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MULTIPLE ACCESS PERFORMANCE  
OF DIRECT SEQUENCE  
SPREAD SPECTRUM SYSTEMS

(直接拡散を用いたスペクトル拡散通信方式の  
多元接続特性に関する研究)

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## PREFACE

This thesis investigates the multiple access performance of direct sequence spread spectrum systems. The communication systems considered in this thesis include both radio frequency and infrared optical wireless digital systems. This thesis consists of five chapters as follows.

Chapter 1 gives a review of previous and recent studies on the problems discussed in this thesis, and provides the background for the thesis.

Chapter 2 introduces a fundamental system model for direct sequence (DS) spread spectrum (SS) digital communication systems. Described are the modulation technique and the multiple access method that are basis throughout this thesis. It also provides the discussion on synchronization of the DS/SS systems.

Chapter 3 discusses on the radio frequency DS spread spectrum multiple access (SSMA) digital communication systems. The main theme of this chapter is concerned with the improvement of multiple access performance of DS/SSMA systems using a modified M-ary frequency shift keying (modified MFSK) as a primary modulation scheme. First, a DS/SS system which employs MFSK as primary modulation is introduced. Then, it is shown that the multiple access performance is effectively improved by modifying the frequency alignment of the MFSK symbols so as to flatten the frequency spectrum of the transmitted signal. Although it has been theoretically known that the performance may be improved by flattening the spectrum, the practical modulation

method for digital DS/SSMA systems has not been given. Numerical comparisons are also presented.

Chapter 4 deals with an application of SSMA to the in-house wireless optical digital communication link. Taking the restrictions and the characteristics of in-house wireless optical link into account, a suitable modulation technique is proposed, and its theoretical error rate expressions based on the Gaussian approximation of the co-channel users' interference and of the shot-noise are derived. Also a synchronization system for the proposed modulation technique is briefly described.

Chapter 5 summarizes the whole conclusions obtained in this thesis.

This thesis is the result of research conducted during the Doctor Course, Research of Engineering, Graduate School of Osaka University.

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## CHAPTER 1

### INTRODUCTION

Not more than a decade ago, those researches on spread-spectrum (SS) communication systems which were primarily limited to the robust military use such as anti-jam and low-probability-of-intercept signaling design were cloaked in deep secrecy. Since the adoption of several documents related to the SS communication systems on XIV-th CCIR Plenary Assembly in 1978<sup>†</sup>, however, the inviolable veil of secrecy in the discipline of SS communication systems has been drastically lifting up especially on the field of emergency distress communication, anti-multipath mobile communication, bandwidth-efficient multiple-accessing, high resolution ranging, and the like.

The cardinal concept of spectrum spreading of a transmitted signal was originally introduced by Costas [71] in 1959. SS signal is generated by first modulating the information digits or the source waveform onto the carrier with the conventional narrow-band modulation technique e.g. PSK (Phase Shift Keying), NBFM (Narrow Band Frequency Modulation), etc. [4]. This procedure is referred to as primary modulation [10]. The primarily modulated waveform is subsequently secondarily modulated to a

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<sup>†</sup> CCIR XIV-th Plenary Assembly, Doc.1/1005-E - 1/1008-E, Kyoto, June 1978.

extraordinarily wide band signal with a various method. The bandwidth expansion ratio of the secondarily modulated waveform to the primarily modulated waveform frequently reaches up to  $10^4$  [7]. This ratio is often referred to as the processing gain [10]. The well-known bandwidth expansion methods are Direct Sequence (DS), Frequency Hopping (FH), and Time Hopping (TH) [11].

The outstanding property of SS signal is that it presents a noise-like wideband waveform with extraordinarily low power density [7]. This property leads to a plenty of advantages which no other modulating method can afford such as

- (i) low-probability-of-intercept (LPI) [9],
- (ii) anti-jam receiving characteristic [73],
- (iii) ability to share the limited frequency resource with other communication business [79],
- (iv) possibility of flexible random multiple accessing (spread spectrum multiple access, SSMA) without any controller [31],
- (v) high resolutional ranging, etc. [9].

Instead of these remarkable advantages, however, some serious disadvantages on circuit complexity [72] and on difficulty of acquiring the desired signal [47] had been preventing the SS systems from applying to the civilian commercial use. Thus in 1960's, SS systems had been researched chiefly in the field of the military communication use for these reasons [7].

Later in 1970's, the great strides made on semiconductor technology had enabled those complex circuits to condense into a single chip LSI (Large Scale Integrated circuit), and the rapid development of new devices, e.g. SAW (Surface Acoustic Wave) filter and CCD (Charge Coupled Device) [76], had brought the easy

and quick signal acquisition technique for the SS receiver.

The above advancement in semiconductor technology produced the possible result of application of SS systems into the civilian commercial usage. In addition, the XIV-th CCIR Plenary Assembly in 1978 stated above also accelerated the research in the field of SS communication systems.

Proposed application area of SS systems at present are as follows [75,79].

- (i) Land mobile communication systems. Included are the cellular mobile telephone service by virtue of anti-multipath characteristics [20,24], personal radio communication systems attached great importance to the LPI quality, and so on.
- (ii) Satellite communication systems, such as small scale random accessing between ground stations and between aircraft, ship, and automobiles. Also the deep space communication is presently performed by SS technique.
- (iii) In-house office communication systems [80]. Codeless telephone systems and portable office data communication links are classified into this category.
- (iv) Special application. Emergency distress communication, deep space satellite ranging, navigation (GPS, Global Positioning System) [42,43], and standard time coordinate delivery etc. are under planning.

Incidentally when an application area of SS technique on mass communication field is taken into proportion, a thoughtful consideration to the efficient utilization of the finite radio frequency spectrum resource as well as not to disturb the public order of conventional radio systems is highly requested.

In this point of view, multifarious efforts on SS modulation technique have been made especially for SSMA to realize the efficient utilization of spectrum resource. These studies are divided broadly into two categories; one is to seek for a superior secondary modulation scheme including the development of better pseudo-noise (PN) sequences [27,29,50] and better hopping-patterns [53], the other is to look out for a preferred combination of primary and secondary modulation scheme. The notable works classified to the former category are summarized in [51].

Concerning to the latter category, lots of works combined with FH have been presented [19-23]. On the contrary, a few works have been published for the combination with DS [34,36,37]. Although there exists some slight difference between FH and DS in their suitable application area, DS is basically considered to be easily implementable than FH because of the difficulty of constructing an accurate wideband frequency hopping synthesizers [7].

Consequently, the chapter 3 of this thesis is wholly devoted to be presenting a good combination of primary modulation scheme with DS digital communication systems in order to realize the efficient utilization of spectrum resource. The concrete primary modulation scheme proposed here is the M-ary Frequency Shift Keying (MFSK) [13].

This idea of applying MFSK as primary modulation is originally presented by Yasuda et al. [36]. They have shown that the number of co-channel users may be greatly increased by using MFSK with long block length. This is a trailblazing work in respect that the demodulation gain of a primarily modulated signal as well as the processing gain of a secondary modulation is utilized for the rejection of undesired co-channel signals.

This thesis proposes another type of DS/SSMA system which employs MFSK as primary modulation. The unique point is to modify the MFSK modulator/demodulator to alter the MFSK modulated symbol alignment such that the carrier frequency of every symbols are apart from each other by far. To distinguish the modified MFSK from the conventional MFSK, thus, modified MFSK is called as "unpacked-MFSK", where the conventional one is named as "packed-MFSK" in this thesis [39-41].

The above modification is essentially intended to flatten the frequency spectrum distribution of the transmitted signal over the allocated bandwidth to the SSMA systems. Similar survey is seen on analog DS systems [37]. Takada et al. [38] have theoretically shown that for DS/SSMA systems, the best frequency utilization is achieved by uniformly distributing the signal spectrum over a given bandwidth. In this manner, the result of this thesis presents the practical suboptimum modulation scheme which Takada et al. have anticipated.

Another topics dealt within the latter part of this thesis is a splendid practical application of SSMA to the in-house wireless optical digital communication systems in office, factory, and home environment.

With recent development and progress of office information network, various kind of data link equipment are about to be set up in the in-house environment. Under these circumstances, problems on drawing the signal cables of data terminal equipment are lately closed up. In consequence, the portability of in-house data communication terminals is ready to be highly required, and a in-house wireless communication way will be in great demand [54].

A variety of communication systems such as small power radio

system, power line data transmission, and optical wireless data transmission and the like have been recently proposed and coming into practical use to this requirement [54,55,80]. In particular, the optical wireless data transmission system has the following advantages compared with the other systems : (i) it is not placed under control by radio regulations, (ii) it does not interfere with the adjacent network because the transmitted signal is easily isolated by the object such as the wall of the room, and (iii) it can be used under the heavy electromagnetic interference [12].

Conventional studies [59-63] and practices on optical wireless systems, however, are limited basically to a point-to-point data transmission, and the multiple access system in which more than one data transmissions are made simultaneously in the same communication link has not yet been studied [67-70].

In order to realize the multiple access system in the optical wireless channel, FDMA (Frequency Division Multiple Access), TDMA (Time Division Multiple Access), and SSMA might be utilized as well as in the case of the conventional radio frequency multiple access systems [6].

Among them, SSMA has the outstanding advantages stated above including (i) nearly complete random access without any network control, (ii) LPI character desirable for personal data communication, and (iii) preferable property so called "graceful degradation" on SSMA.

In addition to these advantages, the serious disadvantage in the radio frequency SSMA that the spectrum efficiency is very low [79] is not important on optical communication systems. Thus the optical wireless data link system using spread spectrum method is one of the ideal technique of in-house multiple access

communications.

However there has been few studies on optical SS systems [63]. Especially, no analysis has been made on receiving characteristics of the optical SSMA.

Chapter 4 first describes the properties of in-house optical wireless channel, and discusses the modulation technique which matches these properties. Then, based on the discussion, a modulation technique using baseband direct sequence followed by a binary pulse position modulation (PPM) is proposed, and the theoretical error rate expressions based on the Gaussian noise approximation of the co-channel users' interference and of the shot noise are presented.

## CHAPTER 2

### MODEL FOR DIRECT SEQUENCE SPREAD SPECTRUM DIGITAL COMMUNICATION SYSTEMS

#### 2.1 Introduction

This chapter introduces the overall model for direct sequence (DS) spread spectrum (SS) digital communication systems. Described are the fundamental construction of the transmitter and the receiver, and the mathematical expressions of modulated signal and of multiple accessing. Also the principle and the construction of the synchronization system for the DS/SS receiver is briefly shown. Described model in this chapter is a basis of the DS/SS systems detailed in the following chapters.

#### 2.2 Fundamental System Model

The overall model for DS/SS digital communication systems is depicted in Fig. 2.1. As explained in the previous chapter, the modulation process of a SS transmitter is divided into two parts, namely the primary modulation and the secondary modulation. The primarily modulated output  $u(t)$  of typical DS/SS transmitter in which binary phase shift keying (BPSK) [8] is utilized as



primary modulation is described as

$$u(t) = \sqrt{2S_0} \cdot \sin[\omega_0 \cdot t + a^{(n)} \cdot \frac{\pi}{2}] \quad 0 \leq t < T_b \quad (2.1)$$

where

$T_b$  : Symbol duration of a transmitted data.

$a^{(n)}$  :  $n$ -th transmitted data symbol. =  $\pm 1$

$S_0$  : Averaged signal power.

$\omega_0$  : Carrier frequency.

$u(t)$  is subsequently secondarily modulated to a widespread signal by multiplying a "Pseudo-Noise (PN)" sequence  $b(t)$ . This sequence is a run of binary digits which takes plus or minus one pseudo-randomly. Furthermore this sequence is independent of the binary digits to be transmitted. The production rate of this pseudo-noise digits is usually much faster than the rate of the symbol digits to be transmitted, and many pseudo-noise digits are assigned to one transmitted symbol. The transmitted signal  $s(t)$  is then expressed as

$$s(t) = \sqrt{2S_0} \cdot \sin[\omega_0 \cdot t + b(t) \cdot a^{(n)} \cdot \frac{\pi}{2}] \quad (2.2)$$

$$b(t) = b_k, \quad k \cdot T_c \leq t < (k+1) \cdot T_c, \quad k = \text{integer}$$

where

$T_c$  : Chip duration of a pseudo-noise digit.  $T_c \ll T_b$

$b_k$  : Each pseudo-noise digit.

The ratio of  $T_b$  to  $T_c$  is often called as "the processing gain" of the system [7]. The term "pseudo-random" means that it is

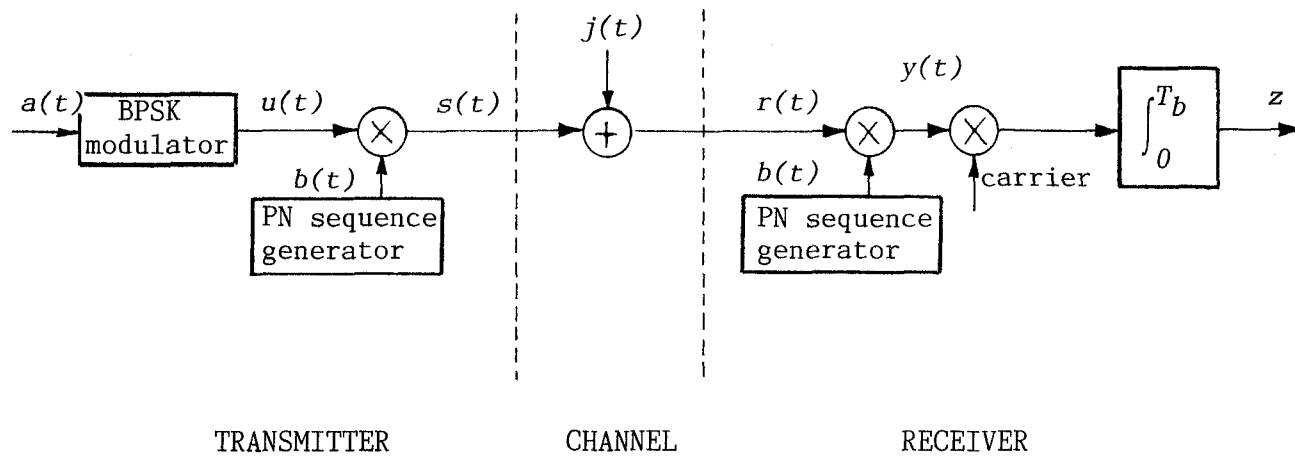


Fig. 2.1 DS/SS digital communication system

expected for the receiver to be able to generate the exactly same sequence as the transmitter used for the secondary modulation. In the strict sense of the word, this noise-like sequence is not random. The well-known PN sequences commonly used are Maximal length codes, Gold codes and Kasami codes [51].

Under the simplest channel model,  $s(t)$  with a certain propagation delay  $\tau$  arrives at the receiver along with additive noise  $j(t)$ . The receiver input  $r(t)$  is then

$$r(t)=s(t-\tau)+j(t) \quad (2.3)$$

At first,  $r(t)$  is despread by again multiplying the PN sequence  $b(t)$  which is exactly same as what is used for spreading at the transmitter except for the effect of the propagation delay  $\tau$ . Let  $\tau=0$  for the convenience, then the despread signal is written as

$$\begin{aligned} y(t) &= r(t) \cdot b(t-\tau) \\ &= \sqrt{2S_0} \cdot \sin[\omega_0(t-\tau) + a^{(n)} \cdot \frac{\pi}{2}] + j(t) \cdot b(t-\tau) \end{aligned} \quad (2.4)$$

where the relation  $b(t)^2=1$  is applied. The despread signal is subsequently demodulated by a conventional coherent demodulator for BPSK. Finally, the demodulator output  $z$  is given by

$$z = \int_0^{T_b} r(t) \cdot b(t) \cdot \sqrt{2/T_b} \cdot \cos \omega_0 t \, dt \quad (2.5)$$

### 2.3 Multiple Accessing

A model for the multiple accessing on DS/SS systems is shown in Fig. 2.2. In this model, asynchronous multiple accessing is assumed, where the signal propagation delay  $\tau$  of each users' signal  $s(t)$  is assumed to be random. This is also referred to as the "random accessing (SSRA)".

Provided that the channel is linear, the input  $r(t)$  to a certain receiver is expressed as the sum of (2.3) for each users' signal.

$$\begin{aligned}
 r(t) &= \sum_{j=0}^N s_j(t-\tau_j) \\
 &= \sum_{j=0}^N \sqrt{2S_j} \cdot \sin[\omega_0(t-\tau_j) + b_j(t-\tau_j) \cdot a_j^{(n)} \cdot \frac{\pi}{2}]
 \end{aligned} \tag{2.6}$$

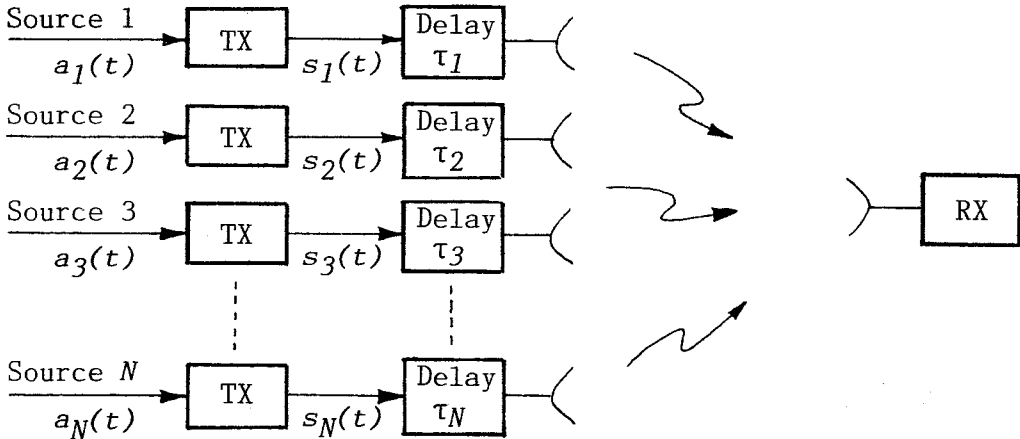


Fig. 2.2 Multiple Accessing

where  $N$  denotes the number of co-channel users. The suffix  $j$  is added on each signal for distinguishing each data, power, PN sequence, and the propagation delay. In order to differentiate a desired signal from the other, every user must use different PN sequence with low cross-correlation coefficient. Now if the receiver is designed to receive the  $i$ -th transmitted signal and assuming that  $\tau_i=0$  without loss of generality, the output  $z$  of the demodulator for the primary modulation is

$$\begin{aligned}
 z &= \sqrt{2S_i} \cdot a_i^{(n)} \cdot b_i^2(t) + \sum_{\substack{j=0 \\ j \neq i}}^N \sqrt{2S_j} \cdot \int a_j^{(n)} \cdot b_j(t-\tau_j) \cdot b_i(t) dt \\
 &= \sqrt{2S_i} \cdot a_i^{(n)} + \sum_{\substack{j=0 \\ j \neq i}}^N \sqrt{2S_j} \cdot a_j^{(n)} \cdot R_{ij}(\tau_j)
 \end{aligned} \tag{2.7}$$

where  $R_{ij}(\tau)$  denotes the cross-correlation function between PN sequences of  $i$ -th and  $j$ -th signals. It is evident, from (2.7),  $R_{ij}(\tau)$  directly relates to the amount of interference due to the other users. In practice, quasi-orthogonal PN sequence such as Gold codes and Kasami codes are applied for DS/SSMA.

Unfortunately, however, the trouble is that the fact of asynchronous multiple accessing prevents the exact evaluation of correlation coefficient between the desired signal and the co-channel users' signal. In consequence, the theoretical error rate expressions for DS/SSMA had been extremely complicated in terms of the partial cross-correlation function of  $R_{ij}(\tau)$  [31].

In the meanwhile, Weber et al. [33] have shown the counter convenient expression of the co-channel users' interference in the spectral notation. According to their results, the spectrum

of the sum of despread co-channel users' signal is

$$\Phi(f) = \frac{T_C^{-1} \cdot S_0}{T_C^2 + \pi^2(f-f_C)^2} \quad (2.8)$$

## 2.4 Synchronization

As mentioned above, any spread spectrum signal receiver must prepare the perfectly same PN sequence which the transmitter used for spreading, including the phase of the PN sequence. But since the propagation delay  $\tau$  is unpredictable in general, the phase of the PN sequence to generate must be estimated from the received signal  $r(t)$  by the synchronization system at the receiver.

The synchronization system is roughly divided into two parts, namely the acquisition subsystem and the tracking subsystem. The roll of these subsystems are as follows.

- (i) Acquisition subsystem : this subsystem estimates the propagation delay  $\tau$  within the accuracy of  $\pm T_C$ . Required feature is to estimate it as soon as possible when the desired signal is present, and not to respond to the received waveform when the desired signal is not present. Typical construction of this subsystem is depicted in Fig. 2.3 [8].
- (ii) Tracking subsystem : this subsystem runs after the acquisition is achieved, holding the estimation error of the propagation delay within  $\pm T_C$  for a long time. The construction of well-known tracking subsystem, named DLL

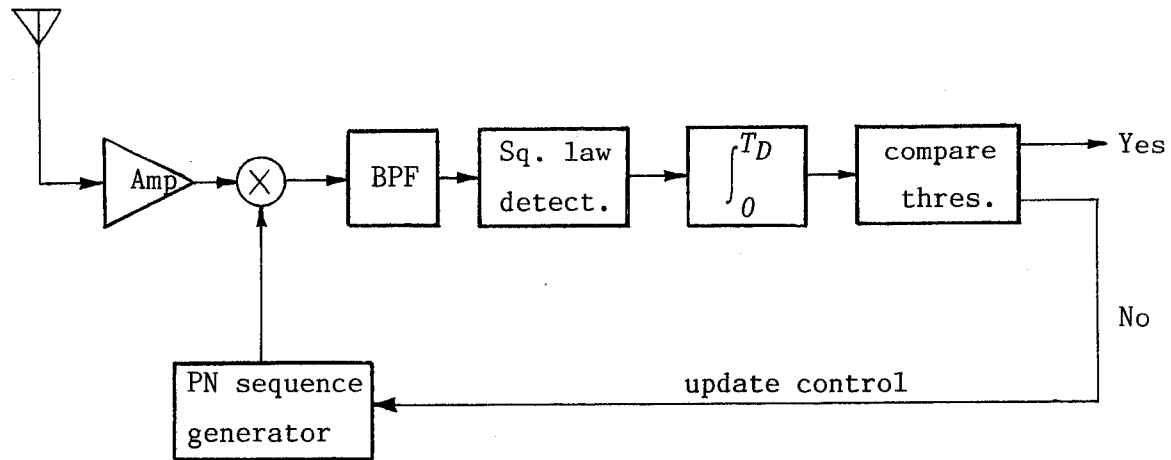


Fig. 2.3 Typical acquisition subsystem for DS/SS

(Delay Locked Loop) for the baseband DS signal is shown in Fig. 2.4 [8].

A principle of both two subsystems is to ingeniously utilize the pointed peak of the auto-correlation function of the PN sequence around  $\tau=0$ . For example, the auto-correlation function  $R_{PN}(\tau)$  for the maximal length code is described as [9]

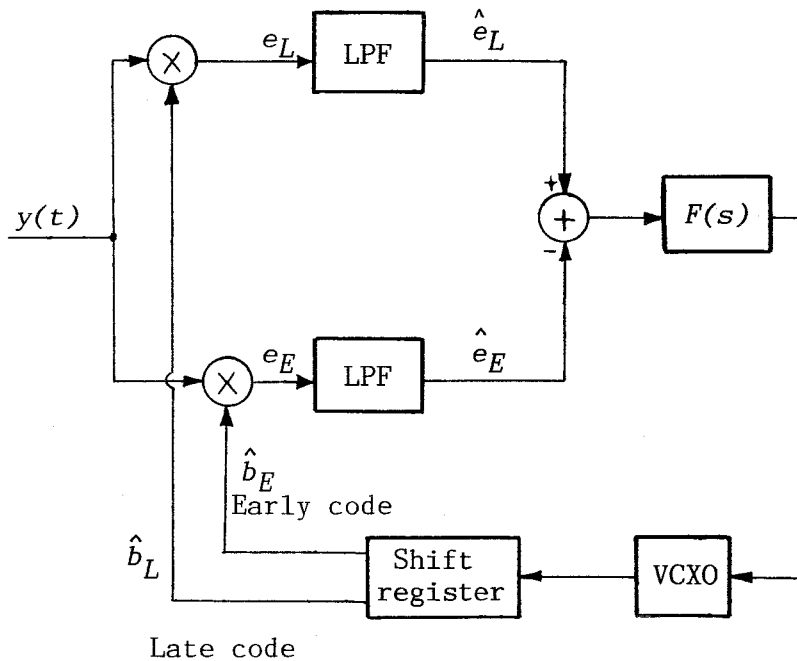


Fig. 2.4 Baseband delay locked loop



$$R_{PN}(\tau) = -\frac{1}{q} + \frac{q+1}{q} \cdot \Lambda(\tau) * \sum_{l=-\infty}^{\infty} \delta(\tau + q \cdot l \cdot T_c) \quad (2.9)$$

$$\Lambda(\tau) = \begin{cases} 1 - \frac{|\tau|}{T_c} & , -T_c \leq \tau < T_c \\ 0 & , \text{elsewhere} \end{cases} \quad (2.10)$$

where \* denotes convolution, and  $q$  denotes the spreading ratio ( $q = T_b/T_c$ ). Fig. 2.5 depicts the shape of  $R_{PN}(\tau)$ .

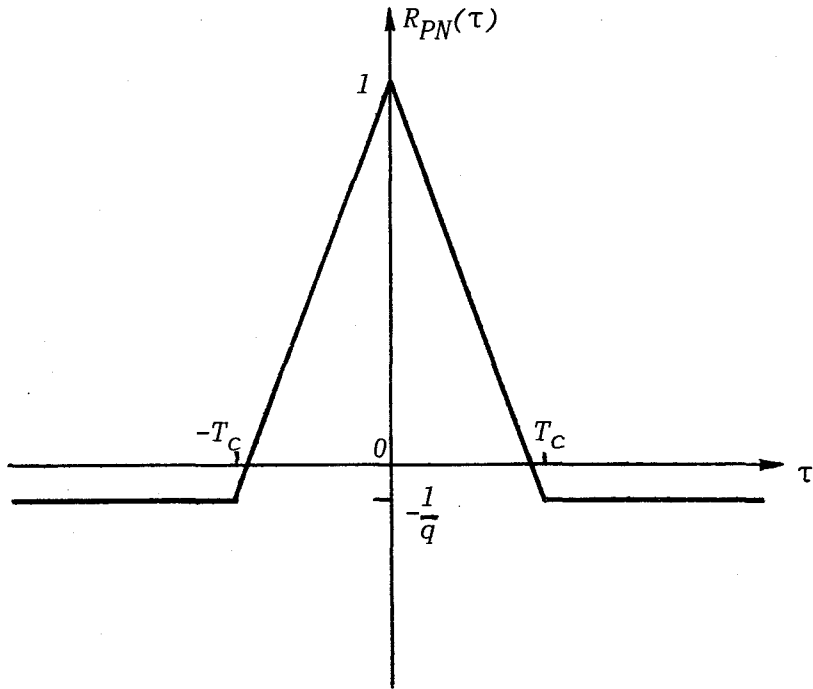


Fig. 2.5 Auto-correlation function  $R_{PN}(\tau)$

## 2.5 Concluding Remarks

A fundamental model for direct sequence spread spectrum digital communication systems is introduced. Presented are the mathematical expressions for the modulation, the demodulation, and the multiple accessing, and the roll of the synchronization systems.

# CHAPTER 3

## IMPROVEMENT OF MULTIPLE ACCESS PERFORMANCE FOR DIRECT SEQUENCE SPREAD SPECTRUM SYSTEMS

### 3.1 Introduction

This chapter is devoted to present a DS/SS digital communication system with good multiple access capability. The term of multiple access capability means how many simultaneous users can access within a given bandwidth and a specified bit error rate. As introduced in chapter 1, the improvement of the multiple access capability is one of the most important problem to be worked out for DS/SSMA.

An effective solution to this problem is to introduce an error correcting code for the channel coding. For example, constraint length = 7, rate = 1/2 convolutional coding and Viterbi decoding with soft decision is shown to improve the multiple access capability several times compared with the uncoded system [35]. But since coding is of no value during acquisition, the acquisition time may be expected to be substantially larger.

Another solution to this problem is employing M-ary FSK (MFSK) instead of binary PSK (BPSK) as primary modulation [36].

The fundamental concept of this system is to positively utilize so called demodulation gain of MFSK as well as the processing gain on despreading in order to reject the co-channel interference due to the simultaneous users. However, the power spectra of these modulated signals i.e. BPSK/DS and MFSK/DS, have the envelope of  $\text{sinc}^2(f)^\dagger$ , therefore the transmitted energy of such signaling is relatively concentrated to its center frequency.

Takada et al. [38] have shown that for BPSK/DS systems, multiple access capability is maximized by flattening the spectrum of transmitted signal over the given bandwidth. The practical method to do so have shown by Sugiyama et al. [37] for the analog DS systems.

In this chapter, a practical method to flatten the spectrum for the digital DS systems using MFSK as primary modulation is presented. Proposed method is simply to change the MFSK symbol allocation over the frequency axis, and it is of no need to modify the conventional MFSK/DS transmitter/receiver constructions.

### 3.2 Use of M-ary Frequency Shift Keying and Its Modification

MFSK, employed as primary modulation scheme in this chapter, is one of the M-ary orthogonal digital modulation technique. Under the white Gaussian noise environment, the larger the number of symbols of MFSK is, the less signal to noise ratio (SNR) is needed to achieve the specified bit error rate [13].

In general, the output of the MFSK modulator which modulates

---

$^\dagger \text{sinc}(f) = \sin(f)/f$

$k$  bits digital data into one FSK symbol is expressed as

$$x(t) = \sqrt{2S_0} \cdot v(t) \cdot \cos(2\pi f_i t + \psi_i), \quad i=1,2,3,\dots,M \quad (3.1)$$

$$v(t) = \begin{cases} 1, & 0 \leq t < k \cdot T_b \\ 0, & \text{elsewhere} \end{cases} \quad (3.2)$$

where  $M$  denotes the total number of the MFSK symbols,  $M=2^k$ ,  $S_0$  is signal power,  $T_b$  is duration of one binary digit to be transmitted,  $f_i$  ( $i=1,2,3,\dots,M$ ) is carrier frequency of each symbol,  $\psi_i$  is initial phase of the symbol.  $k$ -bit digital data is thus modulated onto one of the  $M$  carriers.  $\psi_i$  is assumed to be unknown in this model, hence noncoherent detection at the receiver is considered here. As previously noted, the longer  $k$  becomes, the better the error rate performance of MFSK under the same SNR is achieved, while increasing  $k$  brings tremendous increment of  $M$  (because  $M=2^k$ ) and difficulty of receiver construction.

Carrier frequency of each symbol i.e.  $f_i$  ( $i=1,2,\dots,M$ ) is usually aligned so close to each other keeping their mutual orthogonality. The reason for this alignment is to minimize the overall bandwidth occupied by the MFSK signal. In this case, each space  $\Delta f$  between adjacent symbols is [13]

$$\Delta f = \frac{1}{k \cdot T_b} \quad (3.3)$$

The conventional MFSK modulation scheme obeying the rule (3.3) is called as "packed-MFSK" in this thesis. On the other hand, a modification to the rule (3.3) that each space between adjacent symbols be much wider than (3.3) is presented. This

scheme is named as "unpacked-MFSK". As shown in the following sections, unpacked-MFSK/DS systems have better multiple access capability than the conventional packed-MFSK/DS systems.

### 3.3 Spectral Analyses of Each Systems

#### 3.3.1 Assumption

Before the comparison of two systems, let us summarize the assumptions.

- (i) The number of co-channel users are  $N+1$  including the desired signal. All of those signals have the same statistical characteristics.
- (ii) The bandwidth assigned to this SSMA system is  $W_{SS}$ , and the channel has a white Gaussian thermal noise source with spectral density  $n_0/2$ .
- (iii) Chip duration  $T_c$  is much shorter than the data symbol duration  $T_b$ ,  $T_c \ll T_b$ .
- (iv) Every user transmits  $M$  MFSK symbols equally likely.
- (v) Convenient spectral notation of equation (2.8) derived by Weber et al. [33] is used for the theoretical error rate calculation.
- (vi) Synchronization to the desired signal is perfect.

#### 3.3.2 Packed-MFSK/DS

For convenience of explanation, let  $M=2$  ( $k=1$ ). Symbol

separation  $\Delta f$  is  $1/T_b$  according to (3.3), If we choose

$$\frac{1}{T_c} = \frac{1}{2} \cdot W_{SS} \quad (3.4)$$

the bandwidth of the modulated SS signal will be roughly confined to  $W_{SS}$  (see Fig. 3.1-a ). On the despreading process at the receiver, undesired co-channel users' signal is changed approximately into the widespread Gaussian noise with the following spectral density  $\Phi_D(f)$  [33].

$$\Phi_D(f) = \frac{T_c^{-1} \cdot S_0 \cdot N}{T_c^{-2} + \pi^2 (f - f_c)^2} \quad (3.5)$$

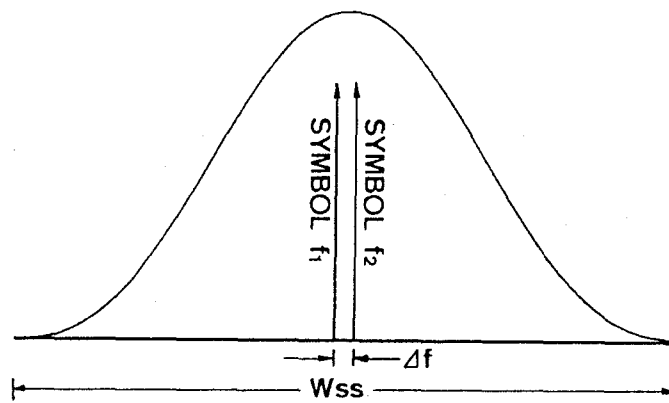
where  $f_c$  is the central frequency of the SS signal. On demodulating the recovered MFSK signal, the noise level only in close vicinity to their symbol carrier frequencies, i.e.  $f_i$  ( $i=1,2$ ), is dominant. Since  $f_i \approx f_c$ , by adding two-sided white Gaussian noise of  $n_0/2$ ,

$$\Phi_D(f) \big|_{f \approx f_c} = \frac{S_0 N}{T_c^{-1}} + n_0 = \frac{2S_0 \cdot N}{W_{SS}} + n_0 \quad (3.6)$$

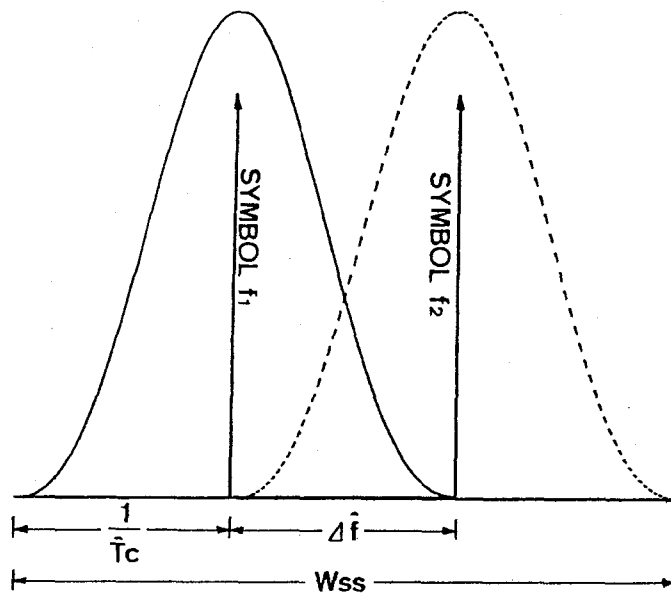
will be the noise level on demodulating MFSK.

### 3.3.3 Unpacked-MFSK/DS

We choose the symbol separation  $\Delta f$  to be much wider than (3.3). In this section, for example, let



(a) Packed -MFSK / DS



(b) Unpacked -MFSK / DS

Fig. 3.1 Power spectra of packed- and unpacked-MFSK/DS



$$\Delta \hat{f} = \frac{1}{3} W_{SS} \quad (3.7)$$

(Sign  $\wedge$  is used to distinguish unpacked- case from packed-MFSK.)  
The chip duration  $\hat{T}_C$  of the PN sequence is then necessary to be

$$\frac{1}{\hat{T}_C} = \frac{1}{3} W_{SS} \quad (3.8)$$

in order to keep  $W_{SS}$  the same as the packed-MFSK/DS case. From (3.4) and (3.8), it is found that the spreading ratio in unpacked-MFSK/DS is forced to be decreased by a factor of 2/3.

On the contrary, according to the assumption (iv) discussed above, it is highly expected that a half of the  $N$  co-channel users transmit symbol  $f_1$ , and another half transmit symbol  $f_2$  at an instance. Thus, noise spectral density  $\hat{\Phi}_D(f)$  around  $f_i$  ( $i=1,2$ ) is

$$\begin{aligned} \hat{\Phi}_D(f) \big|_{f \approx f_i} &= \frac{N \cdot S_0}{2 \cdot \hat{T}_C^{-1}} + \frac{N \cdot S_0 \hat{T}_C^{-1}}{2 \cdot \hat{T}_C^{-2} + \pi^2 \Delta f^2} + n_0 \\ &= \frac{2S_0 \cdot N}{W_{SS}} \cdot \frac{3}{4} \left\{ 1 + \frac{1}{1 + \pi^2} \right\} + n_0 \end{aligned} \quad (3.9)$$

Since

$$\frac{3}{4} \left\{ 1 + \frac{1}{1 + \pi^2} \right\} \approx 0.819 \quad (3.10)$$

it is obviously concluded from (3.6), (3.9), and (3.10) that the noise density on unpacked-MFSK/DS due to the co-channel users' interference is less than that on packed-MFSK/DS. Although

unpacked-MFSK/DS has less spreading ratio than packed-MFSK/DS, the total receiving interference due to the co-channel users is reduced on unpacked-MFSK/DS. This is the effect of spectral flattening of transmitted signal.

In the practical application, when PN sequence such as Gold code [7] or Kasami code [51] is used,  $\hat{T}_C$  must be one of the following discrete value in order to assign a period of PN sequence to one MFSK symbol,

$$\hat{T}_C = \frac{k \cdot T_b}{2^{L-1}}, \quad L = \text{integer} \quad (3.11)$$

where  $L$  denotes the number of shift register stages of PN sequence generator. Moreover, the symbol spacing  $\Delta\hat{f}$ , selected as  $W_{SS}/3$  in the previous example, is arbitrary as long as the relation

$$\Delta\hat{f} + 2\hat{T}_C^{-1} = W_{SS} \quad (3.12)$$

is satisfied. Accordingly, there exists some freedom to choose  $L$ . But since the multiple access capability strongly depends on  $L$ , an optimization on  $L$  must be necessary.

Above discussion is restricted to the special case of  $M=2$ . Here we extend these expressions to arbitrary  $M$ . Then (3.12) will be

$$(M-1) \cdot \Delta\hat{f} + 2\hat{T}_C^{-1} = W_{SS} \quad (3.13)$$

and (3.9) will be

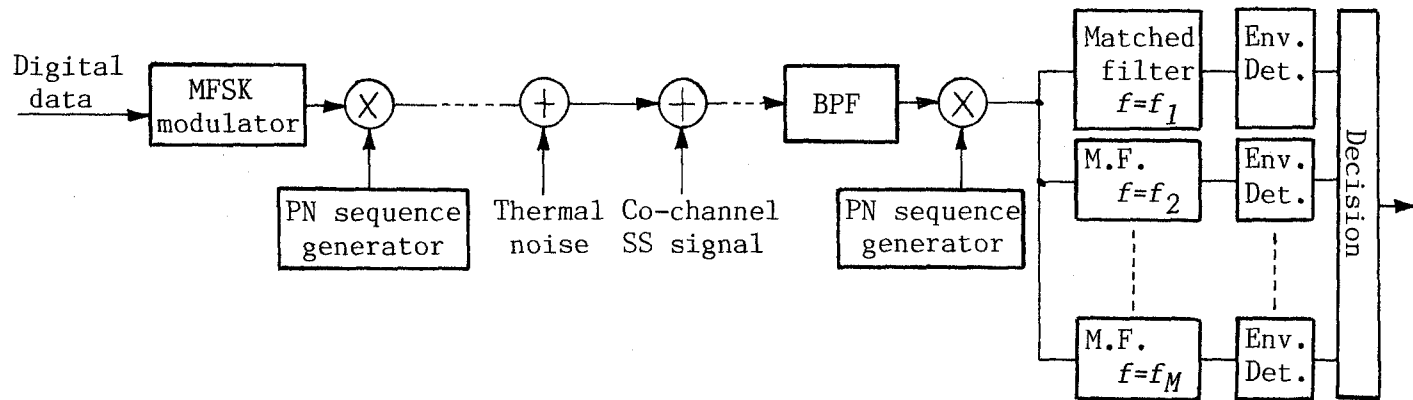


Fig 3.2 Noncoherent MFSK/DS communication system

$$\hat{\Phi}_D(f) \big|_{f \approx f_i} = \sum_{j=1}^M \frac{S_0 \hat{T}_C^{-1}}{\hat{T}_C^{-2} + \pi^2 (i-j)^2 \cdot \Delta f^2} \cdot \frac{N}{M} + n_0 \quad (3.14)$$

where each adjacent symbol of MFSK is assumed to be equally spaced, and  $N \gg M$ .

### 3.3.4 Error Rate Expressions

Fig. 3.2 shows the noncoherent MFSK/DS SSMA systems. Symbol error rate  $P_{SE}$  in terms of  $\hat{\Phi}_D(f)$  in (3.14) is derived as follows.

$$P_{SE} = \frac{1}{M} \sum_{p=1}^M \left\{ 1 - \prod_{\substack{q=1 \\ q \neq p}}^M \left\{ 1 - \frac{\hat{\Phi}_D(f_q)}{\hat{\Phi}_D(f_p) + \hat{\Phi}_D(f_q)} \right\} \right\} \cdot \exp \left\{ \frac{-S_0 \cdot T_b \cdot k}{\hat{\Phi}_D(f_p) + \hat{\Phi}_D(f_q)} \right\} \quad (3.15)$$

(see appendix A for the derivation.)

In unpacked-MFSK/DS systems in general, the output noise power of  $M$  matched filters are different from each other, and is concentrated around the center frequency of the SS signal. This fact prevents the bit error rate  $P_e$  from being expressed with a simple formula in terms of  $P_{SE}$ . Nevertheless, it is expected to derive  $P_e$  in order to compare it with other systems. Thus in the subsequent section, for the convenience, bit error rate  $P_e$  is calculated using a universal expression of  $P_e$  and  $P_{SE}$  for the general M-ary orthogonal signals written below [4].

$$P_e = \frac{1}{2} \frac{M}{M-1} P_{SE} \quad (3.16)$$

### 3.4 Numerical Results and Discussions

Fig. 3.3 shows the numerical result of allowable number of co-channel users for a fixed  $W_{SS}$  and a required  $P_e$  in terms of received signal to thermal noise ratio. Both packed- and unpacked-MFSK/DS as well as the conventional PSK/DS are compared. In PSK/DS, bit error rate expression is given as [7]

$$P_e = \frac{1}{2} \operatorname{erfc} \sqrt{\frac{T_b \cdot S_0}{N \cdot T_C \cdot S_0 + n_0}} \quad (3.17)$$

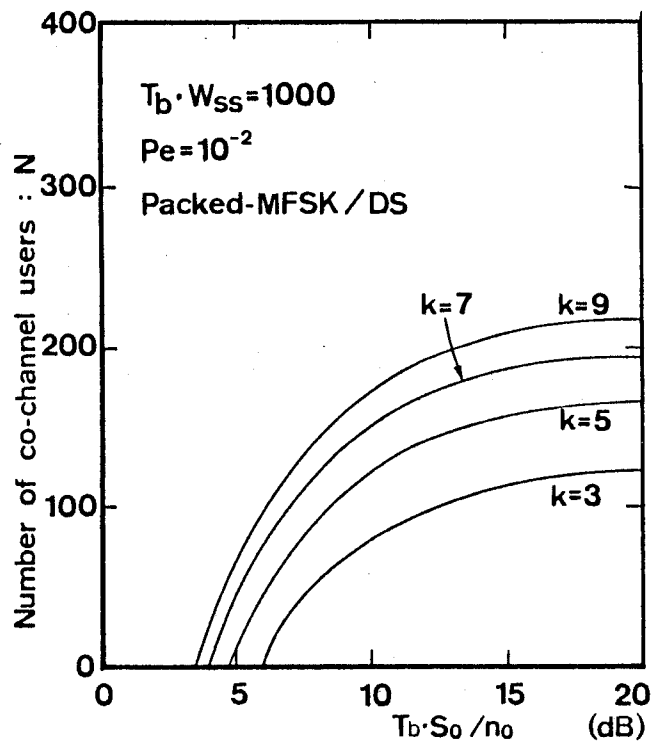
where

$$T_C^{-1} = \frac{W_{SS}}{2} \quad (3.18)$$

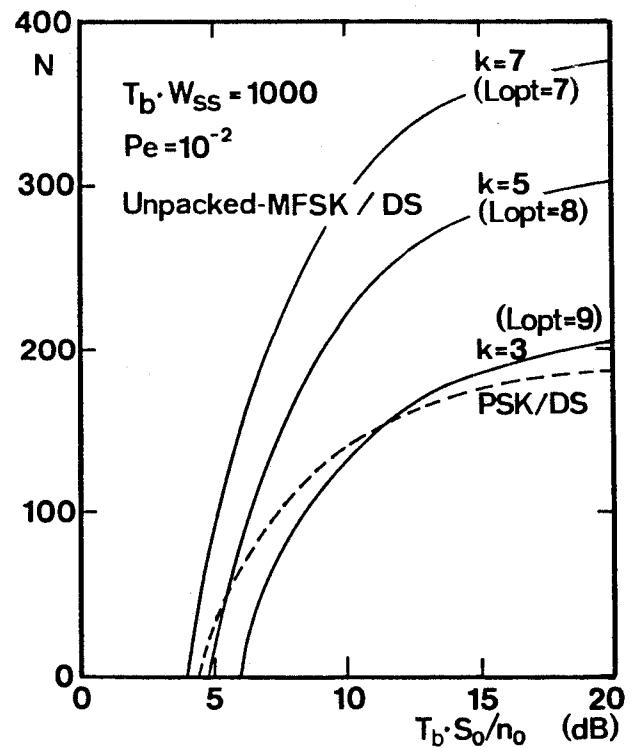
As previously noted, since the number of the shift register stages  $L$  may be set arbitrarily in unpacked-MFSK/DS,  $L$  is optimized to  $L_{opt}$  for maximizing the allowable number of co-channel users.

It is readily found the following characteristics from Fig. 3.3.

- (i) When the block length  $k$  is long enough, both MFSK/DS systems have the superior multiple access capability to the conventional PSK/DS. This is due to the fact that MFSK is

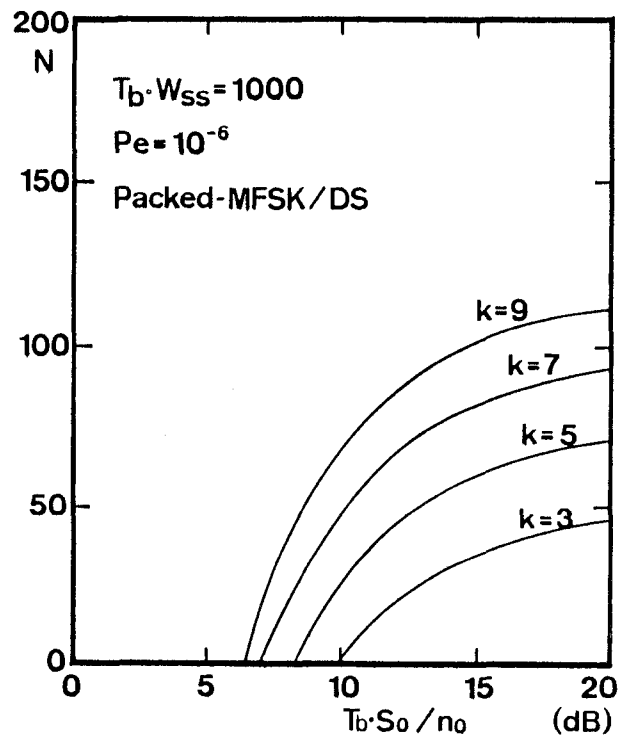


(a)  $P_e = 10^{-2}$ , packed-MFSK/DS

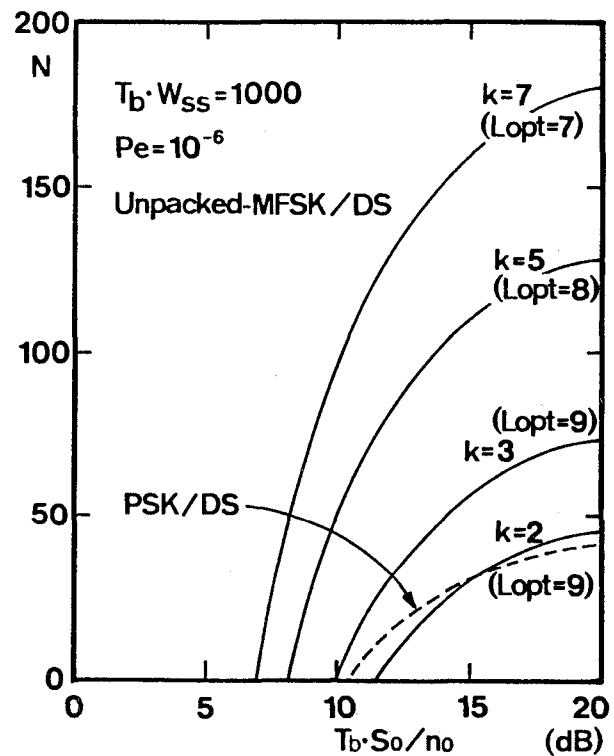


(b)  $P_e = 10^{-2}$ , unpacked-MFSK/DS

Fig. 3.3 Number of available co-channel users



(c)  $P_e = 10^{-6}$ , packed-MFSK/DS



(d)  $P_e = 10^{-6}$ , unpacked-MFSK/DS

Fig. 3.3 Continued

one of the M-ary orthogonal modulation scheme and that it possesses so-called the demodulation gain on receiving.

- (ii) The multiple access capability is increased monotonously as  $k$  increases. Although the number of co-channel users up to  $k=7$  is calculated in Fig. 3.3 for the realizability of the receiver hardware, employing longer  $k$  causes abrupt decrement of processing gain. Hence the multiple access capability will be conversely decreased for longer  $k$  [39].
- (iii) Unpacked MFSK/DS has about twice as large multiple access capability as packed-MFSK/DS of the same  $k$ . Since  $k$  is one of the measure to evaluate the receiver complexity, the receiver complexity will be reduced by employing unpacked-MFSK/DS for a specified number of co-channel users. This is the result of spectral flattening of a transmitted signal, and it is found that a signaling design for reducing the interference to the other users is strongly desired in SSMA.
- (iv) The spreading ratio is forced to be reduced in unpacked-MFSK/DS. This implies that an anti-jam characteristics which any SS systems originally possesses might be sacrificed to some extent.

### 3.5 Concluding Remarks

A digital SSMA system employing MFSK as primary modulation is investigated, and a modification on the symbol alignment of each MFSK symbol is proposed in order to improve the multiple access capability. Proposed modification brings the spectrum distribution of the transmitted signal flat, and improves the multiple access capability up to twice as large as that of the



conventional MFSK/DS systems.

## CHAPTER 4

### IN-HOUSE WIRELESS OPTICAL SPREAD SPECTRUM MULTIPLE ACCESS SYSTEMS

#### 4.1 Introduction

With recent development and progress of office information network the portability of in-house data communication terminals is highly required, and a in-house wireless communication way is in great demand [54] as introduced in chapter 1.

A variety of communication systems such as small power radio system, power line data transmission, and optical wireless data transmission and the like have been recently proposed and coming into practical use to this requirement. In particular, the optical wireless data transmission system has the following advantages :

- (i) it is not placed under control by radio regulations,
- (ii) it does not interfere with the adjacent network because the transmitted signal is easily isolated by the object such as the wall of the room,
- (iii) it can be used under the heavy electromagnetic interference.

Conventional studies and practices on optical wireless system, however, are limited basically to a point-to-point data

transmission, and the multiple access system in which more than one data transmissions are made simultaneously in the same communication link has not yet been studied.

In order to realize the multiple access system in the optical wireless channel, FDMA (Frequency Division Multiple Access), TDMA (Time Division Multiple Access), and SSMA (Spread Spectrum Multiple Access) might be utilized as well as in the case of the conventional radio frequency multiple access systems.

Among them, SSMA has the following outstanding advantages :

- (i) nearly complete random access can be achieved without any network control,
- (ii) new data terminals are easily incorporated into the network without any change of the network,
- (iii) it has a preferable characteristic so called "graceful degradation" on the overload performance,
- (iv) the data terminals with different data transmission rate can be accommodated in the network.

In addition to these advantages, the serious disadvantage in the radio frequency SSMA that the spectrum efficiency is very low [35] is not important on optical communication systems. Thus the optical wireless data link system using spread spectrum (SS) is one of the ideal technique of in-house multiple access communications.

However there has been few studies on optical SS system. Especially, no analysis has been made on receiving characteristics of the optical SSMA.

This chapter first describes the properties of in-house optical wireless channel, and discusses the modulation technique which matches these properties. Then, based on the discussion, a

modulation technique using baseband direct sequence (DS) signal followed by a binary pulse position modulation (PPM) is proposed, and the theoretical error rate expressions based on the Gaussian noise approximation of the co-channel users' interference and of the shot noise are presented.

## 4.2 Selection of Modulation Technique

### 4.2.1 Restrictions on In-House Wireless Systems

There arises several restrictions on an in-house wireless optical channel unlike a optical fiber channel. The modulation technique must be chosen taking account of on the following restrictions.

- (i) As an E/O (electric to optic) device, a Light Emitting Diode (LED) rather than a Laser Diode (LD) is preferred from a viewpoint of the security to the human body. However, since the response frequency of LED is at most 200MHz at present [12], modulation speed is limited up to 200Mbits/sec.
- (ii) The desired optical signal, co-channel users' signal, and ambient lights (background radiation) arrive at the O/E (optic to electric) device in the receiver. Since, the intensity of the received signal varies widely with the distance and the route between transmitter and receiver, a modulation system which needs a control of a decision threshold level in accordance with the intensity of the received signal such as on-off keying (OOK) modulation is not suitable for the wireless optical channel.

As the modulation technique of optical carrier, there are two kinds of intensity modulation schemes, namely a direct intensity modulation (D-IM), and a subcarrier intensity modulation (S-IM) [12]. S-IM is superior in that it can eliminate the low-frequency noise of background radiation such as the flicker noise due to a fluorescent lamp [55]. When the practical data transmission rate and the spreading ratio are assumed<sup>†</sup>, the required driving frequency of LED, however, exceeds the value in the restriction (i) by far. Therefore, D-IM in which a baseband SS signal drives the LED directly must be chosen.

A binary pulse-position-modulated (PPM) waveform [58] is assumed as the waveform of driving LED in this chapter. This is also referred to as the Manchester coding [57], which takes non-negative values, and satisfies the restriction (i).

#### 4.2.2 Spreading Technique

The following three types of spreading technique are well-known [7]: Frequency Hopping (FH), Direct Sequence (DS), and Time Hopping (TH). These are characterized as follows.

- (i) FH.... In FH system its subcarrier frequency is changed periodically in accordance with the hopping pattern. Accordingly, under the restriction (i) discussed above, we cannot spread the transmission data spectrum enough. Since

---

<sup>†</sup> 64k bits/sec as data transmission rate, 1023 as spreading ratio are assumed in this chapter. This system has a capacity of dozens of co-channel users.

FH has an advantage on the near/far problem [24] compared with DS, less margin is required on system design.

- (ii) DS.... Using DS technique, we can spread the transmission data spectrum enough if baseband D-IM is employed. Since DS has a disadvantage on the near/far problem, enough margin on channel design must be required. Hardware construction is very easy.
- (iii) TH.... In general, the peak value of emitting power of LED is limited to some value. Thus in TH, averaged emission power will be extremely reduced if the transmission data is spread widely. This obliges us not to spread the data widely. TH has an advantage on near/far problem as well as FH. In addition, the shot noise due to the co-channel users' signal may be reduced in TH.

In this manner, these three types of spreading format have both advantages and disadvantages. Accordingly in this chapter, a binary PPM as the baseband waveform and baseband DS as the spreading technique are assumed for the hardware simplicity.

### 4.3 System Model

#### 4.3.1 Transmitter

Fig. 4.1 shows the block diagram of the transmitter. The transmitted data,  $a(t)=\pm 1$  of bit rate  $1/T_b$  is first multiplied by the pseudo-noise (PN) sequence,  $b(t)=\pm 1$  with chip rate  $1/T_c$ , and is spread on the baseband region. Here  $q=T_b/T_c$  and  $q$  is a spreading ratio which is far larger than 1. Next, this spread

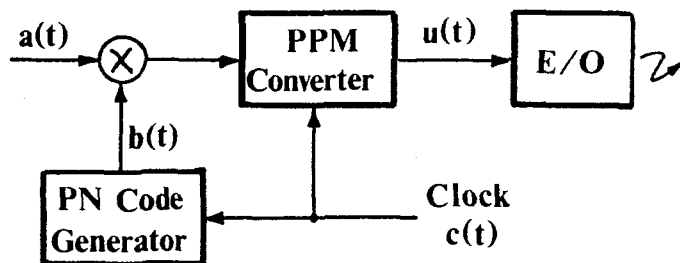


Fig. 4.1 Transmitter diagram

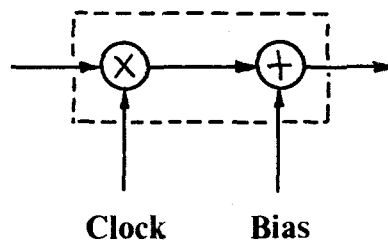


Fig. 4.2 PPM converter

signal is converted to the binary PPM waveform by a clock signal  $c(t)$  (Fig. 4.2). The output of PPM converter,  $u(t)$ , is

$$u(t) = a(t) \cdot b(t) \cdot c(t) + 1 \quad (4.1)$$

where

$$c(t) = \begin{cases} -1, & T_s \leq t < T_s + \frac{1}{2}T_c \\ 1, & T_s + \frac{1}{2}T_c \leq t < T_s + T_c \end{cases} \quad (4.2)$$

and  $T_s$  is the starting time of each chip,

$$T_s \in \{1 \cdot T_c\}, \quad 1=1,2,3,4,\dots$$

Fig. 4.3 shows the generating process of the PPM waveform. Finally,  $u(t)$  is intensity-modulated onto the optical carrier by driving LED directly. Since the multiple access is assumed in this chapter, subscripts which distinguish each users are added in (4.1). Thus, for example, the optical signal intensity  $I_{sj}(t)$  radiated from the  $j$ -th user's LED is

$$\begin{aligned} I_{sj}(t) &= I_{cj} \cdot u_j(t) \\ &= I_{cj} \cdot \{a_j(t) \cdot b_j(t) \cdot c(t) + 1\} \end{aligned} \quad (4.3)$$

where  $I_{cj}$  is averaged signal intensity of  $j$ -th user.



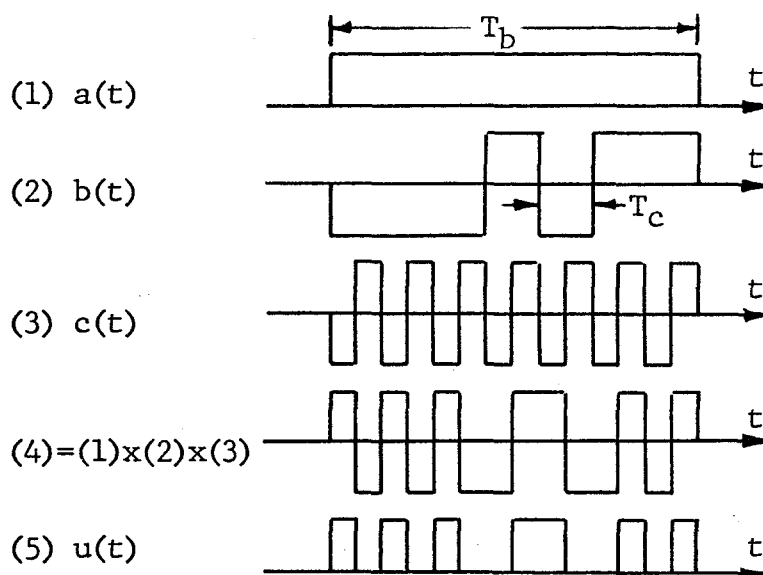


Fig. 4.3 Generation of PPM waveform

#### 4.3.2 Channel

Roughly speaking, there are two types of in-house wireless optical systems ; The one uses the line-of sight path which interconnects a transmitter with a receiver directly, and the other uses the diffusively reflected radiation. Since in the latter system heavy multipath phenomena may be observed, the channel is modeled as a filter with impulse response  $h_p(\tau)$  [55].

$$h_p(\tau) = \begin{cases} \frac{2\tau_0^2}{\tau^3 \sin^2(\text{FOV})} & , \tau_0 \leq \tau < \frac{\tau_0}{\cos(\text{FOV})} \\ 0 & , \text{elsewhere} \end{cases} \quad (4.4)$$

where

$\tau_0$  : minimum transmission delay between transmitter and receiver,

FOV : the field of view of receiver.

Although this kind of system can offer very flexible interconnecting capability, the multipath effect expressed by (4.4) limit the transmission data rate and/or the number of co-channel users.

However, once a channel between a satellite on the ceiling and a terminal-head on the desk is assumed, we need not take the influence of multipath into account, but the propagation loss of free space alone. Hence, the former type of system which is considered to be fundamental is discussed in this chapter.

Another factor to be taken into account is the propagation loss between transmitter and receiver. When the propagation channel is a long atmospheric one, we must pay attention to the loss of absorption, scattering, and scintillation due to the atmosphere as well as the free space propagation loss.

In the in-house wireless optical channel, however, it is almost always enough to consider the free space propagation loss alone as follows [12].

$$I_r = \frac{G_a \cdot A}{4\pi^2 L_p^2} I_t \quad (4.5)$$

where

$I_r$  : received intensity of the optical signal,

$I_t$  : transmitted intensity of the optical signal,

$G_a$  : relative gain of the transmitter's antenna to the isotropic radiated antenna,

$A$  : receiving area of O/E device,

$L_p$  : distance between transmitter and receiver.

$G_a$  is defined by the directivity of the transmitter's antenna. Suppose it has three dimensional directivity of  $\theta$  centered on its maximum radiated direction,

$$G_a = \frac{2}{1 - \cos\theta} \quad (4.6)$$

is derived easily. For example, let

$$L_p = 20m \quad (\text{medium size office})$$

$$\theta = \pi/6 \quad (\text{medium directivity})$$

$$A = 1cm^2 \quad (\text{easily attainable for PIN diode})$$

then

$$I_r \approx 3 \times 10^{-6} I_t$$

It is equivalent to the loss of about 55dB.

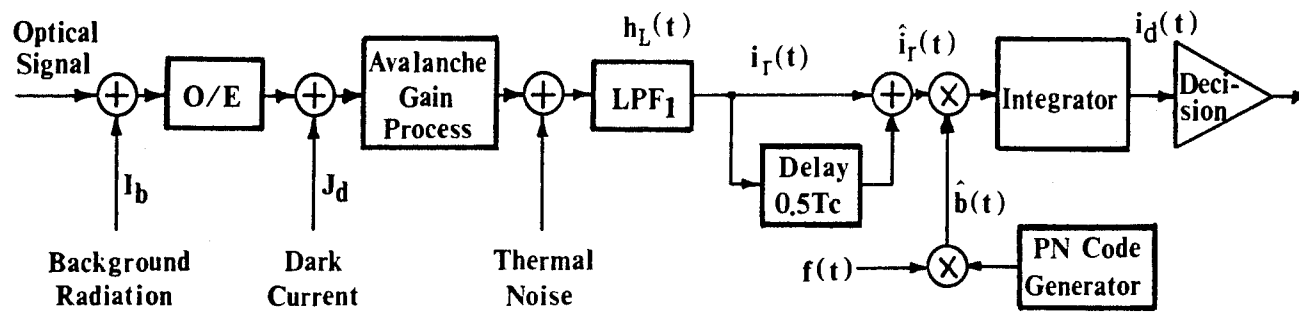


Fig. 4.4 Receiver diagram

### 4.3.3 Receiver

Fig. 4.4 shows the block diagram of the receiver construction. Here an Avalanche Photo Diode (APD) is employed as O/E device for the generalization of error rate expressions. In case a PIN photo diode is used instead of an APD, the following expressions still holds as the particular case of APD gain  $G=1$ .

Because of optical SSMA, several users' signal as well as ambient light arrive at the input of the receiver. Suppose that these signals and ambient light are mutually independent<sup>†</sup>, the averaged output of the  $LPF_1$  is given as follows [56].

$$\langle i_r(t) \rangle = \left[ \frac{\eta}{h\nu} \cdot G \cdot e \left\{ \sum_{j=0}^N I_{cj} \cdot u_j(t - \tau_j) + I_b \right\} + G \cdot J_d \right] * h_L(t) \quad (4.7)$$

where

- $\langle \rangle$  : ensemble average over the randomness of O/E and avalanche process,
- $*$  : convolution,
- $e$  : electron charge,
- $\eta$  : quantum efficiency of APD,
- $G$  : averaged APD gain,
- $h\nu$  : energy of a photon,
- $J_d$  : dark current of APD,
- $h_L(t)$ : impulse response of  $LPF_1$ ,
- $\tau_j$  : propagation delay of the  $j$ -th signal,
- $I_b$  : intensity of background radiation,
- $N$  : number of co-channel users.

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<sup>†</sup> This assumption is reasonable if LEDs are used as E/O.

$LPF_1$  represents the frequency characteristic of O/E process, and generally the cut off frequency is assumed to be extremely higher than the bandwidth of desired signal.

Since  $i_r(t)$  in (4.7) is a non-negative PPM waveform biased with ambient light and dark current, reversion from PPM to bipolar waveform, such as non-return-to-zero (NRZ) or return-to-zero (RZ), is needed before despreading. For the simplicity's sake, reversion to RZ waveform is assumed here.

This is done by first subtracting the delayed version of  $i_r(t)$  from  $i_r(t)$ ,

$$\hat{i}_r(t) = i_r(t) - i_r(t - \frac{1}{2}T_c) \quad (4.8)$$

As shown in (4.10), the PN sequence  $\hat{b}_0(t)$  for despreading the desired signal ( $j=0$ ) is generated by using both the original PN sequence  $b_0(t)$  and the gating signal  $f(t)$ ,

$$f(t) = \frac{1}{2}\{c(t)+1\} \quad (4.9)$$

$$\hat{b}_0(t) = b_0(t) \cdot f(t) \quad (4.10)$$

When the synchronization to the desired signal is maintained perfectly, the received signal is successfully despread by multiplying  $\hat{i}_r(t)$  by  $\hat{b}_0(t)$ , and then integrating over  $T_b$  second. Thus the despread output  $i_d(t)$  in Fig. 4.4 is given by

$$i_d(t) = \frac{1}{T_b} \int_{t-T_b}^t \hat{i}_r(t') \cdot \hat{b}_0(t' - \tau_0) dt' \quad (4.11)$$

Fig. 4.5 depicts the despreading process.

#### 4.4 Error Rate Performance

Theoretical expression for the error rate at the output of the receiver is derived in this section.

$\langle i_r(t) \rangle$ , the ensemble average of  $i_r(t)$ , was given in (4.7), whereas  $i_r(t)$  also contains the shot noise  $n_{rs}(t)$  and the thermal noise  $n_{rt}(t)$ . Therefore,

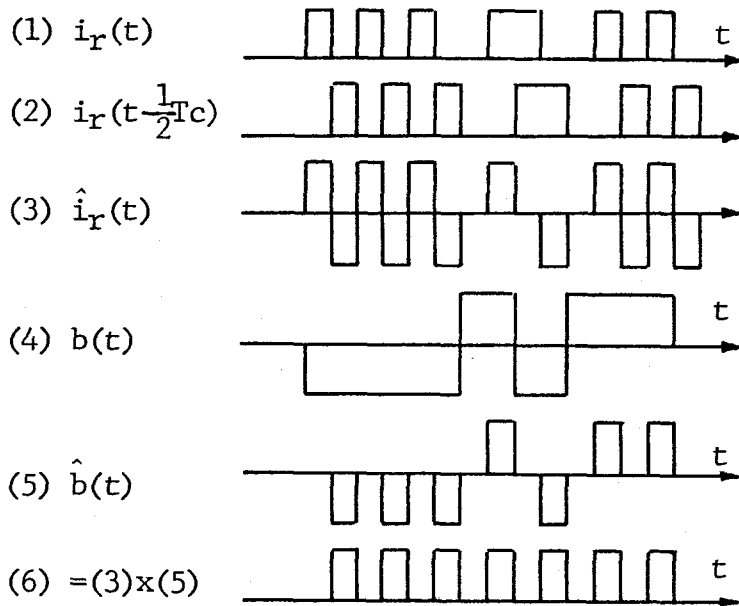


Fig. 4.5 Despreading process

$$i_r(t) = \langle i_r(t) \rangle + n_{rs}(t) + n_{rt}(t) \quad (4.12)$$

It is shown in [56] that  $n_{rs}(t)$  can be modeled as a band-limited white Gaussian noise with variance

$$\text{Var}[n_{rs}(t)] = e^2 G^{2+x} \frac{\eta}{h\nu} I_r(t) * h_L(t)^2 \quad (4.13)$$

where  $x$  is the excess noise factor of APD, and

$$I_r(t) = \sum_{j=0}^N I_{cj} \cdot u_j(t - \tau_j) + I_b \quad (4.14)$$

In (4.14),  $I_b$  contains the influence of the dark current  $J_d$ .

Now let  $H_L(f)$  be the Fourier transform of  $h_L(t)$ , and suppose that  $H_L(f)$  be an ideal rectangular low-pass filter with cutoff frequency  $B$ .

$$H_L(f) = \begin{cases} 1, & |f| \leq B \\ 0, & \text{elsewhere} \end{cases} \quad (4.15)$$

Then (4.7) and (4.13) are simplified as

$$\langle i_r(t) \rangle = \frac{\eta}{h\nu} \cdot G \cdot e \cdot I_r(t) \quad (4.16)$$

$$\text{Var}[n_{rs}(t)] = 2e^2 G^{2+x} \frac{\eta}{h\nu} I_r(t) \cdot B \quad (4.17)$$



Furthermore,  $\hat{i}_r(t)$  is written as follows.

$$\begin{aligned}\hat{i}_r(t) = & \langle i_r(t) \rangle + n_{rs}(t) + n_{rt}(t) - \langle i_r(t - \frac{1}{2}T_c) \rangle \\ & - n_{rs}(t - \frac{1}{2}T_c) - n_{rt}(t - \frac{1}{2}T_c)\end{aligned}\quad (4.18)$$

By using (4.18), (4.10), and letting  $\tau_0=0$  without loss of generality, the integrated component of (4.11),  $\hat{i}_r(t)\hat{b}_0(t)$  will be represented by

$$\begin{aligned}\hat{i}_r(t) \cdot \hat{b}_0(t) = & 2\alpha \cdot f(t) \cdot I_{c0} \cdot a_0(t) \cdot b_0(t)^2 \\ & + 2\alpha f(t) \sum_{j=1}^N \{I_{cj} a_j(t - \tau_j) b_j(t - \tau_j) c(t - \tau_j) b_0(t)\} \\ & + f(t) b_0(t) \{n_{rs}(t) + n_{rt}(t) - n_{rs}(t - \frac{1}{2}T_c) - n_{rt}(t - \frac{1}{2}T_c)\}\end{aligned}\quad (4.19)$$

where

$$\alpha = \frac{\eta}{h\nu} G \cdot e \quad (4.20)$$

and the relation  $c(t - \frac{1}{2}T_c) = -c(t)$  is utilized.

In (4.19), the first term represents the desired signal component, the second term represents the interference due to the signals of co-channel users, and the third term represents the shot noise and the thermal noise. The dc component due to ambient light is eliminated by subtraction  $i_r(t) - i_r(t - \frac{1}{2}T_c)$ .

Finally, (4.19) is integrated over  $T_b$  sec to generate  $i_d(t)$ , and sampled at  $t=T_b$ . As discussed above,  $i_d(t)$  contains four terms, namely the desired signal, the interference due to the

signals of co-channel users, the shot noise, and the thermal noise. Denote these four terms as  $s(t)$ ,  $n_i(t)$ ,  $n_s(t)$ , and  $n_t(t)$ , respectively. Therefore

$$i_d(t) = s(t) + n_i(t) + n_s(t) + n_t(t) \quad (4.21)$$

Here

$$s(t) = \alpha \cdot I_{c0} \cdot a_0(t) \cdot \frac{1}{T_b} \int_{t-T_b}^t 2f(t') \cdot b_0(t')^2 dt' \quad (4.22)$$

Weber et al. [33] have shown that  $n_i(t)$  at  $t=T_b$  can be approximated by the Gaussian random variable with variance

$$\begin{aligned} \text{Var}[n_i(T_b)] &= \sum_{j=1}^N \frac{1}{T_b} \frac{(\alpha I_{cj})^2}{T_c^{-1}} \\ &= \sum_{j=1}^N \frac{(\alpha I_{cj})^2}{q} \end{aligned} \quad (4.23)$$

$n_s(t)$  in (4.21) is written as follows by using  $n_{rs}(t)$ .

$$n_s(t) = \frac{1}{T_b} \int_{t-T_b}^t \{n_{rs}(t') - n_{rs}(t' - \frac{1}{2}T_c)\} f(t') b_0(t') dt' \quad (4.24)$$

Although  $n_{rs}(t)$  is essentially the non-stationary noise which depends on the intensity of received optical signal, we may approximate it as stationary when the number of co-channel users is large. Thus the variance of  $n_{rs}(t)$  is simplified to

$$\text{Var}[n_{rs}(t)] = 2e^2 G^{2+x} \frac{\eta}{h\nu} \left\{ \sum_{j=0}^N I_{cj} + I_b \right\} B \quad (4.25)$$

Since  $B \cdot T_b \gg 1$  is assumed, frequency component higher than  $1/T_b$  will be eliminated by the integration in (4.24), then

$$\text{Var}[n_s(t)] = 2e^2 G^{2+x} \frac{\eta}{h\nu} \left\{ \sum_{j=0}^N I_{cj} + I_b \right\} \frac{1}{T_b} \quad (4.26)$$

Now we are ready to give the error rate expression. Since it is evident that  $s(t)$  is antipodal and that each noise component  $n_i(t)$ ,  $n_s(t)$ , and  $n_t(t)$  in  $i_d(t)$  is mutually independent, the decision error rate  $P_e$  is simply expressed by a complement error function  $\text{erfc}()$ .

$$P_e = \frac{1}{2} \text{erfc} \sqrt{\gamma} \quad (4.27)$$

$$\gamma = \frac{(\alpha I_{c0})^2}{\text{Var}[n_i(T_b)] + \text{Var}[n_s(T_b)] + \text{Var}[n_t(T_b)]} \quad (4.28)$$

## 4.5 Numerical Results

### 4.5.1 Error Rates and Overload Characteristics

Table 1 shows the parameters used in numerical calculations. First, error rate performance on multiple access is calculated as

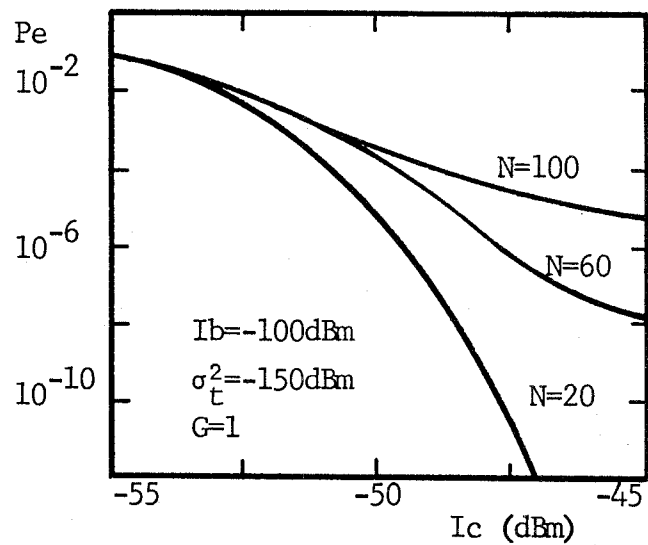


Fig. 4.6 Error rate performance

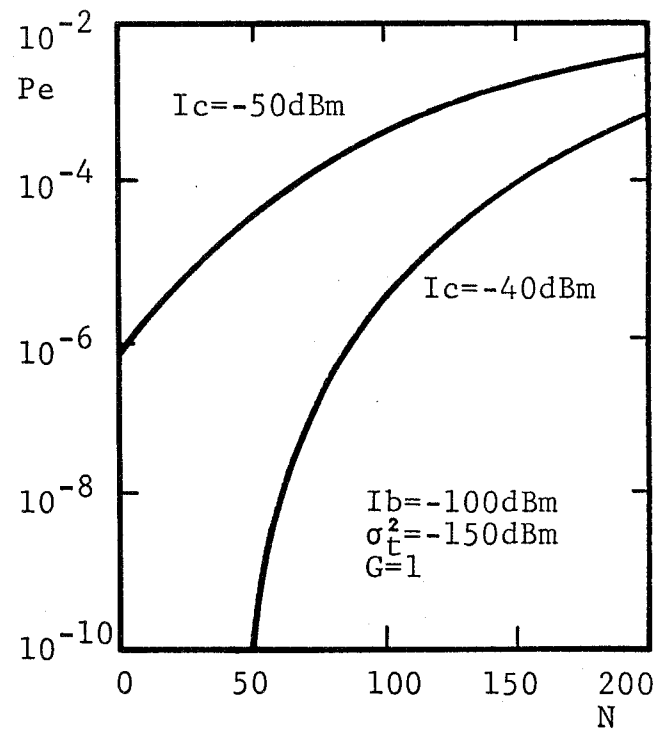


Fig. 4.7 Overload characteristic

a function of received desired signal intensity  $I_C$  (Fig. 4.6). For simplicity of analysis, the received signal intensity of each user  $I_{Cj}$  is equated to that of the desired signal  $I_{C0}$ . So the subscript on  $I_{C0}$  is omitted ( $I_C$ ).

Next, error rate as a function of the number of co-channel users  $N$  is depicted in Fig. 4.7 to show the overload characteristic. As shown in Fig. 4.7, SSMA has a desirable characteristic so-called "graceful degradation" that as the number of co-channel users increases, the error performance is not deteriorated suddenly.

#### 4.5.2 Required Optical Power

In designing SSMA system, data that describe the required received optical power as a function of the number of co-channel users in order to attain the predetermined transmission quality are valuable. In this section, the required signal intensity on multiple access is presented.

Suppose that a desired signal,  $N$  equal-powered co-channel users' signals, and ambient light of intensity  $I_b$  be received. The required signal intensity  $I_{CR}$  to attain a fixed value of the error rate  $P_{eR}$  is given as follows using (4.27).

$$\gamma_R = \{\text{erfc}^{-1}(2P_{eR})\}^2 \quad (4.29)$$

$$\gamma_R = \frac{(\alpha I_{CR})^2}{\frac{(\alpha I_{CR})^2 N}{q} + \beta\{(N+1)I_{CR} + I_b\} + \sigma_t^2} \quad (4.30)$$

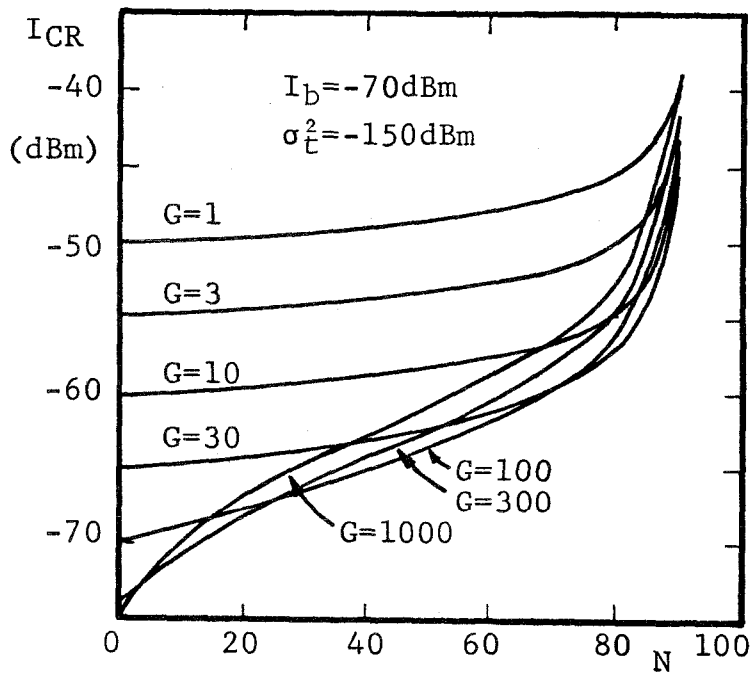


Fig. 4.8 Required optical power

(a)  $I_b = -70\text{dBm}$ ,  $\sigma_t^2 = -150\text{dBm}$

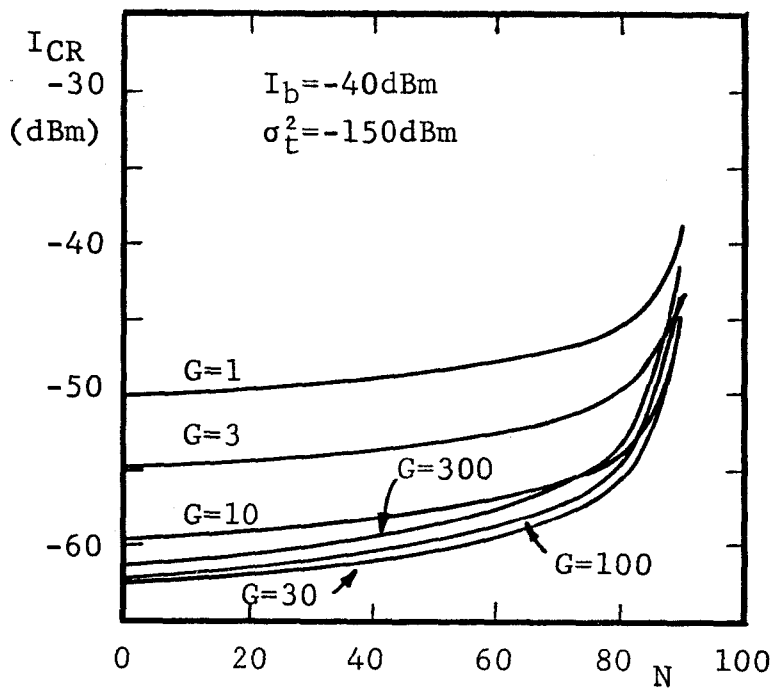


Fig. 4.8 Continued

(b)  $I_b = -40 \text{ dBm}$ ,  $\sigma_t^2 = -150 \text{ dBm}$

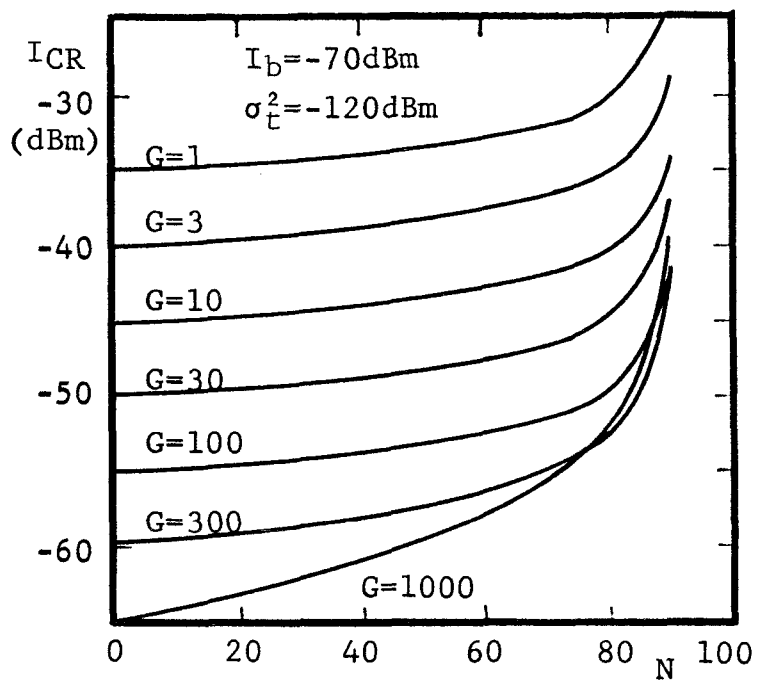


Fig. 4.8 Continued

(c)  $I_b = -70 \text{ dBm}$ ,  $\sigma_t^2 = -120 \text{ dBm}$



where

$$\beta = 2e^2 G^{2+x} \frac{\eta}{h\nu} \cdot \frac{1}{T_b}$$

$$\sigma_t^2 = \text{Var}[n_t(T_b)]$$

From (4.30),

$$I_{CR} = \frac{(N+1)\gamma_R \beta + \sqrt{(N+1)^2 \gamma_R^2 \beta^2 + 4\alpha^2 \{1 - \frac{N\gamma_R}{q}\} \gamma_R \{\sigma_t^2 + \beta I_b\}}}{2\alpha^2 \{1 - \frac{N\gamma_R}{q}\}} \quad (4.31)$$

Fig. 4.8 depicts  $I_{CR}$  in (4.31) as a function of  $N$  on  $P_{eR}=10^{-6}$ . These figures show distinctive features depending on the following situation concerning the dominant noise source.

- (a) Shot noise due to the received signal is dominant.
- (b) Shot noise due to ambient light is dominant.
- (c) Thermal noise is dominant.

In Fig. 4.8,  $I_{CR}$  approaches infinity at  $N=90$ . This indicates that more than 90 simultaneous multiple access is impossible on  $P_{eR}=10^{-6}$ . The maximal number of simultaneous users  $N_{max}$  is given by first solving (4.30) on  $N$ , and then taking limitation  $I_{CR} \rightarrow \infty$ .

$$N = \frac{\frac{\alpha^2 I_{CR}^2}{\gamma_R} - \beta(I_{CR} + I_b) - \sigma_t^2}{\frac{\alpha^2 I_{CR}^2}{q} + \beta I_{CR}} \quad (4.32)$$

$$N_{max} = \lim_{I_{CR} \rightarrow \infty} N = \frac{q}{\gamma_R} = \frac{q}{\{\text{erfc}^{-1}(2P_{eR})\}^2} \quad (4.33)$$

Fig. 4.9 shows  $N_{max}$  normalized by the spreading ratio  $q$  as a function of  $P_{eR}$ . From Fig. 4.9, minimum spreading ratio  $q$  to satisfy the required transmission quality and the required number of simultaneous users can be derived.

#### 4.5.3 Optimization of APD Gain

As shown in Fig. 4.8, the required optical power  $I_{CR}$  strongly depends on the APD gain  $G$ . In order to clarify the dependency of  $I_{CR}$  on  $G$ ,  $I_{CR}$  as a function of  $G$  is shown in Fig. 4.10. As described in [12], it is shown that, from Fig. 4.10, an optimum value of APD gain exists.

Generally in SSMA, the optimization may be accomplished by the following three strategies.

- (i) To minimize  $I_{CR}$  with the given  $P_{eR}$  and  $N$ .
- (ii) To maximize  $N$  with the given  $P_{eR}$  and  $I_C$ .
- (iii) To minimize  $P_e$  with the given  $I_C$  and  $N$ .

These optimization are accomplished theoretically by first

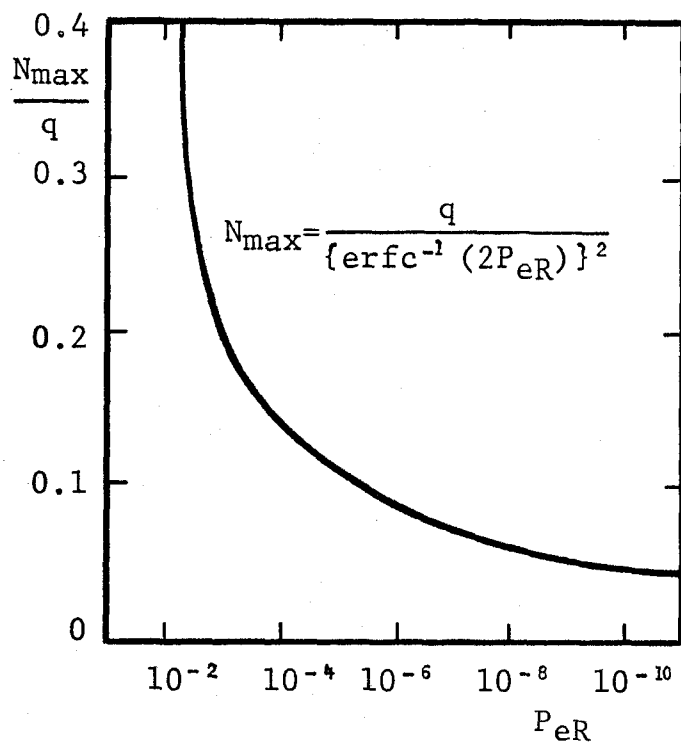


Fig. 4.9 Maximum number of simultaneous users

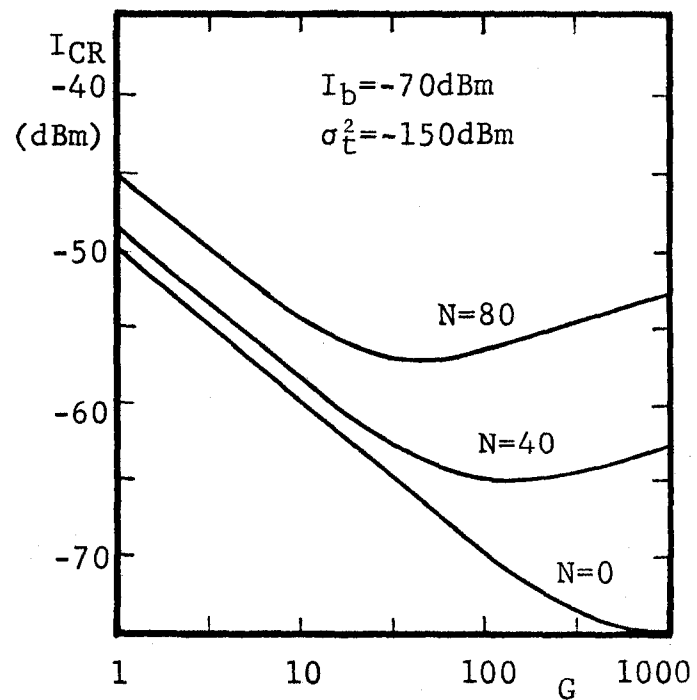


Fig. 4.10 Optimization of APD gain

differentiating (4.31), (4.32), and (4.27), respectively, on  $G$ , and then letting the obtained equations be equated to zero. All these equations are  $(2+x)$ th order, and difficult to solve analytically. However, especially in case of (iii), the optimum gain  $G_{opt}$  is derived easily as follows.

$$\frac{\partial P_e}{\partial G} = \frac{\partial}{\partial G} \left\{ \frac{1}{2} \text{erfc} \sqrt{\gamma} \right\} = 0 \quad (4.34)$$

$$G_{opt} = \left[ \frac{2\sigma_t^2}{2e^2 \frac{\eta}{h\nu} \cdot \frac{1}{T_b} \cdot x \cdot \{(N+1)I_c + I_b\}} \right]^{\frac{1}{2+x}} \quad (4.35)$$

In designing SSMA, we may estimate  $N_{max}$  and/or  $I_{CR}$  according to the strategy (i) or (ii). Also while SSMA is operating and the optical signal intensity varies with time, we could determine APD gain dynamically according to the strategy (iii).

## 4.6 Synchronization Systems

### 4.6.1 Cross-Correlation Property

While the propagation delay  $\tau$  between transmitter and receiver was assumed to be known in the previous discussion, it is synchronization system that estimates  $\tau$ . The synchronization system is generally divided roughly into the acquisition subsystem and the tracking subsystem. Acquisition is to reduce the estimation error to a fixed small range (for example, less

than  $\pm T_c$ ), and tracking is to keep the error within the range.

The synchronization on DS-SS system is performed taking advantage that the cross-correlation function of received signal and local PN sequence has a sharp peak at its origin. On the contrary, binary PPM waveform instead of general NRZ waveform is assumed in the proposed modulation format. Thus it is not trivial whether or not the cross-correlation of received signal and local PN sequence has a sharp peak. Consequently, it is necessary to examine the cross-correlation function of  $\hat{i}_r(t)$  and  $\hat{b}_0(t)$ .

In the following discussion, only a single user SS system is considered (i.e.  $N=0$ ) for the sake of simplicity. Therefore the subscripts for distinguishing each users are omitted. Furthermore it is assumed that  $b(t)$  is a pure random code.

First  $\hat{i}_r(t)$  and  $\hat{b}(t)$  are rewritten as follows.

$$\begin{aligned}\hat{i}_r(t) &= \{b(t) + b(t - \frac{1}{2}T_c)\}c(t) \\ \hat{b}(t) &= b(t) \cdot f(t)\end{aligned}\tag{4.38}$$

Here, constants are neglected, and the relation  $a(t) \approx a(t - \frac{1}{2}T_c)$  is used. Moreover, it is assumed that  $a(t)=1$  without loss of generality. The cross-correlation function  $R(\epsilon)$  of  $\hat{i}_r(t)$  and  $\hat{b}(t)$  is defined as

$$R(\epsilon) = \int_0^{T_b} \hat{i}_r(t) \cdot \hat{b}(t+\epsilon) dt\tag{4.39}$$

and the calculated result is

$$R(\epsilon) = \begin{cases} 1 - \frac{|\epsilon|}{T_c}, & |\epsilon| < \frac{1}{2}T_c \\ \frac{|\epsilon|}{T_c} - 1, & \frac{1}{2}T_c \leq |\epsilon| < T_c \\ 0, & \text{elsewhere} \end{cases} \quad (4.40)$$

where  $R(\epsilon)$  is normalized so that  $R(0)=1$ . Fig. 4.11 shows the shape of  $R(\epsilon)$ . See appendix B for the derivation of (4.40).

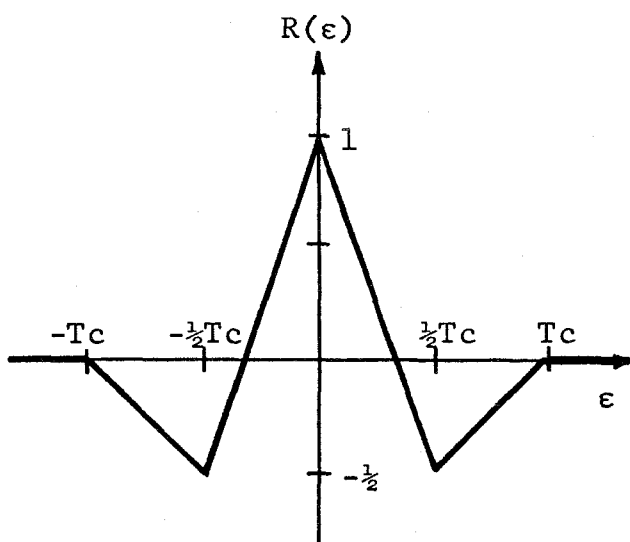


Fig. 4.11 Cross-correlation function  $R(\epsilon)$

#### 4.6.2 Problems on Acquisition

According to Fig. 4.11, the function  $R(\epsilon)$  has a sharp peak around  $\epsilon = 0$ . This implies that an acquisition system for the conventional NRZ waveform can work for the PPM waveform. But there are another problems as follows.

First,  $R(\epsilon)$  is positive only in  $|\epsilon| < \frac{1}{3}T_c$ . Thus  $|\epsilon|$  must be less than  $\frac{1}{3}T_c$  in order to demodulate the signal correctly. On the other hand  $|\epsilon|$  may be less than  $T_c$  for NRZ waveform [8].

Second,  $R(\epsilon)$  have another negative peaks on  $\epsilon = \pm \frac{1}{2}T_c$ . Since the polarity of  $R(\epsilon)$  cannot be utilized on acquisition, the acquisition subsystem must distinguish three peaks on  $\epsilon = 0$ ,  $\epsilon = \pm \frac{1}{2}T_c$ . This means that the noise margin on detecting the maximum peak is reduced.

These two problems both badly affect to the time to complete the acquisition. But this is a weak point caused by the fact that the conventional acquisition system for NRZ is used forcibly for PPM. So the discussion above will not conclude that the use of PPM generally makes it difficult to acquire the desired signal.

Further study on an acquisition subsystem matched to PPM waveform will be necessary.

#### 4.6.3 Discussion on Tracking

A Delay Lock Loop (DLL) [8] is usually used on tracking of a SS signal, whereas the DLL for the proposed modulation format is illustrated in Fig. 4.12.

DLL works to renew  $\hat{\tau}$ , the estimated version of the propagation delay  $\tau$ , so as to make  $e_\epsilon$ , the differentiated value

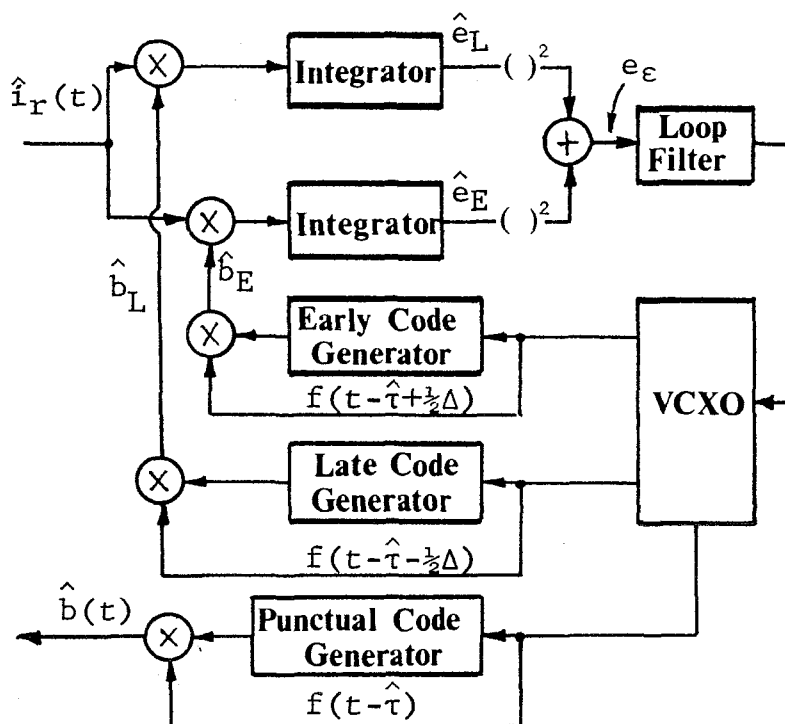
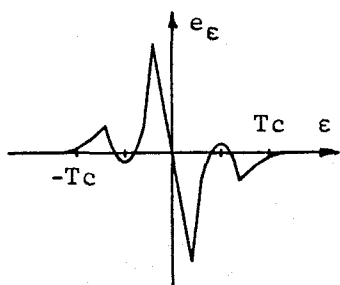
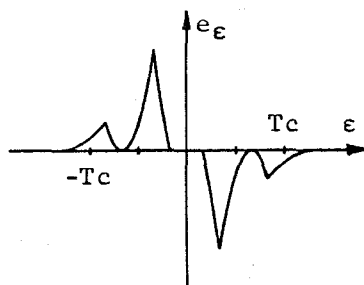


Fig. 4.12 Schematic diagram of DLL

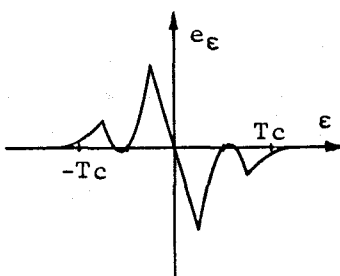




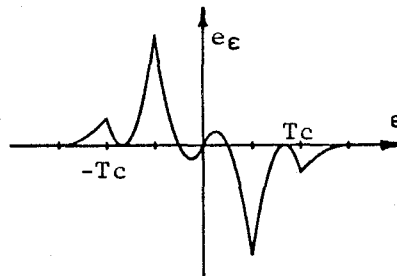
(a)  $\Delta = 0.4T_c$



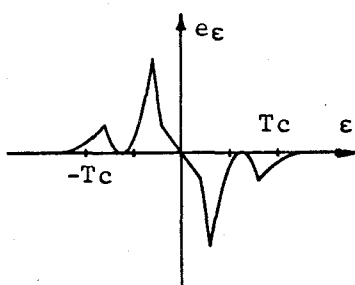
(d)  $\Delta = 0.667T_c$



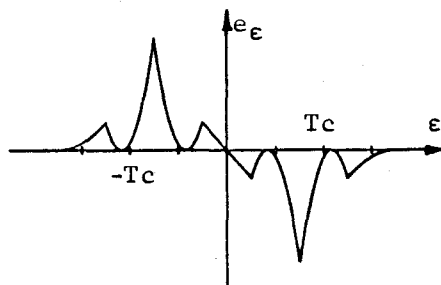
(b)  $\Delta = 0.5T_c$



(e)  $\Delta = T_c$



(c)  $\Delta = 0.6T_c$



(f)  $\Delta = 1.5T_c$

Fig. 4.13 S-curve

of two correlators' output, zero. Each correlator outputs the correlation coefficient of incoming signal and locally generated PN sequence, respectively.

In Fig. 4.12, early-code  $\hat{b}_E$  and late-code  $\hat{b}_L$  are given as

$$\begin{aligned}\hat{b}_E &= b(t-\hat{\tau}+\frac{1}{2}\Delta) \cdot f(t-\hat{\tau}+\frac{1}{2}\Delta) \\ \hat{b}_L &= b(t-\hat{\tau}-\frac{1}{2}\Delta) \cdot f(t-\hat{\tau}-\frac{1}{2}\Delta)\end{aligned}\tag{4.41}$$

where  $\Delta$  is the phase difference between  $\hat{b}_E$  and  $\hat{b}_L$ . The incoming signal  $\hat{i}_r(t)$  is first multiplied by  $\hat{b}_E$  and  $\hat{b}_L$  respectively, and then integrated. The outputs,  $\hat{e}_E$  and  $\hat{e}_L$ , of the integrators are [8],

$$\begin{aligned}\hat{e}_E &= \sqrt{P} \cdot a(t) \cdot R(\tau-\hat{\tau}+\frac{1}{2}\Delta) + n_E(t) \\ \hat{e}_L &= \sqrt{P} \cdot a(t) \cdot R(\tau-\hat{\tau}-\frac{1}{2}\Delta) + n_L(t)\end{aligned}\tag{4.42}$$

where

$$\begin{aligned}n_E(t) &= \frac{1}{T_b} \int_{t-T_b}^t n(t') \cdot \hat{b}_E dt' \\ n_L(t) &= \frac{1}{T_b} \int_{t-T_b}^t n(t') \cdot \hat{b}_L dt'\end{aligned}\tag{4.43}$$

and  $n(t)$  is the sum of the thermal noise and shot noise. Also  $P$  is the averaged power of the desired signal in  $\hat{i}_r(t)$ .

Next  $\hat{e}_E$  and  $\hat{e}_L$  are squared respectively and then

differentiated to generate  $e_{\epsilon}$ .

$$\begin{aligned}
 e_{\epsilon} = & P \cdot a(t)^2 [R(\epsilon - \frac{1}{2}\Delta)^2 - R(\epsilon + \frac{1}{2}\Delta)^2] \\
 & + P \cdot a(t) [n_L(t) \cdot R(\epsilon - \frac{1}{2}\Delta) - n_E(t) \cdot R(\epsilon + \frac{1}{2}\Delta)] \\
 & + n_L(t)^2 - n_E(t)^2
 \end{aligned} \tag{4.44}$$

where  $\epsilon = \tau - \hat{\tau}$ .

The first term of (4.44) determines the shape of so-called S-curve [8] which dominate the dynamics of DLL. While S-curve is a function  $e_{\epsilon}$  of  $\epsilon$ , S-curve for the proposed modulation format varies widely as shown in Fig. 4.13 according to  $\Delta$ .

Since the requirements to S-curve are

- (i) The slope around its origin is large.
- (ii) Absolute value of  $e_{\epsilon}$  is large enough in the wide range of  $\epsilon$ .

Considering these requirement and that the tracking system may cover the range  $|\epsilon| < \frac{1}{3}T_c$  as previously stated,  $\Delta \approx 0.5T_c$  may be appropriate. Although a DLL for NRZ is robust in choosing  $\Delta$ , the DLL for PPM is sensitive.

#### 4.6 Concluding Remarks

We have discussed the possibility of the wireless optical digital SSMA. First, suitable modulation format is studied taking the restrictions of in-house use into account. Then the error rate performance of the proposed modulation format is derived.

Results obtained are as follows.

- (i) A baseband direct sequence signaling with binary PPM waveform is suitable.
- (ii) Fundamental error rate expressions based on Gaussian approximation of shot noise and of co-channel interference are derived.
- (iii) Required received optical power for multiple access is calculated, and the optimization of APD gain is discussed.

Table 4.1 Constants for Numerical Calculations

Wavelength of optical carrier	$\lambda$	0.85 $\mu m$
Quantum efficiency	$\eta$	0.75
Information data rate	$1/T_b$	64k bits/sec
Spreading ratio	$q$	1023
Excess noise factor	$x$	0.35
Required error rate	$P_{eR}$	$1 \times 10^{-6}$

## CHAPTER 5

### CONCLUSIONS

This thesis investigated the multiple access performance of direct sequence spread spectrum systems. Firstly, a modification to the conventional direct sequence SSMA systems was proposed for improving the multiple access capability. Theoretical error rate expressions for the proposed modulation/demodulation schemes, named "unpacked-MFSK/DS" systems, were derived, and the comparison of multiple access capability with other conventional schemes was made. Secondly, a splendid practical application of SSMA to the in-house wireless optical communication systems in office, factory, and home environment was presented. Taking the properties and restrictions on in-house wireless optical channel into account, a suitable modulation scheme for SSMA was discussed, and theoretical error rate expressions for the proposed modulation/demodulation schemes were obtained. Also it was discussed that the optimization of the receiver gain in terms of error probabilities and that the correlative performance of the proposed signaling for synchronization.

The following summarize the principal results obtained in this thesis.

Improvement of multiple access performance:

- (1) A modulation scheme employing M-ary FSK as primary

modulation instead of conventional binary PSK is introduced. A modification to the symbol alignment of MFSK signal for flattening the power spectrum of the transmitted signal over the given bandwidth is presented.

- (2) Theoretical error rate expressions were derived and it was found that the proposed modulation scheme has superior multiple access performance.

#### In-house wireless optical SSMA:

- (1) The suitable modulation scheme was discussed taking the properties and restrictions of in-house wireless optical communication systems into account. In consequence, a baseband direct sequence signaling with binary pulse position modulated waveform was found to be suitable.
- (2) Fundamental error rate expressions necessary for the practical communication link designing were derived. These include the error rate performance, overload characteristic, multiple access capability, and optimization of APD gain.

## APPENDIX A

### SYMBOL ERROR RATE FOR THE NONCOHERENT MFSK DEMODULATOR

The symbol error rate  $P_{SE}$ , when  $\hat{\Phi}_D(f)$  distributes uniformly, is well-known to be [13]

$$P_{SE} = \frac{M-1}{2} \cdot \exp\left(\frac{-k \cdot T_b \cdot S_0}{\hat{\Phi}_D(f)}\right) \quad (\text{A.1})$$

In the proposed systems, however,  $\hat{\Phi}_D(f)$  does not distribute uniformly. Thus it is necessary to derive  $P_{SE}$  under the arbitrary  $\hat{\Phi}_D(f)$ .

Now, suppose symbol  $f_p$  is sent from the transmitter. In this case, the envelope of the output for the  $p$ -th matched-filter obeys the following Rician distribution function [4].

$$Q_p(r) = \frac{r}{J_p} \cdot I_0\left(\frac{\sqrt{2S} \cdot r}{J_p}\right) \cdot \exp\left(-\frac{r^2 + 2S}{2J_p}\right) \quad (\text{A.2})$$

where  $I_0()$  denotes the zero-th order modified Bessel function of the first kind,  $S$  and  $J_p$  denotes the signal power and noise power at the output of the matched filter, respectively. The envelope of the rested  $M-1$  matched filters' output will be expressed as the following Rayleigh distribution function [4].

$$Q_q(r) = \frac{r}{J_q} \cdot \exp\left(-\frac{r^2}{2J_q}\right) \quad (\text{A.3})$$

where  $J_q$  denotes the noise power at the output of the  $q$ -th matched filter.

Symbol decision error occurs when the output of the  $p$ -th matched filter is not maximum among  $M$  filter outputs. This conditional probability when  $p$ -th symbol is sent is

$$\begin{aligned} P_{SE|p} &= 1 - \prod_{\substack{q=1 \\ q \neq p}}^M \left( 1 - \int_0^\infty Q_p(r_1) \cdot \int_{r_1}^\infty Q_q(r_2) dr_2 dr_1 \right) \\ &= 1 - \prod_{\substack{q=1 \\ q \neq p}}^M \left( 1 - \frac{J_q}{J_p + J_q} \cdot \exp \frac{-S}{J_p + J_q} \right) \end{aligned} \quad (\text{A.4})$$

Furthermore,  $P_{SE}$  will be given by averaging (A.4) over all  $p=1,2,\dots,M$ . In addition, since the SNR  $S/J_p$  after the matched filter is equal to the energy contrast ratio  $kT_b S_0 / \hat{\Phi}_D(f)$ ,

$$\begin{aligned} P_{SE} &= \frac{1}{M} \sum_{p=1}^M \left\{ 1 - \prod_{\substack{q=1 \\ q \neq p}}^M \left\{ \frac{\hat{\Phi}_D(f_q)}{\hat{\Phi}_D(f_p) + \hat{\Phi}_D(f_q)} \right. \right. \\ &\quad \left. \left. \cdot \exp \frac{-S_0 \cdot T_b \cdot k}{\hat{\Phi}_D(f_p) + \hat{\Phi}_D(f_q)} \right\} \right\} \end{aligned} \quad (\text{A.5})$$



## APPENDIX B

### DERIVATION OF $R(\epsilon)$

Integration of (4.39) will be rewritten as

$$\begin{aligned}
 R(\epsilon) &= \int_0^T \hat{i}_r(t) \cdot \hat{b}(t+\epsilon) dt \\
 &= \sum_{k=0}^{q-1} \left( \int_{k \cdot T_c}^{(k+1)T_c} b(t) b(t+\epsilon) c(t) f(t+\epsilon) dt \right. \\
 &\quad \left. + \int_{k \cdot T_c}^{(k+1)T_c} b(t - \frac{1}{2}T_c) b(t+\epsilon) c(t) f(t+\epsilon) dt \right) \quad (B.1)
 \end{aligned}$$

Let  $b(t)$  be expressed as

$$b(t) = \sum_{k=0}^{q-1} \bar{b}_k \cdot s(t - k \cdot T_c) \quad (B.2)$$

where

$$s(t) = \begin{cases} 1, & 0 \leq t < T_c \\ 0, & \text{elsewhere} \end{cases} \quad (B.3)$$

$$\bar{b}_k \in \{-1, 1\} \quad (B.4)$$

For the special case  $0 \leq \epsilon < \frac{1}{2}T_C$ , substituting (B.2) into (B.1) yields

$$\begin{aligned}
 R(\epsilon) = & \sum_{k=0}^{q-1} \left\{ \bar{b}_k^2 \int_{k \cdot T_C}^{(k+1)T_C - \epsilon} c(t)f(t+\epsilon) dt \right. \\
 & + \bar{b}_k \bar{b}_{k+1} \int_{(k+1)T_C - \epsilon}^{(k+1)T_C} c(t)f(t+\epsilon) dt \\
 & + \bar{b}_{k-1} \bar{b}_k \int_{k \cdot T_C}^{(k+\frac{1}{2})T_C} c(t)f(t+\epsilon) dt \\
 & + \bar{b}_k^2 \int_{(k+\frac{1}{2})T_C}^{(k+1)T_C - \epsilon} c(t)f(t+\epsilon) dt \\
 & \left. + \bar{b}_k \bar{b}_{k+1} \int_{(k+1)T_C - \epsilon}^{(k+1)T_C} c(t)f(t+\epsilon) dt \right\}
 \end{aligned} \tag{B.5}$$

(see Fig. A.1) Since  $b(t)$  is assumed to be a pure random sequence, the averages of  $\bar{b}_{k-1}\bar{b}_k$  and  $\bar{b}_k\bar{b}_{k+1}$  are zero. Thus (B.5) is simplified by calculating the first and the fourth integrating terms only. Normalizing so as to  $R(0)=1$ ,

$$R(\epsilon) = \sum_{k=0}^{q-1} \left( \frac{1 - \frac{4\epsilon}{T_C}}{2q} + \frac{1 - \frac{2\epsilon}{T_C}}{2q} \right)$$

$$= 1 - \frac{3\epsilon}{T_c}, \quad 0 \leq \epsilon < \frac{1}{2}T_c \quad (\text{B.6})$$

is obtained. In the similar manner, (4.40) is derived by calculating outside the range  $0 \leq \epsilon < \frac{1}{2}T_c$ .

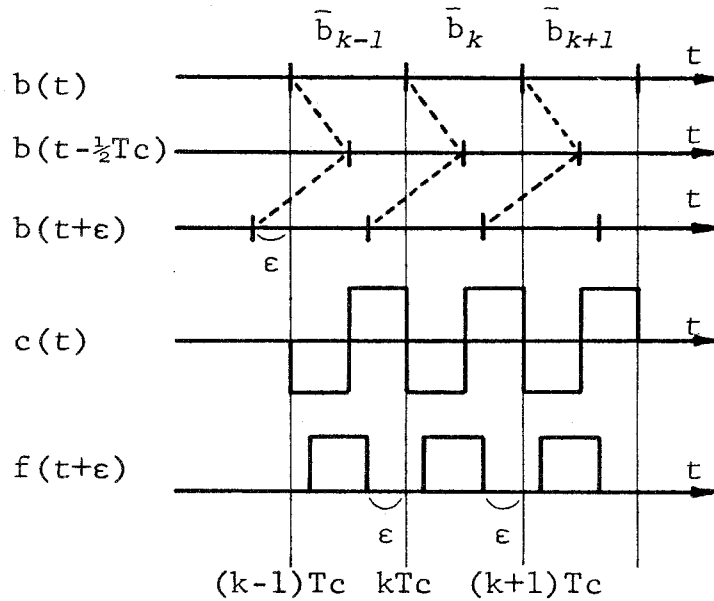


Fig. A.1 Integration of  $c(t)$  and  $f(t+\epsilon)$

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