PAPER *Special Section on Nonlinear Theory and Its Applications*

Effect of Nonlinear Amplification on a Spread Spectrum Signal and Receiver Configurations

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SUMMARY This paper discusses the characteristics of the nonlinearly amplified spread-spectrum (SS) signals. We evaluate the symbol error-rate performance with the conventional receiver, changing the length of the spreading sequence. In addition, we also propose the receiver with MLSE. The configuration of the MLSE for the nonlinearly amplified signals is generally complicated; however we show that the complexity of the MLSE receiver can be reduced, as the number of required reference sequences in the receiver for an SS signal is small. As the result, it is shown that the error rate performance of the nonlinearly amplified SS signal can be improved by this proposed receiver and that the degradation caused by the nonlinear amplification can be made negligibly small with a sufficiently long spreading sequence.

key words: *nonlinear amplification, spread spectrum, radio communication, MLSE*

1. Introduction

There are many communication systems in which the efficient use of the power is important. Mobile communication systems, portable telephones, and wireless LAN systems where batteries are employed as the power supply are examples of these systems. In such systems, the high power amplifiers (HPAs) of transmitters are often used at or near their saturation points which exhibit nonlinear amplification characteristics. This nonlinearity, together with the filter for bandwidth limitation before the HPA, causes the degradation of the error-rate performance.

In recent studies, it is reported that spread-spectrum (SS) signals are robust to nonlinearity of the channel [1]–[4]. This is because the effects of distortion can be reduced by despreading process at the receiver. They discuss the performance of the systems with nonlinear amplification with the fixed-length spreading sequences. In this paper, we show the performance of a QPSK/SS signal; that is, an SS signal employing quadrature phase shift keying, with nonlinear amplification by changing the length of its spreading code.

In order to improve the performance of the nonlinearly amplified non-spread signal, there are several proposals of correlation or maximum likelihood sequence estimation (MLSE) receivers, which makes a decision by comparing incoming signals with references prepared in the receiver. In Ref. [5], the basic aspect of the performance of the correlation receiver is shown; however, the configuration of the receiver is not simple in spite of the very simplified model in which only the effect of the main lobe of each transmitted data pulse is considered and the correlation is made on the basis of symbol by symbol. In Ref. [6], MLSE is discussed for the nonlinearly amplified signal sequences, but its configuration is rather complicated since it has to prepare or generate huge number of symbol sequences and make the comparison of incoming signals with those sequences at the same time.

In SS communication systems, in contrast to the conventional non-spreading systems, the signal is composed of long sequences of spreading codes. Since the number of the possible signal sequences, which are determined by the known spreading codes, is small, the configuration of the receiver becomes much simpler.

In the followings, we first discuss the performance of a nonlinearly amplified SS signal received by a conventional receiver. The MLSE receiver is then defined and its performance is shown. From the results with the conventional and MLSE receivers, we discuss the cause of the performance degradation in the nonlinear transmission system. Following the results of a single SS signal case, we also discuss the case where multiple nonlinearly amplified SS signals are transmitted in the same frequency band at the same time.

2. System Model (Transmitter and Channel)

Figure 1 shows the equivalent low-pass model of a transmitter and channels for QPSK/SS signals. The input to the transmitter of the k-th user is the sequences of binary (± 1) data, where each data pulse has the length T_s. In-phase and quadrature components of the m-th transmitted data are denoted by $b^I_{k,m}$ and $b^Q_{k,m}$, respectively. These sequences of data are then multiplied by the spreading sequences $a^I_{k,n}$ and $a^Q_{k,n}$ with chip length T_c, where n is integer in the range of $-\frac{N-1}{2}$ to $\frac{N-1}{2}$ and N(odd number) is the length of the spreading sequences, resulting $T_s = NT_c$.

After the spreading and QPSK modulation, the sig-

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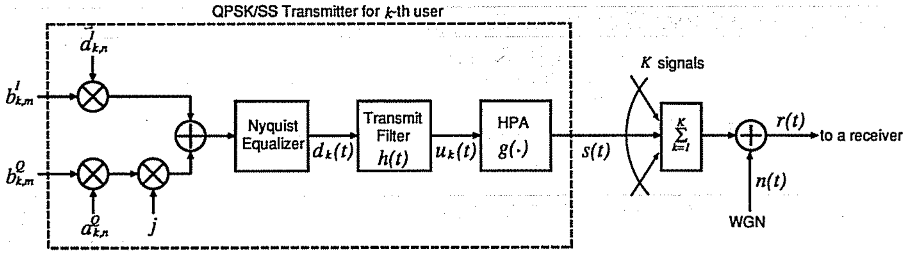

Fig. 1 System model (transmitter and channel).

nal is filtered by the transmit filter with the impulse response $h(t)$ preceded by the Nyquist equalizer and yields

$$u_k(t) = \sum_{m=-\infty}^{+\infty} \sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} (b^I_{k,m}a^I_{k,n} + jb^Q_{k,m}a^Q_{k,n}) \cdot h(t - nT_c - mT_s). \quad (1)$$

Note that this is the conventional QPSK if $N=1$. The filter output $u_k(t)$ is then nonlinearly amplified by the HPA whose input-output characteristic is represented by a function $g(\cdot)$.

In this paper, the number of the simultaneous transmitted signal is expressed by K. We consider that these signals are asynchronous, and thus each signal has independent delay τ_k and initial carrier phase ϕ_k, where the subscript k stands for "the k-th user". The distributions of τ_k and ϕ_k are assumed to be uniform over $[0, NT_c]$ and $[0, 2\pi]$, respectively.

The receiver receives these K QPSK/SS signals with Additive White Gaussian noise (AWGN).

3. Conventional Receiver

The configuration of the conventional receiver discussed in this paper is shown in Fig. 2. In this receiver, the incoming signal $r(t)$ is first band-limited by the receive filter which has the same frequency response as that of the transmit filter. The output of the filter is then demodulated and sampled at every T_c seconds. The samples of in-phase and quadrature components are then multiplied by the spreading codes and summed up for a cycle of the code, N. The outputs of this despreading procedure are then applied to the decision circuits.

In order to make the discussion simple and the effect of the nonlinear amplification clear, in this section, we consider the case where there is only one user ($K=1$) in the system. Thus the signal at the receiver is given by

$$r(t) = g\Big(u_k(t)\Big) + n(t) \quad (2)$$

where the first term is the signal and $n(t)$ denotes the equivalent low-pass model of AWGN with power spectral density $N_0/2$.

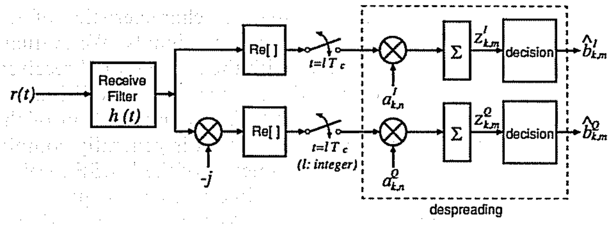

Fig. 2 Conventional receiver.

After despreading process for m-th bit, the sample of in-phase component $Z^I_{k,m}$ is expressed as

$$\begin{aligned} Z^I_{k,m} &= \sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} a^I_{k,n}\mathrm{Re}\Big[h(t) \otimes r(t)\Big] \\ &= \sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} a^I_{k,n}\mathrm{Re}\Big[h(t) \otimes g\Big(u_k(t)\Big)\Big] \\ &\quad + \sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} a^I_{k,n}\mathrm{Re}\Big[h(t) \otimes n(t)\Big] \\ &= Z^I_{D,m} + Z^I_{N,m}, \end{aligned} \quad (3)$$

where the notation $\otimes$ denotes convolution.

In this equation, first term, $Z^I_{D,m}$, is desired signal component and the second term is the noise component. It is easy to show analytically that the latter is a random variable of zero-mean Gaussian distribution with the variance of $\sigma^2 = N_0/(NT_c)$. On the other hand, the former is difficult to be expressed in close form because of the interaction of the inter-symbol interference (ISI) and the nonlinear amplification. The value of $Z^I_{k,m}$ varies even for the same data $b^I_{k,m}$ according to the preceding and succeeding data symbols. We thus derive the value of $Z^I_{k,m}$ by computer simulation considering the effects of ISI and use the result for the the evaluation of the symbol error rate performance as follows.

For a given sequence of the preceding and succeeding symbols, the value $Z^I_{D,m}$ can be calculated by computer simulation. Since the bit error on in-phase or quadrature component of QPSK can be regarded as that of BPSK, the bit error rate performance for $b^I_{k,m}$ can be given as

Table 1 The assumptions for the simulation.

Spreading Sequence	Gold Sequence
Transmit Filter (ISI considerd)	Ideal BPF with bandwidth $1/T_c$ (preceding 10 pulses) (succeeding 10 pulses)
Nonlinearity of the HPA	BPHL
Receive Filter	Ideal BPF with bandwidth $1/T_c$
Synchronization	Perfect

$$BER' = \mathrm{erfc}\left(\frac{Z^I_{D,m}}{\sqrt{2}\cdot\sigma}\right), \quad (4)$$

where erfc() represents complementary error function,

$$\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}}\int_x^\infty e^{-t^2}dt. \quad (5)$$

In Eq. (4), without loss of generality, $b^I_{k,m} = 1$ is assumed.

Since the performance for the quadrature component is considered to be same as that for the in-phase component and error events on both components are independent, the symbol error rate performance $P_e^{''}$, becomes

$$P_e^{'} = 2BER' - (BER')^2. \quad (6)$$

This is then averaged over all possible patterns of the data which give the effect for the observed symbol, and the symbol error rate performance (P_e) is expressed as

$$P_e = \mathrm{E}[P_e^{'}], \quad (7)$$

where E[ ] denotes expectation.

The assumptions for the simulation is shown in Table 1. In the simulation, we assume that the nonlinearity of HPA is represented by Band-Pass Hard Limiter (BPHL) expressed as

$$g(x) = \begin{cases} -1 & x < 0 \\ 0 & x = 0 \\ 1 & 0 < x. \end{cases} \quad (8)$$

This is one of typical nonlinearities and is the good approximation of distorted nonlinear amplifiers such as a class-C amplifier [4]. In order to make the effect of ISI clear, we employ a simple ideal BPF. As discussed in Ref. [3], this is the filter which gives large affection to the system performance of SS signals. As is shown in the table, the effect of each ten pulses, preceding and succeeding, is considered. In other words, the effect of each pulse is truncated in the range of ±10 side lobes. We have confirmed that this truncation does not give any significant difference in the numeric results.

The numeric examples of the performance of the conventional receiver is shown in Fig. 3, where E_b is the energy per one bit, expressed as

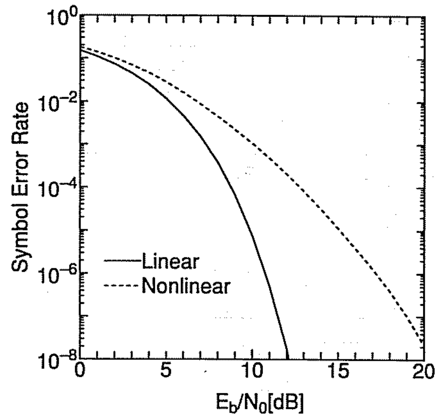

Fig. 3 (a) Symbol error rate (N=1: QPSK).

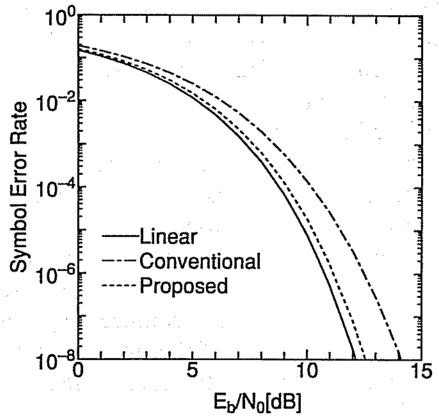

Fig. 3 (b) Symbol error rate (N=7).

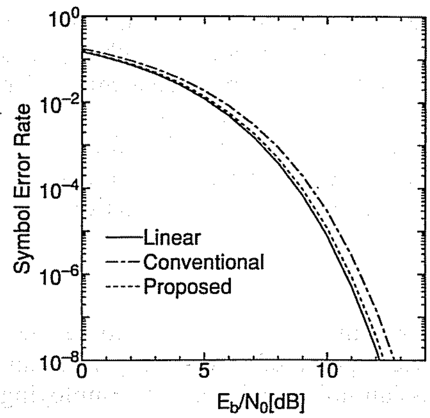

Fig. 3 (c) Symbol error rate (N=31).

$$E_b = \frac{1}{4}\int_{(m-1/2)T_s}^{(m+1/2)T_s} |g(u_k(t))|^2 dt. \quad (9)$$

Figure 3 (a) shows the performance of an SS signal of $N = 1$; that is, conventional QPSK signal. This figure shows large degradation of the performance caused by nonlinear amplification of HPA. When N becomes larger, the performance of QPSK/SS signal with the conventional receiver becomes better as shown in

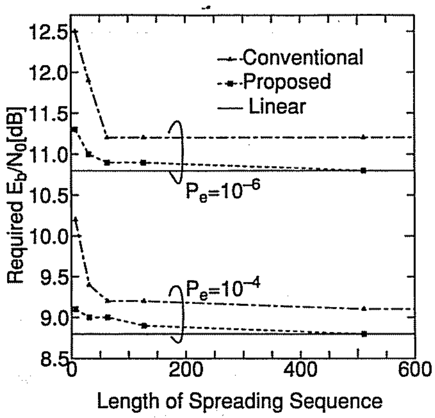

Fig. 4 Required E_b/N_0 (single user).

Figs. 3 (b), (c). This is shown more clearly in Fig. 4, where the E_b/N_0 necessary to achieve $P_e = 10^{-4}$ or $= 10^{-6}$ is given as the function of N. From this figure, we can observe that the performance with the conventional receiver improves according to the increase of N. This performance improvement can be explained as the effect of spreading out the distortion components caused by the nonlinear amplification.

Although there is a great performance improvement by the employment of spread spectrum technique, from Fig. 4, it is clear that the performance degradation still remains. The performance of the nonlinearly amplified signal is worse than that of the ideal linearly amplified one even with very large N.

There are two factors as the cause of this residual performance degradation. The first one is the distortion of the signal constellation, which reduces the Euclidean distance between the symbols; the other one is the loss of receiving power caused by mismatching of the receive filter to the received signal. Among these two factors, the receiving end can do nothing for the former; on the other hand, the latter can be compensated by using MLSE discussed in the following section.

4. MLSE Receiver

The receiver with MLSE is known to be the optimum from the view point of error-rate performance under white Gaussian noise environment. Employing this, we can compensate the loss of the received power of nonlinearly amplified signal.

In the MLSE receiver, the received signal is multiplied by the reference signals which should be prepared in the receiver for all possible sequences. As each pulse of the non-spread QPSK signal takes the value arbitrarily, the number of these references according to the observation duration, increases exponentially and the length of the reference becomes long. Thus the realization of MLSE becomes difficult. On the contrary, when SS technique is employed, the number of necessary references becomes small and the configuration of MLSE

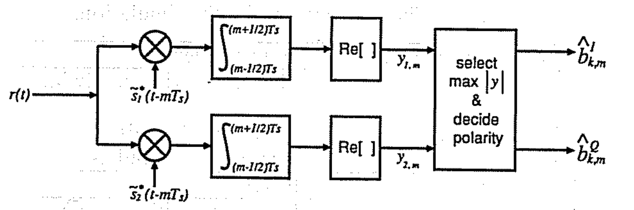

Fig. 5 Proposed receiver.

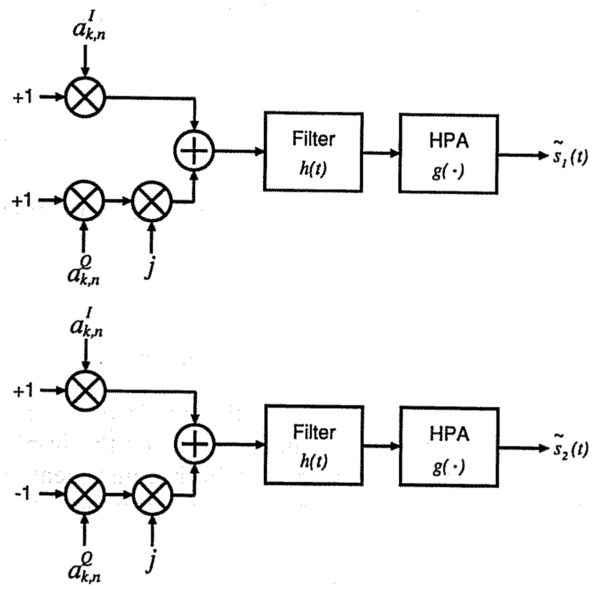

Fig. 6 Generator of $\tilde{s}_1(t)$ and $\tilde{s}_2(t)$.

receiver becomes much simpler.

The configuration of the proposed receiver is shown in Fig. 5. In this receiver, the incoming signal is compared with the reference signals $\tilde{s}_1^*$ and $\tilde{s}_2^*$, which are prepared considering the effect of the nonlinear amplification at the transmitting end. Figure 6 shows the configuration of generators of $\tilde{s}_1$ and $\tilde{s}_2$. The reference signals for the m-th symbol duration are given by

$$\tilde{s}_1^*(t-mT_s) = \begin{cases} g^*\left(\sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} (a_{k,n}^I + j a_{k,n}^Q) \cdot h(t-nT_c-mT_s) \right) & ; (m-1/2)T_s \leqq t \leqq (m+1/2)T_s \\ 0 & ; \text{otherwise} \end{cases} \quad (10)$$

$$\tilde{s}_2^*(t-mT_s) = \begin{cases} g^*\left(\sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} (a_{k,n}^I - j a_{k,n}^Q) \cdot h(t-nT_c-mT_s) \right) & ; (m-1/2)T_s \leqq t \leqq (m+1/2)T_s \\ 0 & ; \text{otherwise} \end{cases} \quad (11)$$

Table 2 Decision.

levels of $y_{1,m}$ and $y_{2,m}$	$(\hat{b}^I_{k,m}, \hat{b}^Q_{k,m})$
$\lvert y_{1,m}\rvert > \lvert y_{2,m}\rvert$ and $y_{1,m} > 0$	(+1, +1)
$\lvert y_{1,m}\rvert > \lvert y_{2,m}\rvert$ and $y_{1,m} < 0$	(−1, −1)
$\lvert y_{1,m}\rvert < \lvert y_{2,m}\rvert$ and $y_{2,m} > 0$	(+1, −1)
$\lvert y_{1,m}\rvert < \lvert y_{2,m}\rvert$ and $y_{2,m} < 0$	(−1, +1)

where the notation $*$ means conjugate. Note that these references are time limited in their symbol duration and thus there are small differences between the received signal waveform and references, and that this difference becomes relatively small if N increases. The decision is made by the absolute values and the polarity of the resultant $y_{1,m}$ and $y_{2,m}$ as is shown in Table 2.

In this receiver, the inputs to the decision circuit, $y_{1,m}$ and $y_{2,m}$, are expressed as

$$\begin{aligned} y_{j,m} &= \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} r(t)\tilde{s}_j^*(t-mT_s)dt\right] \\ &= \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} g\Big(u_k(t)\Big)\tilde{s}_j^*(t-mT_s)dt\right] \\ &\quad + \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} n(t)\tilde{s}_j^*(t-mT_s)dt\right] \end{aligned} \quad (12)$$

where j is 1 or 2. In this equation, the first term is the desired signal component, and the second term is the noise component. We represent the first term by $s_{j,m}$ and the second term by $n_{j,m}$. The signal components, $s_{1,m}$ and $s_{2,m}$ are the deterministic constant values for the given spreading sequences and transmit symbols. On the other hand, the noise components, $n_{j,m}$ are the random values. Since the noise components are the cross correlations between AWGN, $n(t)$, and corresponding reference signals, $\tilde{s}_j^*(t-mT_s)$, their joint distribution function, $p_n(n_{1,m}, n_{2,m})$, becomes zero-mean two dimensional Gaussian. The derivation of the auto and cross covariance of the noise components are shown in Appendix. As the result, $p_n(n_{1,m}, n_{2,m})$ is expressed as

$$\begin{aligned} &p_n(n_{1,m}, n_{2,m}) = \\ &\frac{T_s^2}{2\pi N_0\sqrt{16E_b^2-\gamma^2T_s^2}}\exp\left\{-\frac{2E_bT_s^2}{N_0(16E_b^2-\gamma^2T_s^2)}\right. \\ &\left.\cdot\left(n_{1,m}^2+n_{2,m}^2-\frac{\gamma T_s n_{1,m}n_{2,m}}{2E_b}\right)\right\}, \end{aligned} \quad (13)$$

where γ is given by (A· 10)

$$\gamma=\frac{1}{T_s}\mathrm{Re}\left[\int_{-1/2T_s}^{1/2T_s}\tilde{s}_1(t)\tilde{s}_2^*(t)dt\right]. \quad (14)$$

The inputs to the decision circuit, $y_{1,m}$ and $y_{2,m}$, are thus also random values, which have the distribution $p(y_{1,m}, y_{2,m})$ expressed as

$$p(y_{1,m}, y_{2,m}) = p_n(n_{1,m}+s_{1,m}, n_{2,m}+s_{2,m}). \quad (15)$$

Using this distribution, the symbol error rate performance (P_e'), for given preceding and succeeding symbols becomes as

$$\begin{aligned} P_e' &= \int_{-\infty}^{0}\int_{-\infty}^{+\infty} p(y_{1,m}, y_{2,m})dy_{2,m}dy_{1,m} \\ &+ \int_{0}^{+\infty}\int_{-\infty}^{-y_{1,m}} p(y_{1,m}, y_{2,m})dy_{2,m}dy_{1,m} \\ &+ \int_{0}^{+\infty}\int_{y_{1,m}}^{+\infty} p(y_{1,m}, y_{2,m})dy_{2,m}dy_{1,m}. \end{aligned} \quad (16)$$

As in the previous section, if we calculate the error rate P_e' for all possible symbol data sequences and take the average of them over these sequences, we have the average symbol error rate P_e under the same condition used in Sect. 3, which is shown in Fig. 3 and Fig. 4.

From these figures, we can confirm that the performance of the proposed receiver is superior to that of the conventional one and is almost same as that for the linearly amplified signal when N becomes sufficiently large.

The MLSE receiver could remove the effects of power loss caused by mismatching of receiver to the incoming nonlinearly amplified signal perfectly, if the references and received signal are exactly same. There is, however, difference between these waveforms, since we do not take account of the effect of preceding and following symbols for the reference signal. As the length of this difference is constant even when N changes, the effects of this mismatching becomes smaller when N becomes larger. For this reason, the power loss becomes negligibly small, when N is sufficiently large. This is the reason why the the performance depends on N.

As described above, under sufficiently large N, the nonlinearly amplified signal can be received by the proposed receiver without the affection of nonlinear amplification. Therefore, we can conclude that the effect of power loss of the received signal at the receiver is a dominant factor of the performance degradation caused by nonlinear amplification for the conventional receiver.

5. Error Rate Performance with an Interfering Signal

Now let us consider the performance where multiple signals are transmitted at the same time. In this section, we discuss the simplest case where two signals arrive at the receiver with the same power. In this case, the input signal to the receiver, $r(t)$, is expressed as

$$\begin{aligned} r(t) &= g(u_D(t-\tau_D))e^{j\phi_D} \\ &\quad + g(u_U(t-\tau_U))e^{j\phi_U} + n(t), \end{aligned} \quad (17)$$

where subscripts D and U represent "desired" and "undesired": the first and the second terms of this equation are desired signal and interfering signals, respectively.

When the signal (17) is applied to the conventional receiver discussed in Sect. 3, the sample $Z_{k,m}^I$, after despreading process for m-th bit of in-phase component is expressed as

$$Z_{k,m}^I = Z_{D,m}^I + Z_{U,m}^I + Z_{N,m}^I, \tag{18}$$

where $Z_{D,m}^I$ and $Z_{N,m}^I$ are signal and noise components, respectively. The second term $Z_{U,m}^I$ represents interference component which is denoted as

$$Z_{U,m}^I = \sum_{n=-\frac{N-1}{2}}^{+\frac{N-1}{2}} a_{k,n}^I \mathrm{Re}\Big[h(t) \otimes g\Big(u_U(t-(\tau_U-\tau_D))\Big)e^{j(\phi_U-\phi_D)}\Big]. \tag{19}$$

Since the difference of the initial phase $\phi_U - \phi_D$, and of the delay $\tau_U - \tau_D$, are random variables, $Z_{U,m}^I$ itself takes random value. We define the distribution of $Z_{U,m}^I$ as $p(Z_{U,m}^I)$ and calculate it by computer simulation considering the cross-correlation of the codes for desired and interfering signals. Replacing $Z_{D,m}^I$ of (4) with $Z_{D,m}^I + Z_{U,m}^I$ and taking the average on this distribution, we can express the bit error rate BER as

$$BER = \frac{1}{2}\int_{-\infty}^{+\infty} p(Z_{U,m}^I) \cdot \mathrm{erfc}\left(\frac{Z_{D,m}^I + Z_{U,m}^I}{\sqrt{2}\cdot\sigma}\right) dZ_{U,m}^I, \tag{20}$$

and symbol error rate as

$$P_e = 2BER - BER^2. \tag{21}$$

When the signal (17) is the input to the proposed receiver of Sect. 4, the inputs to the decision circuit, $y_{1,m}$ and $y_{2,m}$, are

$$\begin{aligned} y_{j,m} &= \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} r(t)\tilde{s}_j^*(t-mT_s)dt\right] \\ &= \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} g(u_D(t))\tilde{s}_j^*(t-mT_s)dt\right] \\ &\quad +\frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} g\Big(u_U(t-(\tau_U-\tau_D))\Big) \cdot e^{j(\phi_U-\phi_D)}\tilde{s}_j^*(t-mT_s)dt\right] \\ &\quad +\frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} n(t)\tilde{s}_j^*(t-mT_s)dt\right] \\ &= s_{j,m} + i_{j,m} + n_{j,m}, \end{aligned} \tag{22}$$

where the second term $i_{j,m}$ is the interference component. The values of this components for $j=1$ and 2 are random and have joint probability density function $p_i(i_{1,m}, i_{2,m})$. We calculate this distribution by computer simulation, and, derive the symbol error rate performance P_e as follows

$$\begin{aligned} P_e = \int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} & p_i(i_{1,m}, i_{2,m}) \\ \cdot\Bigg\{ & \int_{-\infty}^{0}\int_{-\infty}^{+\infty} p(y_{1,m}+i_{1,m}, y_{2,m}+i_{2,m})dy_{2,m}dy_{1,m} \\ & + \int_{0}^{+\infty}\int_{-\infty}^{-y_{1,m}} p(y_{1,m}+i_{1,m}, y_{2,m}+i_{2,m})dy_{2,m}dy_{1,m} \\ & + \int_{0}^{+\infty}\int_{y_{1,m}}^{+\infty} p(y_{1,m}+i_{1,m}, y_{2,m}+i_{2,m})dy_{2,m}dy_{1,m} \\ \Bigg\} & di_{1,m}di_{2,m}. \end{aligned} \tag{23}$$

Figure 7 shows the numerical results of the required E_b/N_0 versus the length of the spreading sequences. Although the effect of white noise can not be compensated by the spectrum spreading technique, the effect of interference can be reduced. Thus, it is shown that the performance depends on the length of the code even without nonlinear amplification. If we concentrate our attention on the comparison of the performances of linearly amplified case and between two types of the receivers under nonlinear amplification, almost the same conclusion as in the previous sections can be drown; e.g., the performance of the proposed receiver exceeds the conventional receiver. We can also confirm that the performance improves according to the increase of N; however, there is a difference. When interfering signal exists, even with very large N, slight performance degradation caused by nonlinear amplification still remains. This residual degradation even with the proposed receiver can be explained as follows.

The effect of the interference depends on the cross-correlation between desired and undesired (interfering) signals. In this paper, we employ the Gold sequence as the spreading code. This code has excellent cross-correlation characteristic; however, when the signal is

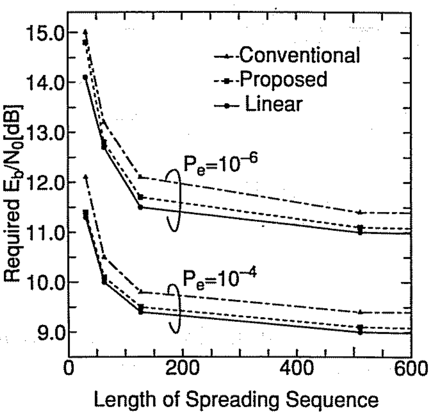

Fig. 7 Required E_b/N_0 (2 users).

nonlinearly amplified, the cross-correlation feature is also distorted and the performance is degraded. This is the cause of the residual degradation.

This performance degradation caused by nonlinearly amplified interfering signals may tend to be large when the number of interfering signal increases. Therefore, the effect of the interference should be considered in the design of the receiver in the multiple access system with nonlinear amplification.

6. Conclusion

In this paper, the characteristics of the nonlinearly amplified spread spectrum signals have been discussed. With the conventional receiver, we show that the performance is improved according as the increase in the code length and that there is still a residual performance degradation caused by the nonlinear amplification. We then proposed a simple MLSE receiver. In this receiver, the power loss caused by mismatching of the receive filter to the received signal can be become negligibly small for the sufficiently long spreading sequences. The resultant symbol error rate performance indicates that the performance for the proposed receiver is better than that for the conventional receiver and the degradation of the error rate performance can be reduced when the length of the spreading code is sufficiently long. The result suggests that the power loss caused by mismatching of the receive filter to received signal is a dominant factor of the performance degradation caused by nonlinear amplification. We also discuss the case where a nonlinearly amplified interfering signal exists. In this case, even with the proposed receiver, nonlinear amplification affects the performance.

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Appendix

As defined in (12), the noise components of the inputs to decision circuit of MLSE reciever are

$$n_{j,m} = \frac{1}{T_s}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} n(t)\tilde{s}_j^*(t-mT_s)dt\right], \quad \text{(A· 1)}$$

where $j=1,2$. As $n(t)$ in the equation is AWGN, $n_{j,m}$ become zero mean Gaussian random values.

Let

$$n(t) = n_C(t) + jn_S(t) \quad \text{(A· 2)}$$

and

$$\begin{aligned}\tilde{s}_j^*(t) &= (\tilde{s}_{jC}(t) + j\tilde{s}_{jS}(t))^* \\ &= \tilde{s}_{jC}(t) - j\tilde{s}_{jS}(t).\end{aligned} \quad \text{(A· 3)}$$

Then the auto covariance of the noise components become

$$\begin{aligned}\mu_{jj} &= \mathrm{E}[(n_{jm})^2] \\ &= \frac{1}{T_s^2}\mathrm{E}\left[\left\{\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s}(n_C(t)\tilde{s}_{jC}(t-mT_s)\right.\right. \\ &\qquad \left.\left. + n_S(t)\tilde{s}_{jS}(t-mT_s))dt\right\}^2\right],\end{aligned} \quad \text{(A· 4)}$$

for $j=1,2$. Since $n(t)$ is the equivalent low-pass model of AWGN of power spectrum density $N_0/2$, $n_C(t)$ and $n_S(t)$ are orthogonal and

$$\mathrm{E}[n_C(t)n_C(\lambda)] = \mathrm{E}[n_S(t)n_S(\lambda)] = N_0\delta(t-\lambda). \quad \text{(A· 5)}$$

Considering this, we can reduce (A· 4) as

$$\begin{aligned}\mu_{jj} &= \frac{N_0}{T_s^2}\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} |\tilde{s}_{jC}(t-mT_s)|^2 \\ &\qquad + |\tilde{s}_{jS}(t-mT_s)|^2 dt \\ &= \frac{N_0}{T_s^2}\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} \tilde{s}_j(t-mT_s)\tilde{s}_j^*(t-mT_s)dt,\end{aligned} \quad \text{(A· 6)}$$

and from (9), (10) and (11)

$$\mu_{jj} = \frac{4E_bN_0}{T_s^2}. \quad \text{(A· 7)}$$

Similary the cross covariance of the noise components becomes

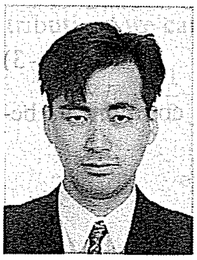
$$\mu_{12} = \mathrm{E}[n_{1,m} \cdot n_{2,m}] = \frac{N_0}{T_s^2}\mathrm{Re}\left[\int_{(m-\frac{1}{2})T_s}^{(m+\frac{1}{2})T_s} [\tilde{s}_1(t-mT_s)\tilde{s}_2(t-mT_s)]dt\right]. \quad (A\cdot 8)$$

Since the result of integration does not depend on m, (A· 8) can be expressed as

$$\mu_{12} = \frac{N_0\gamma}{T_s}, \quad (A\cdot 9)$$

where

$$\gamma=\frac{1}{T_s}\mathrm{Re}\left[\int_{-1/2T_s}^{1/2T_s} \tilde{s}_1(t)\tilde{s}_2^*(t)dt\right]. \quad (A\cdot 10)$$



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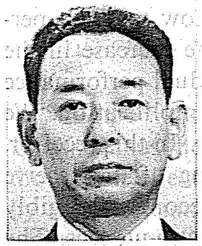

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