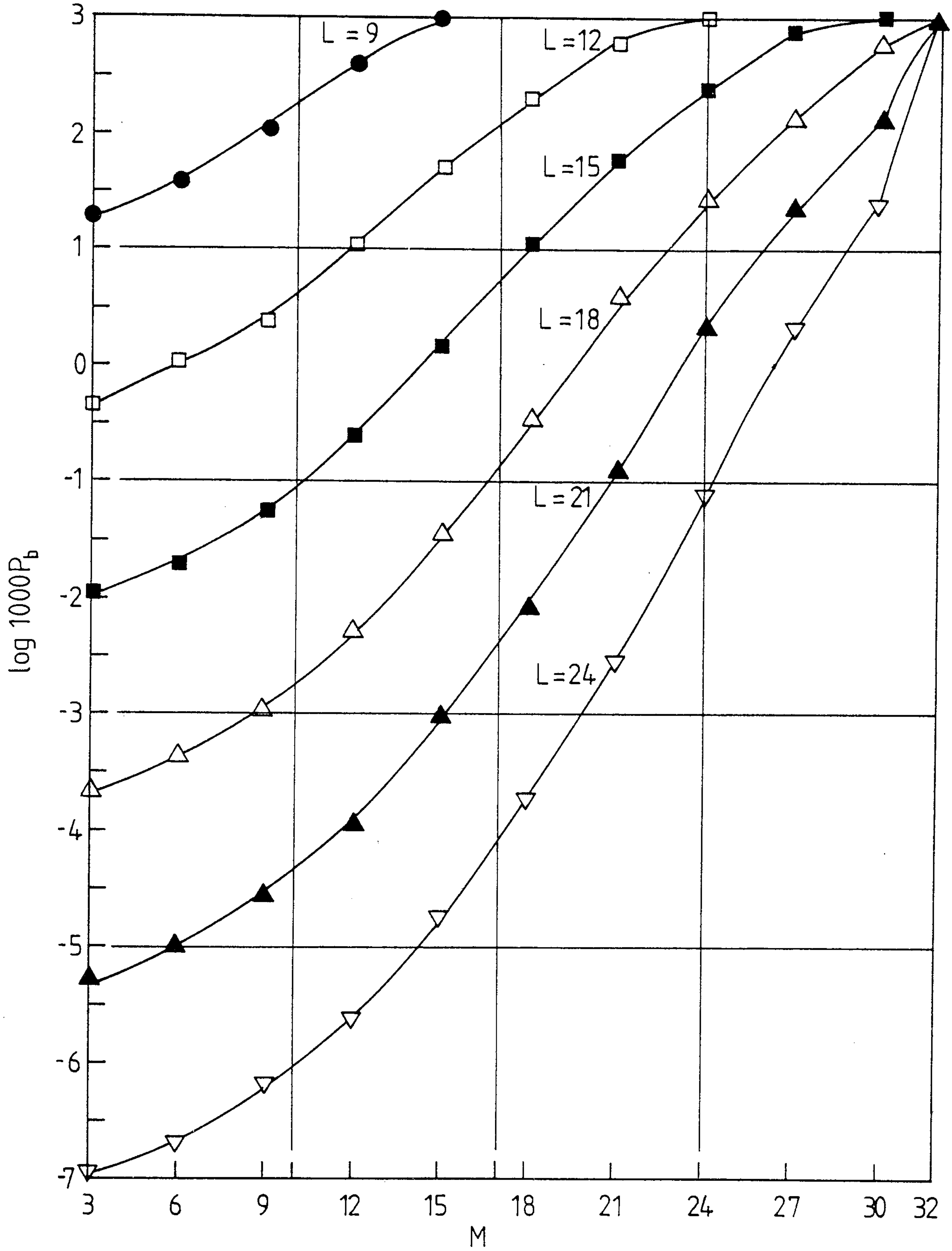


Fig 6.21a Bit error-probability as functions of the number of users for a FH-MFSK system with Q = 32, in the presence of A.W.G.N. for S/N = 9·32 dB, threshold level = 0·31504 V

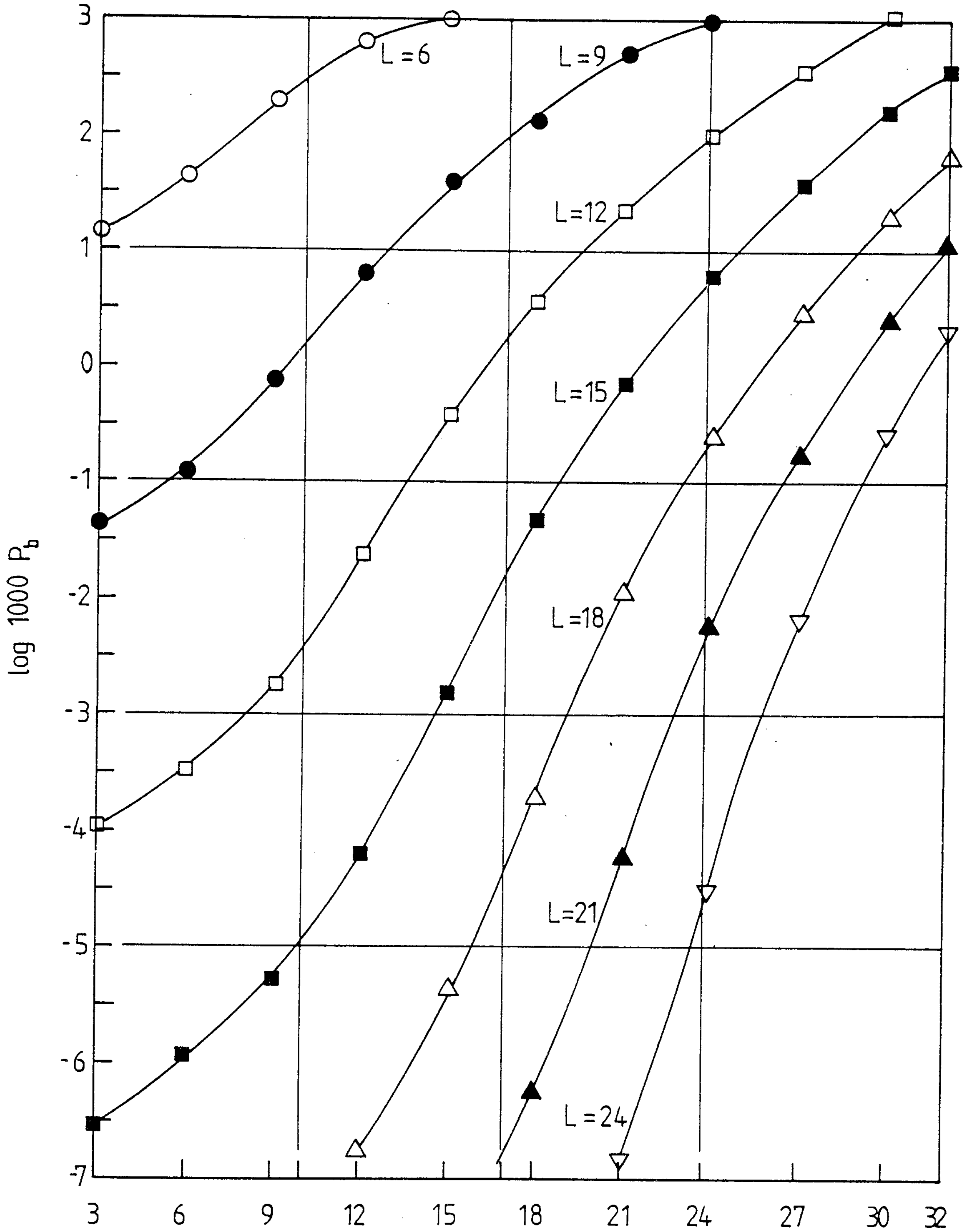


Fig 6 .21b Bit error-probability as functions of the number of users for a FH-MFSK system with Q = 32, in the presence of A.W.G.N. for S/N = 9·32 dB, threshold level = 0·3939 V

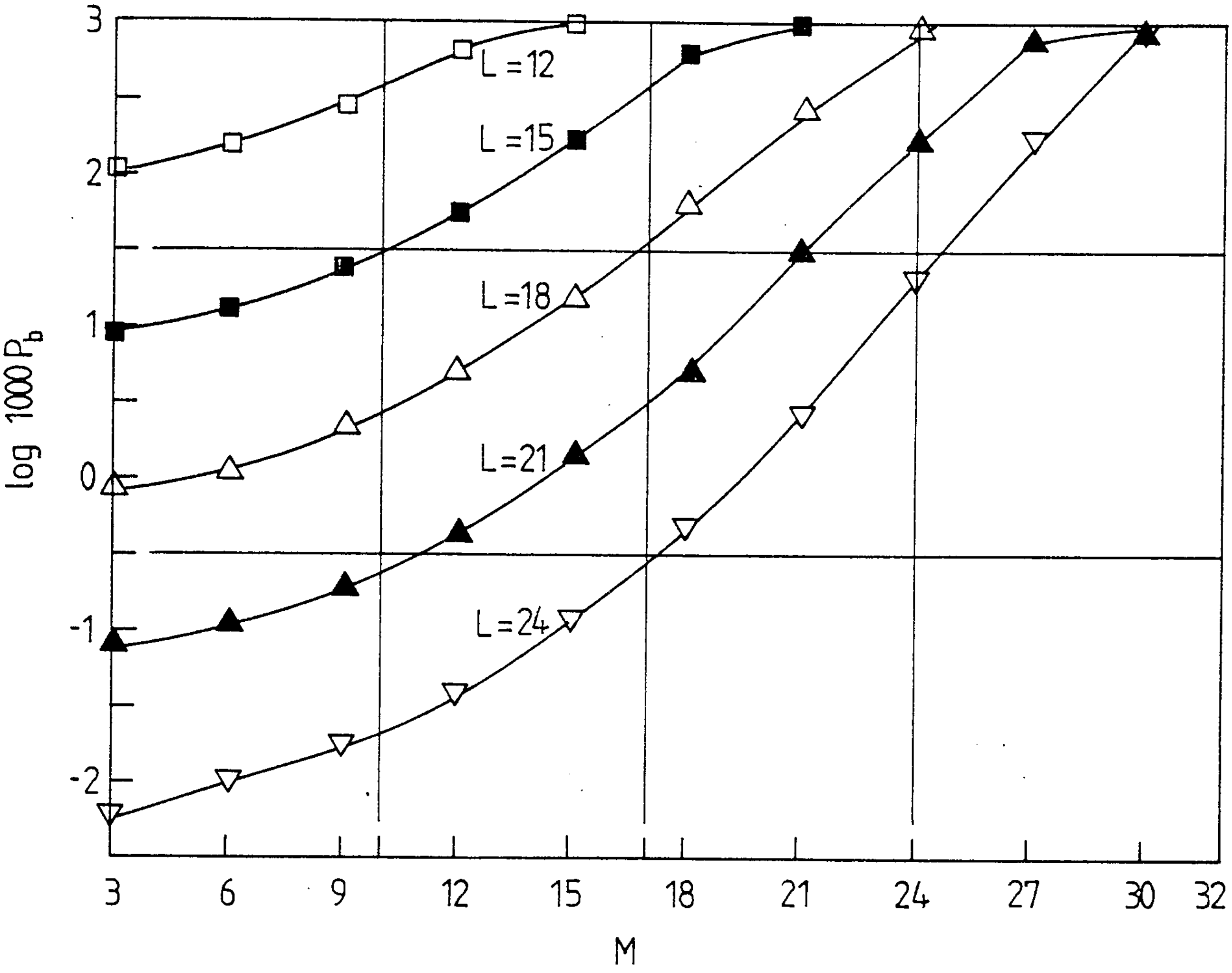


Fig 6.22a Bit error-probability as functions of the number of users for a FH-MFSK system with Q = 32, in the presence of A.W.G.N. for S/N = 7·4 dB , threshold level = 0·31504V

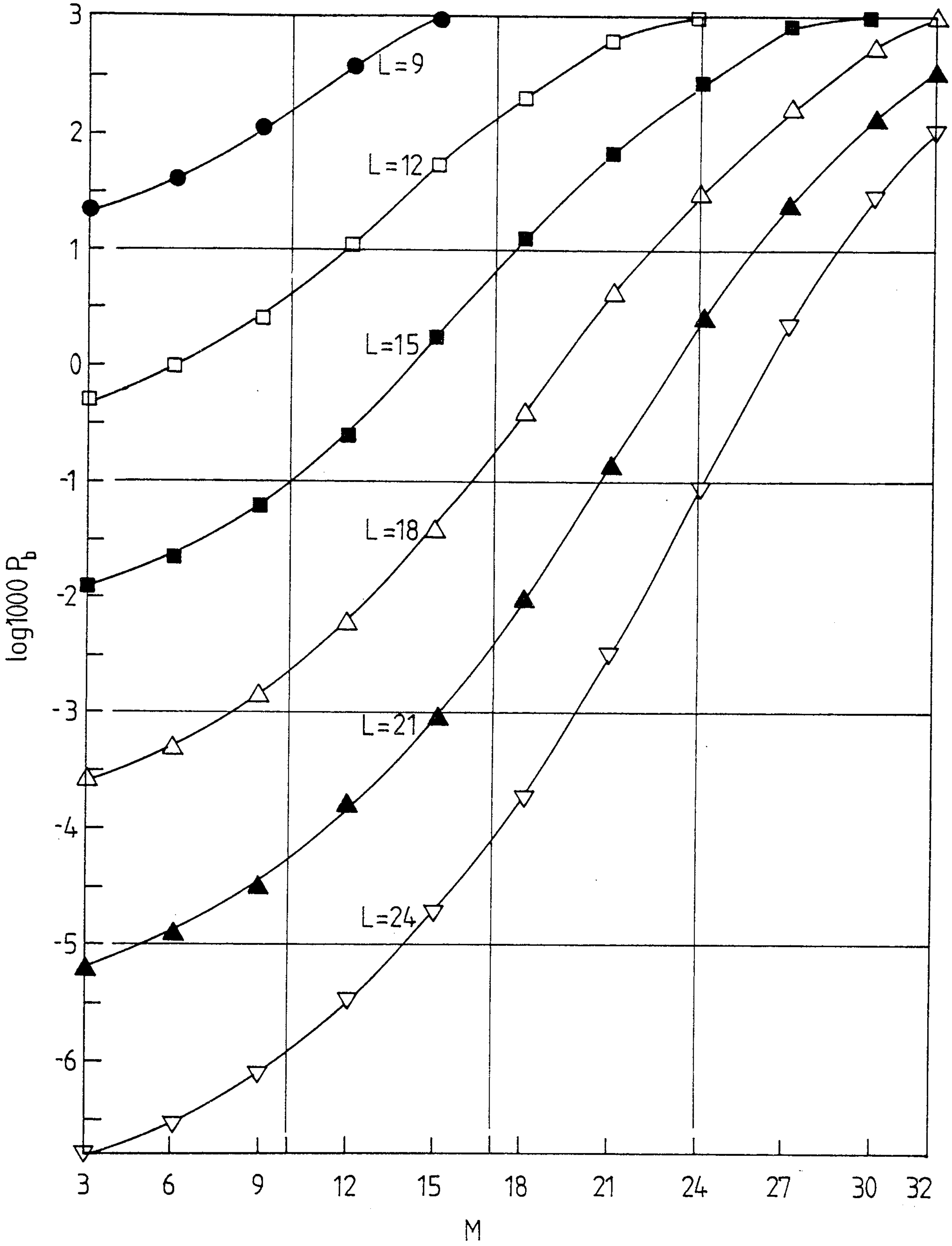


Fig 6.22b Bit error-probability as functions of the number of users for a FH-MFSK system with Q=32, in the presence of A.W.G.N. for S/N = 7·4 dB, threshold level = 0·3939 V

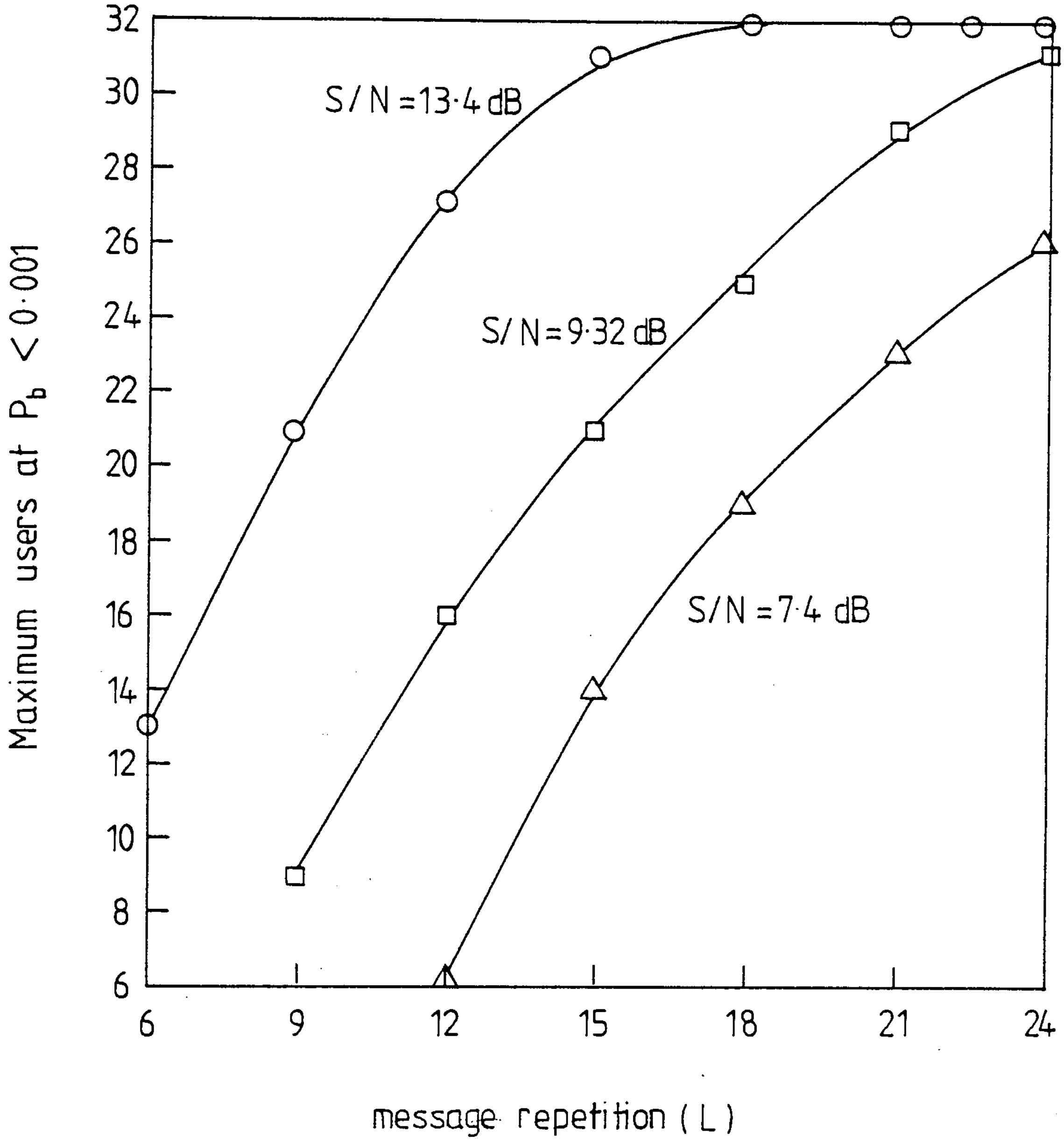


Fig 6.23 Maximum number of users versus message repetition for a FH-MFSK system with Q = 32, in the presence of A.W.G.N. for threshold level = 0·3939 V.

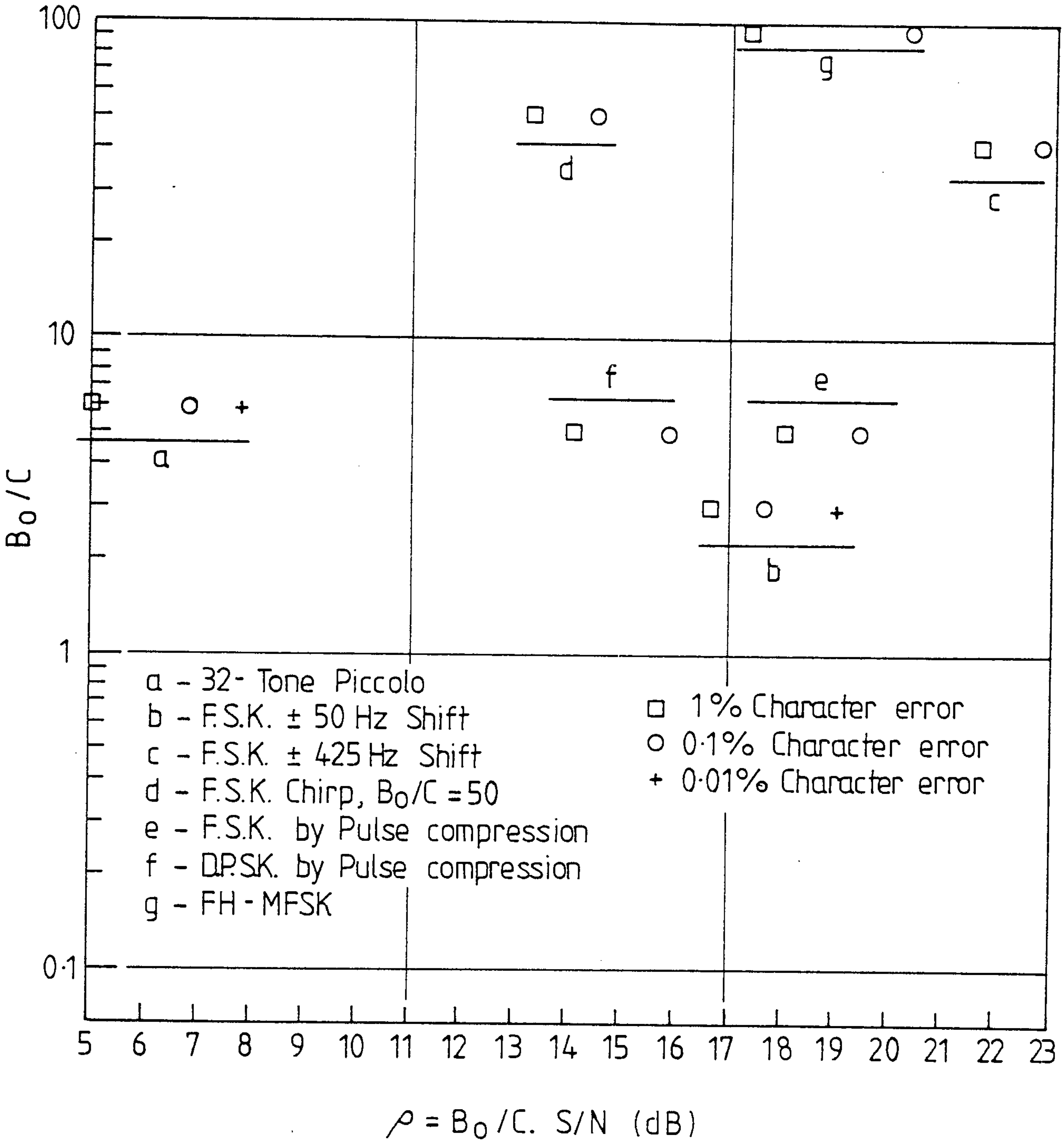


Fig 6.24 Comparison of experimental performance of various digital systems under noise conditions.

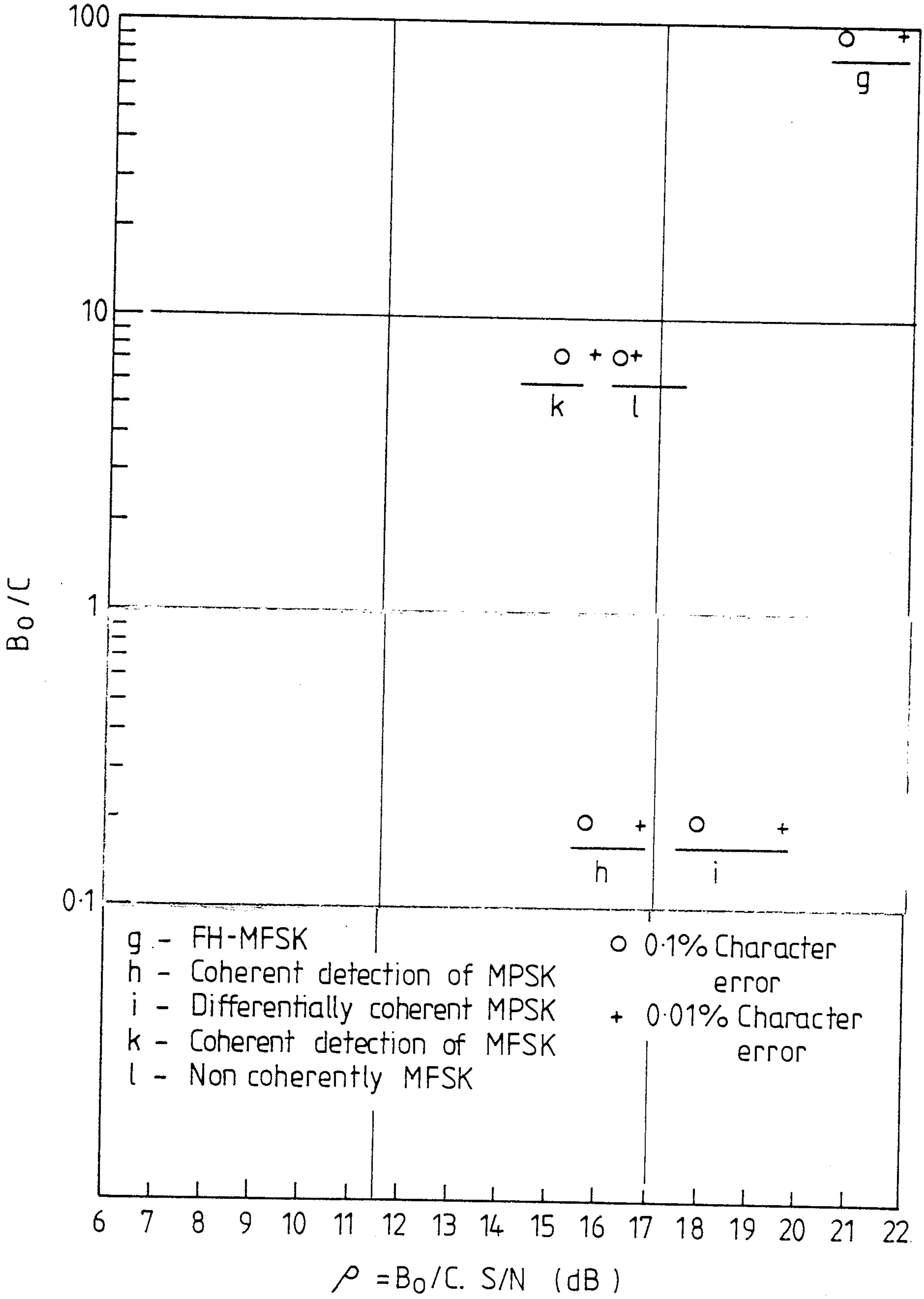


Fig 6.25 Comparison of theoretical performance of various M-ary systems under noise conditions.

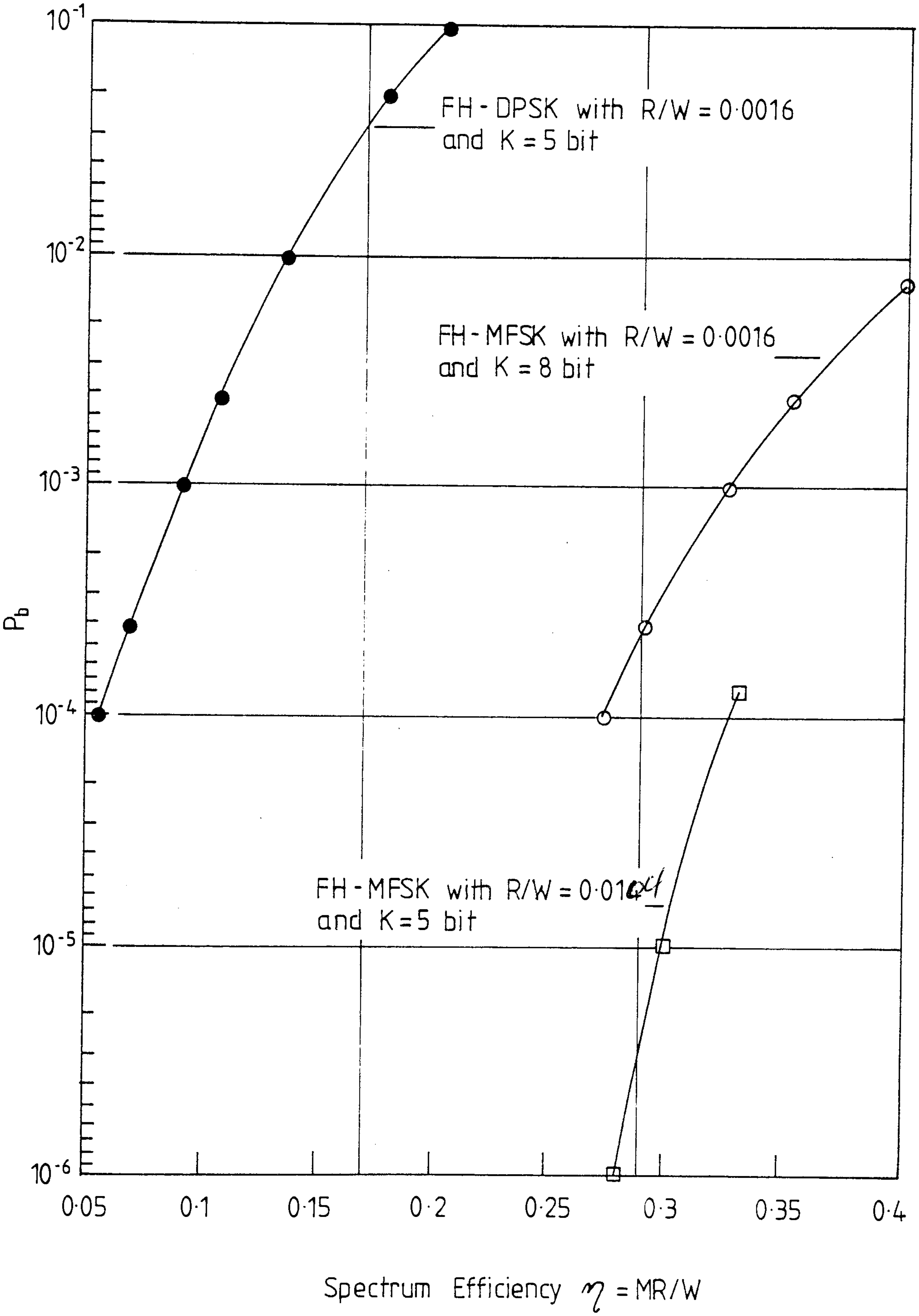


Fig 6.26 Comparison of theoretical performance of FH - MFSK and FH-DPSK under noiseless conditions.

CHAPTER SEVEN

CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

7.1 CONCLUSIONS

The aim of this research was to investigate the possibility of data transmission using FH-MFSK techniques in the presence of A.W.G.N. The FH-MFSK system may be implemented utilising SAW devices and digital techniques, which have now reached a high level of development that allows the design and construction of a compact system at a relatively low cost.

A systematic study was carried out on types of addresses in order to minimize the interference between users which share the FH-MFSK system. It was found that Einarsson addresses have minimum cross-correlation between users as mentioned in section 3.5. Table 3.1 may be used to find Einarsson addresses for 32 users.

The FH-MFSK transmitter was implemented on a microprocessor by using a table look-up method, and then tested. The results of testing were found acceptable, since no error was detected in the generation of the required frequency corresponding to the transmitted data.

An optimum spectral analysis for FH-MFSK signals in the presence of A.W.G.N., was designed using the FFT as mentioned in section 4.5. The FFT was also implemented using a microprocessor. Integer numbers were used for the FFT implementation for simplicity and to speed up the calculation. The accuracy of the FFT using this implementation was found to be ±6%. The effects produced by deletion and delay of samples of FH-MFSK signals on the output of the FFT, were determined. These effects simply degrade the output of FFT. The amount of degradation was found to be proportional to the number of deleted and delayed samples. Expressions for the FFT in the presence of deleted and delayed samples were derived. These expressions were used to determine the theoretical capability of the FFT in order to make a comparison with the experimental capability. Only small differences between them were found.

An experimental system was used to determine the error-rate performance of the FH-MFSK system. This system was operated under the supervision of the microprocessor system, thus giving a high level of versatility to the system. The experiments showed that if message repetition and threshold level values increase, the error will be decreased. For signal-to-noise ratios greater than 9.32 dB with message repetition = 12 and threshold level = 0.3939V, the bit error-probability was found less than 10^{-3}.

All experimental results of error-rate were compared with theoretical results calculated using the analysis for the FH-MFSK system derived in section 3.6.3. Experimental results were found to indicate slightly inferior performance to that determined theoretically as mentioned in section 6.6.

Error-rate results concerning deleted and delayed samples of the received signal show that the FH-MFSK system is tolerant to the deletion and delaying of samples in the presence of A.W.G.N. The performance of the FH-MFSK system in the presence of deletion or delaying of samples highlighted the need for a compensation in the level of input signal if a good performance with a relatively low signal-to-noise ratio is required. The amount of the compensation for a received signal is equal to the amount of the degradation in the output of the FFT. However, the effect of deletion and delaying of samples on the received FH-MFSK signal is simply to degrade its amplitude as mentioned in section 6.4.1.

The performance of the FH-MFSK system with multi-users in the presence of A.W.G.N. was determined theoretically using the analysis derived in section 3.6.3. It was calculated as functions of signal-to-noise ratio, number of repetition messages, threshold level and bit error-probability. If either the signal-to-noise ratio, the number of message repetitions or the threshold level increases, the number of users will be increased for a given value of bit error-probability. For signal-to-noise ratio greater than 13.4 dB, the number of message repetitions = 15, threshold level = 0.3939V and bit error-probability less than 10^{-3}, the number of users which can share the same system, was found to be 32.

In comparison with other systems, the FH-MFSK system performed somewhat better than other systems as shown in figures 6.24 and 6.25. The comparison indicates that the FH-MFSK system has a high processing gain (Bo/C) and high signal energy per bit/Noise power per unit bandwidth (ρ). However, the spectrum efficiency of the FH-MFSK system is larger than for the FH-DPSK system with linear combining, as shown in Fig. 6.26.

7.2 SUGGESTIONS FOR FURTHER WORK

It is hoped that this research will lead to further development of the system. There are several areas in which further work is required.

(1) A second stage of decoding of the complete rows in the detection matrix can improve the performance in order to minimize the error[78]. In a previous analysis of the FH-MFSK system, it was assumed that if there is more than one complete row the receiver has no way of finding the correct message and has to choose one of them at random.

(2) The performance of FH-MFSK system for data transmission has been established under conditions of received signals in the presence of A.W.G.N. However, in many radio applications the channel characteristics are often strongly non-stationary. These communication channels are known as fading channels as mentioned in section (2.4). Thus it is of practical interest to establish the performance of FH-MFSK for data transmission under simulated conditions of signal fading. The experimental system described in Chapter 5 may be modified to investigate fading channel effects.

(3) The performance benefits associated with the use of any spread spectrum techniques derive from the correlation properties of the spreading codes used[10,23,24]. Obtaining these benefits requires that there be synchronization between the code arriving at the receiver and the code generated locally. This synchronization is a critical feature of a practical system. The synchronization problem of spread spectrum must be investigated in two parts: acquisition and tracking. Acquisition is a process of adjusting the relative timing and relative frequency of the reference signal and the received signal to the pull-in range of the tracking synchronizer. After a receiver has acquired code synchronization, it can begin tracking in order to maintain its synchronization. Thus a

search for a synchronization scheme would be necessary.

(4) The FH-MFSK system with multi-users was studied theoretically as mentioned in section 6.5. It is unlikely at least initially that the practical testing of multi-users will be possible. This being the case, simulation of the FH-MFSK with multi-users in the presence of A.W.G.N. and fading channel, would probably be the only way the system could be evaluated. The results obtained from simulation would be invaluable in producing an efficient practical system.

(5) Coding is used in digital radio communication to reduce the error for both fixed and mobile links. The incorporation some types of coding into an FH-MFSK system would be possible.

(6) A rectangular window was used in the FFT implementation. To improve the spectral anlaysis of FH-MFSK signals, a window which has lower side lobes in its frequency response should be chosen. Thus it is of practical interest to find the optimum window for spectral analysis of FH-MFSK signals.

(7) A low-rate data transmission system using the FH-MFSK technique may be designed as a real time system, using the FH-MFSK transmitter and the decoder as described in Chapter 5, together with a high speed

FFT processor. High speed FFT processors are now available, for example the TD 1010J, produced by TRW Inc.[70], which takes 0.26 mseconds to perform a 64 point transformation.

APPENDICES

APPENDIX A

Elements of $GF(2^5)$

The Galois field of (2^5) element may be formed as field of polynomial over $GF(2)$ modulo an irreducible polynomial $P(z)$ of degree 5,(5,29). The polynomial $P(z) = z^5+z^3+1$ is irreducible.

If the element $b = z = 00010$ is primitive, the non-zero field elements can be found as follows:

(1) $b^0 = z^0 = 1$

(2) $b^1 = z^1 = z$

(3) $b^2 = b.b = z^2$

(4) $b^3 = b.b^2 = z^3$

(5) $b^4 = b.b^3 = z^4$

(6) $b^5 = b.b^4 = z^5$

$= 2z^5 + z^3 + 1$ (adding $P(z)$)

$= z^3 + 1$

(7) $b^6 = b.b^5 = z(z^3+1) = z^4 + z$

(8) $b^7 = b.b^6 = z(z^4+z) = z^5 + z^2$

$= 2z^5 + z^3 + z^2 + 1$ (adding $P(z)$)

$= z^3 + z^2 + 1$

(9) $b^8 = b.b^7 = z(z^3+z^2+1) = z^4 + z^3 + z$

(10) $b^9 = b.b^8 = z(z^4+z^3+z) = z^5 + z^4 + z^2$

$= 2z^5 + z^4 + z^3 + z^2 + 1$ (adding $P(z)$)

$= z^4 + z^3 + z^2 + 1$

(11) $b^{10} = b.b^9 = z(z^4+z^3+z^2+1) = z^5 + z^4 + z^3 + z$

$= 2z^5 + z^4 + 2z^3 + z + 1$ (adding $P(z)$)

$= z^4 + z + 1$

(12) $b^{11} = b.b^{10} = z(z^4+z+1) = z^5 + z^2 + z$

$= 2z^5 + z^3 + z^2 + z + 1$ (adding $P(z)$)

$= z^3 + z^2 + z + 1$

(13) $b^{12} = b.b^{11} = z(z^3+z^2+z+1) = z^4 + z^3 + z^2 + z$

(14) $b^{13} = b.b^{12} = z(z^4+z^3+z^2+z) = z^5 + z^4 + z^3 + z^2$

$= 2z^5 + z^4 + 2z^3 + z^2 + 1$ (adding $P(z)$)

$= z^4 + z^2 + 1$

(15) $b^{14} = b.b^{13} = z(z^4+z^2+1) = z^5 + z^3 + z$

$= 2z^5 + 2z^3 + z + 1$ (adding $P(z)$)

$= z + 1$

(16) $b^{15} = b.b^{14} = z(z+1) = z^2 + z$

(17) $b^{16} = b.b^{15} = z(z^2+z) = z^3 + z^2$

(18) $b^{17} = b.b^{16} = z(z^3+z^2) = z^4 + z^3$

(19) $b^{18} = b.b^{17} = z(z^4+z^3) = z^5 + z^4$

$= 2z^5 + z^4 + z^3 + 1$ (adding $P(z)$)

$= z^4 + z^3 + 1$

(20) $b^{19} = b.b^{18} = z(z^4+z^3+1) = z^5 + z^4 + z$

$= 2z^5 + z^4 + z^3 + z + 1$ (adding $P(z)$)

$= z^4 + z^3 + z + 1$

(21) $b^{20} = b.b^{19} = z(z^4+z^3+z+1) = z^5 + z^4 + z^2 + z$

$= 2z^5 + z^4 + z^3 + z^2 + z + 1$ (adding $P(z)$)

$= z^4 + z^3 + z^2 + z + 1$

(22) $b^{21} = b.b^{20} = z(z^4+z^3+z^2+z+1) = z^5 + z^4 + z^3 + z^2 + z$

$= 2z^5 + z^4 + 2z^3 + z^2 + z + 1$ (adding $P(z)$)

$= z^4 + z^2 + z + 1$

(23) $b^{22} = b.b^{21} = z(z^4+z^2+z+1) = z^5 + z^3 + z^2 + z$

$= 2z^5 + 2z^3 + z^2 + z + 1$ (adding $P(z)$)

$= z^2 + z + 1$

(24) $b^{23} = b.b^{22} = z(z^2+z+1) = z^3 + z^2 + z$

(25) $b^{24} = b.b^{23} = z(z^3+z^2+z) = z^4 + z^3 + z^2$

(26) $b^{25} = b.b^{24} = z(z^4+z^3+z^2) = z^5 + z^4 + z^3$

$= 2z^5 + z^4 + 2z^3 + 1$ (adding $P(z)$)

$= z^4 + 1$

(27) $b^{26} = b.b^{25} = z(z^4+1) = z^5 + z$

$= 2z^5 + z^3 + z + 1$ (adding $P(z)$)

$= z^3 + z + 1$

(28) $b^{27} = b.b^{26} = z(z^3+z+1) = z^4 + z^2 + z$

(29) $b^{28} = b.b^{27} = z(z^4+z^2+z) = z^5 + z^3 + z^2$

$= 2z^5 + 2z^3 + z^2 + 1$ (adding $P(z)$)

$= z^2 + 1$

(30) $b^{29} = b.b^{28} = z(z^2+1) = z^3 + z$

(31) $b^{30} = b.b^{29} = z(z^3+z) = z^4 + z^2$

(32) $b^{31} = b.b^{30} = z(z^4+z^2) = z^5 + z^3$

$= 2z^5 + 2z^3 + 1$ (adding $P(z)$)

$= 1$

$= b^0$

APPENDIX B

Fourier Transform of a Sinusoidal Signal in the Presence of Deleted Samples Effect

Let

$$x(t) = \begin{cases} A\cos 2\pi f_n t & 0 \leq t \leq T_o - \delta' \\ 0 & \text{elsewhere} \end{cases} \qquad \text{(B-1)}$$

where

A is the amplitude

f_n is the nth frequency

T_o is the duration of the FFT window

δ' is the deleted duration of the signal $x(t)$

The Fourier transform of $x(t)$ is

$$X(f) = \int_0^{T_o - \delta'} x(t)\, e^{-j2\pi ft}\, dt \qquad \text{(B-2)}$$

Let

$$K_o T_o = T_o - \delta' \qquad \text{(B-3)}$$

where

$K_o T_o$ is the undeleted duration of the signal $x(t)$

Hence,

$$X(f) = \int_0^{K_o T_o} x(t)\, e^{-j2\pi ft}\, dt \qquad \text{(B-4)}$$

substituting for $x(t)$ in equation (B-4) we get

$$X(f) = A\int_0^{K_oT_o} \cos 2\pi f_n t \; e^{-j2\pi ft}\, dt$$

$$X(f) = \frac{A}{2}\int_0^{K_oT_o} \left[e^{-j2\pi(f-f_n)t} + e^{-j2\pi(f+f_n)t}\right] dt \tag{B-5}$$

$$X(f) = \frac{A}{2}\left[\frac{e^{-j2\pi(f-f_n)K_oT_o}}{-j2\pi(f-f_n)} + \frac{e^{-j2\pi(f+f_n)K_oT_o}}{-j2\pi(f+f_n)}\right]$$

$$- \frac{A}{2}\left[\frac{1}{-j2\pi(f-f_n)} + \frac{1}{-j2\pi(f+f_n)}\right] \tag{B-6}$$

Simplifying

$$X(f) = \frac{AK_oT_o}{2}\left[e^{-j\pi(f-f_n)K_oT_o}\,\frac{\sin\pi(f-f_n)K_oT_o}{\pi(f-f_n)K_oT_o}\right.$$

$$\left. + e^{-j\pi(f+f_n)K_oT_o}\,\frac{\sin\pi(f+f_n)K_oT_o}{\pi(f+f_n)K_oT_o}\right]$$

for $f = f_n$

$$\therefore \quad X(f_n) = \frac{AK_oT_o}{2} \tag{B-7}$$

Let $\bar{\delta} = N_D\, T_s$

where

N_D is the number of deleted samples

$T_s = \frac{T_o}{64}$ is the sampling period.

Substituting for $\bar{\delta}$ in (B-3) and (B-7) we get

$$X(f_n) = \frac{T_o A}{2} \left[1 - \frac{N_D}{64}\right] \qquad \text{(B-8)}$$

APPENDIX C

Fourier Transform of a Sinusoidal Signal in the Presence of Delayed Samples Effect

Let

$$x(t) = \begin{cases} A\cos 2\pi f_n(t-\delta) & \delta \leq t \leq T_o \\ 0 & \text{elsewhere} \end{cases} \qquad \text{(C-1)}$$

where

A is the amplitude

f_n is the nth frequency

T_o is the duration of the FFT window

δ is the delayed duration of the signal $x(t)$

The Fourier transform of $x(t)$ is

$$X(f) = \int_{\delta}^{T_o} x(t)\, e^{-j2\pi ft}\, dt \qquad \text{(C-2)}$$

Substituting for $x(t)$ in (C-2) we get

$$X(f) = A\int_{\delta}^{T_o} \cos 2\pi f_n(t-\delta)\, e^{-j2\pi ft}\, dt$$

$$X(f) = \frac{A}{2} \int_{\delta}^{T_o} \left[e^{-j2\pi f_n\delta}\, e^{-j2\pi(f-f_n)t} + e^{j2\pi f_n\delta}\, e^{-j2\pi(f_n+f)t} \right] dt$$

(C-3)

$$X(f) = \frac{A}{2}\left[e^{-j2\pi f_n\delta}\left(\frac{e^{-j2\pi(f-f_n)T_o} - e^{-j2\pi(f-f_n)\delta}}{-j2\pi(f-f_n)}\right)\right]$$

$$+\frac{A}{2}\left[e^{j2\pi f_n\delta}\left(\frac{e^{-j2\pi(f+f_n)T_o} - e^{-j2\pi(f+f_n)\delta}}{-j2\pi(f+f_n)}\right)\right] \qquad \text{(C-4)}$$

Simplifying

$$X(f) = \frac{A(T_o-\delta)}{2}\left[e^{-j2\pi f_n\delta}e^{-j2\pi(f-f_n)(T_o+\delta)/2}\,\frac{\sin 2\pi(f-f_n)(T_o-\delta)/2}{2\pi(f-f_n)(T_o-\delta)/2}\right.$$

$$\left.+\, e^{j2\pi f_n}\, e^{-j2\pi(f+f_n)(T_o+\delta)/2}\,\frac{\sin 2\pi(f+f_n)(T_o-\delta)/2}{2\pi(f+f_n)(T_o-\delta)/2}\right]$$

$$\text{(C-5)}$$

for $f = f_n$

$$\therefore\ X(f_n) = \frac{A(T_o-\delta)}{2}\ e^{-j2\pi f_n\delta} \qquad \text{(C-6)}$$

Let

$$\delta = N_A T_s$$

where

N_A is the number of delayed samples

$T_s = \frac{T_o}{64}$ is the sampling period.

Substituting for δ in (C-6) we get

$$X(f_n) = \frac{AT_o}{2}\left[1 - \frac{N_A}{64}\right] e^{-j2\pi f_n\delta} \qquad \text{(C-7)}$$

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