論　文

干涉과 雜音이 存在하는 Hard-Limiting 衛星채널上에서의 DS-BPSK 信號의 誤率特性

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Error Rate Performance of DS-BPSK Signal transmitted through a Hard-Limiting Satellite Channel in the presence of Interference and Noise

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요 약　동일채널간섭(cochannel interference)과 다운링크(downlink) 가우스성잡음이 존재하는 환경하에서 비선형 위성 트랜스폰더(transponder)를 통과하는 DS-BPSK(Direct Sequence Binary Phase Shift Keying)신호의 오율식을 구했다. 이때 위성 트랜스폰더의 입력으로는 DS-BPSK 신호와 스펙트럼 확산된 광대역의 동일채널간섭신호와의 합성파를 가정하였다. 구해진 오율식에 의한 계산결과는 반송파 대 간섭파 전력 비(CIR), 다운링크 신호 대 잡음 전력 비(downlink SNR) 그리고 처리이득(process gain)을 파라미터로 하여 그래프로 나타내고 분석했다. 그 결과, DS-BPSK 신호와 간섭신호가 하드 리미터(hard limiter)특성의 트랜스폰더를 통과하게 되면 수신기의 복조단에서는 처리이득을 증가시키더라도 개선되어 지지 않는 협대역의 상호 변조적 성분이 생긴다는 것을 알 수 있었다. 오율면에서는 CIR이 낮을 경우(약 10dB이하)에는 CIR의 증가에 따른 오율 개선도가 현격하지만 약 20dB이상의 경우에서는 별다른 개선효과가 없었다. 또한 처리이득의 경우는 일정한 오율값에 대해 처리이득을 약 10배 증가시키므로써 약 10dB 정도의 다운링크SNR 개선을 얻을 수 있었다.

ABSTRACT　The error rate equation of DS-BPSK (Direct Sequence Binary Phase Shift Keying) signal transmitted through the nonlinear satellite transponder has been derived in the cochannel interference and downlink Gaussian noise environment. The input to the satellite transponder is the superposition of DS-BPSK signal with one interferer which is a cochannel wide-band PN signal. The error rate performance of DS-BPSK system has been evaluated and shown in figures in terms of carrier to interference power ratio(CIR), downlink signal to noise power ratio(downlink SNR) and process gain. In the analysis, it has been shown that the use of a hard limiter in DS-BPSK satellite system leads to the generation of narrow-band intermodulation products which is independent of the process gain. Also it is known that the error rate performance can be improved in the low levels (below 10 dB) of CIR as the CIR increases. As the process gain varies from 10 to 100 the curve gives the about 10 dB gain in downlink SNR to maintain a fixed error rate.

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1. INTRODUCTION

The growth of satellite communications capacity and capability has been revolutionary, a result of the flexibility provided by multiple-access, global coverage digital satellite systems. The unique ability of telecommunication satellites has opened a new era for regional and global communications. However, there are some problems in realizing satellite systems, that is, 1) Interferences between channels - cochannel interference [1] and adjacent channel interference,[2] 2) Jamming signals, 3) Uplink and downlink noises,[3] 4) Intermodulation products,[4] etc. The performance of satellite systems may be degraded seriously by those factors. Also there are several features in satellite communications normally not encounted in terrestrial communications. These include the nonlinear property of power limited amplifier,[5] and easy exposure to any radio frequency interference.

Pseudonoise(PN) spread spectrum techniques have been widely studied in satellite communications because of certain features, that is, multiple access capability, and lower susceptibility to radio frequency interference and signal hiding.[6]

This paper presents the results of the performance analysis of a direct sequence(DS) spread spectrum binary-phase-shift-keying(BPSK) system designed to operate through nonlinear satellite communication systems in the presence of interference and downlink noise. The performance of the DS(Direct Sequence) signal transmitted through the nonlinear satellite channel in the environment of CW interference and noise has been investigated by T.C. Huang,[3] etc. For spread-spectrum signals, however, the classical "transform method"[10] cannot give a complete analysis because it does not supply phase information about the output signals.[9] Consequently, the transform method must be modified to enable a useful analysis of PN spread-spectrum systems. Thus we analyze the error rate performance of DS-BPSK signal by using analysis method in

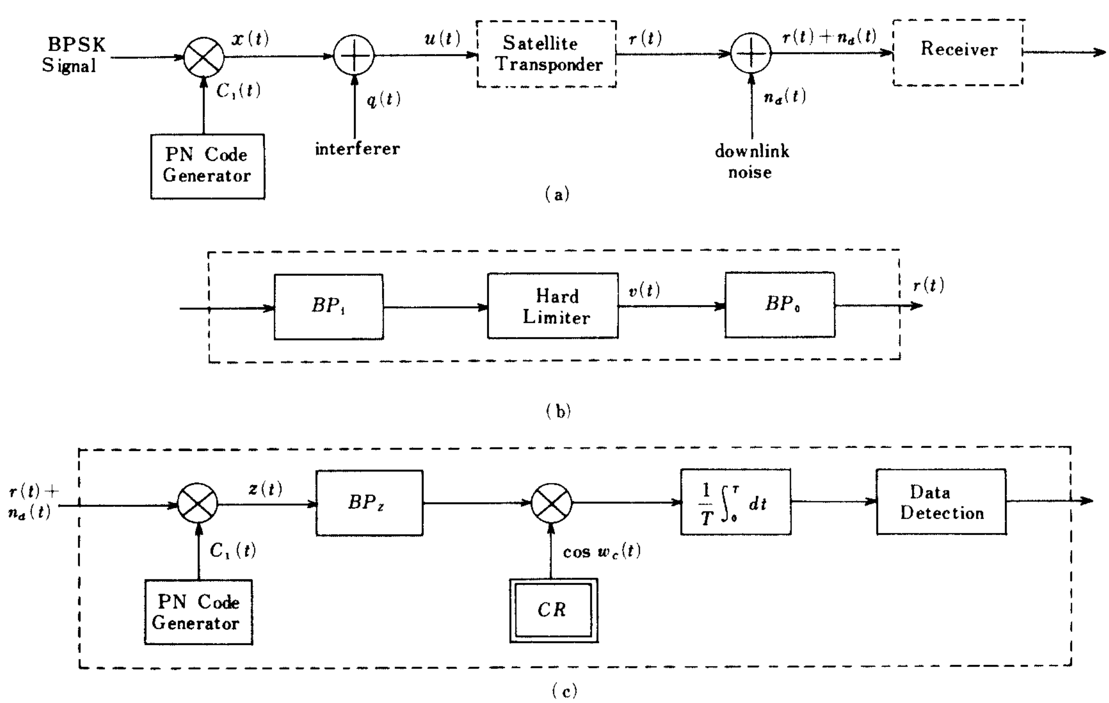

Fig. 1 Analysis model (a) overall configuration, (b) satellite transponder, (c) receiver.

Ref. (9) in this paper. We assumed here that the interference signal is a cochannel wide band PN signal, for example, the signal of another subscriber in a spread spectrum multiple access(SSMA) system.

2. ANALYSIS MODEL

Fig. 1 shows the block diagram of the analysis model. An overall configuration is given in Fig. 1 (a). In a satellite transponder given in Fig. 1 (b), it consists of the input bandpass filter BP_i, the hard limiter, and the output bandpass filter BP_O. The bandpass filters are assumed to be ideal filters with the bandwidth equal to the bandwidth of the DS-BPSK signal $x(t)$, and the input to the satellite transponder consists of the superposition of the DS-BPSK signal $x(t)$ with one interferer $q(t)$ which is a cocnannel wide-band PN signal.

2.1 DS-BPSK signal: $x(t)$

DS-BPSK signal $x(t)$ may be represented as

$$x(t) = AC_1(t)\cos\{w_c t + \phi_x\} \tag{1}$$

where A : data,
$C_1(t)$: PN code signal,
w_c : carrier frequency,
ϕ_x : phase.

2.2 Interference signal: $q(t)$

The interference signal $q(t)$ is assumed to be modulated with amplitude B and carrier frequency w_c which is the same as carrier frequency of DS-BPSK signal, and also assumed to be a wide-band signal with its own PN code $C_2(t)$ and thus a PN spread signal, for example, the signal of another subscriber in a spread-spectrum-multiple-access(SSMA) system. Then it may be expressed as

$$q(t) = B\, C_2(t)\cos\{w_c t + \phi_q\} \tag{2}$$

where B : data,
$C_2(t)$: PN code signal,
ϕ_q : phase.

Here we now make a major simplifying assumption, namely that the PN codes $C_1(t)$ and $C_2(t)$ have the same chip rate and the phases ϕ_x and ϕ_q are constant. Actually the phase ϕ_x and ϕ_q are low frequency signals compared to the chip rate of the PN sequences and the carrier frequency f_c, and thus the effect of the phases can be neglected in the analysis of this paper (especially in the analysis of a PN spread spectrum system).

2.3 Hard - Limiter

2.3.1 Input signal: $u(t)$

The input to the satellite transponder is the superposition of the DS-BPSK signal $x(t)$ with the interferer $q(t)$, thus

$$u(t) = x(t) + q(t) = A\, C_1(t)\cos w_c t + B\, C_2(t)\cos w_c t. \tag{3}$$

2.3.2 Output signal: $v(t)$

The output v as a function of the input u of a hard limiter with a transfer characteristic $g(u)$ can be represented by means of the inverse Fourier transform of the transfer function $G(jw)=2/jw$ and thus the output of the hard limiter can be written as (See Appendix-1)

$$\begin{aligned} v(t) &= \frac{1}{2\pi}\int_{-\infty}^{\infty} G(jw) \exp\{jw(AC_1(t)\cos w_c t + BC_2(t)\cos w_c t)\}\,dw \\ &= \frac{1}{j\pi}\int_{-\infty}^{\infty} \frac{1}{w}\{J_o(Aw)\,J_o(Bw) \\ &+ 2J_o(Bw)\sum_{i=1}^{\infty}(j)^i C_1^i(t)\,J_i(Aw)\cos iw_c t \\ &+ 2J_o(Aw)\sum_{k=1}^{\infty}(j)^k C_2^k(t)\,J_k(Bw)\cos kw_c t \\ &+ 4\sum_{i=1}^{\infty}\sum_{k=1}^{\infty}(j)^{i+k} C_1^i(t)\,C_2^k(t)\,J_i(Aw) \\ &\cdot J_k(Bw)\cos iw_c t\ \cos kw_c t\}\,dw \end{aligned} \tag{4}$$

where $J_i(Aw)$ denotes the bessel function of the first kind, order i and argument Aw.

The bandpass filter BP_0 removes higher harmonics from $v(t)$ to produce $r(t)$.

$$r(t)=\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{1}{w}\{2J_0(Bw)\,C_1(t)\,J_1(Aw)\cos w_ct$$
$$+2J_0(Aw)\,C_2(t)\,J_1(Bw)\cos w_ct$$
$$+2\sum_{i=1}^{\infty}(-1)^i C_1^{i+1}(t)\,C_2^i(t)\,J_{i+1}(Aw)\cdot J_i(Bw)\cos w_ct$$
$$+2\sum_{k=1}^{\infty}(-1)^k C_1^k(t)\,C_2^{k+1}(t)\cdot J_k(Aw)\,J_{k+1}(Bw)\cos w_ct\}\,dw. \quad (5)$$

2.4 Receiver

2.4.1 Input signal: $z(t)$

The despread signal $z(t)$ in Fig. 1 (c) can be represented as

$$z(t)=r(t)\,C_1(t)+n_d(t)\,C_1(t)$$
$$=\frac{2}{\pi}\int_{-\infty}^{\infty}\frac{1}{w}J_0(Bw)\,J_1(Aw)\,dw\cos w_ct$$
$$+\frac{2}{\pi}\int_{-\infty}^{\infty}\frac{1}{w}J_0(Aw)\,J_1(Bw)\,dw\cdot C_1(t)\,C_2(t)\cos w_ct$$
$$+\frac{2}{\pi}\sum_{i=1}^{\infty}(-1)^i C_1^i(t)\,C_2^i(t)\cos w_ct\cdot\int_{-\infty}^{\infty}\frac{1}{w}J_{i+1}(Aw)\,J_i(Bw)\,dw$$
$$+\frac{2}{\pi}\sum_{k=1}^{\infty}(-1)^k C_1^{k+1}(t)\,C_2^{k+1}(t)\cos w_ct\cdot\int_{-\infty}^{\infty}\frac{1}{w}J_k(Aw)\,J_{k+1}(Bw)\,dw$$
$$+n_d(t)\,C_1(t) \quad (6)$$

where $n_d(t)$ denotes a downlink Gaussian noise.

The representation of $z(t)$ according to (6) enables the identification of each term, that is, information signal term, interfering term, intermodulation products term, and downlink noise term.

(a) Information signal term.

$$z_x(t)=A_x\cos w_ct \quad (7)$$

where
$$A_x=\frac{4}{\pi}\int_0^{\infty}\frac{1}{w}J_0(Bw)\,J_1(Aw)\,dw$$
$$=-\sum_{n=0}^{\infty}\frac{4}{\pi}\frac{(2n-1)\,[(2n-3)!!]^2}{2^{2n}\,(n!)^2}\left(\frac{B^2}{A^2}\right)^n.$$

(B^2/A^2) denotes interference power to carrier power ratio.

(b) Interference signal term

$$z_q(t)=A_q\,C_1(t)\,C_2(t)\cos w_ct \quad (8)$$

where
$$A_q=\frac{4}{\pi}\int_0^{\infty}\frac{1}{w}J_0(Aw)\,J_1(Bw)\,dw$$
$$=\sum_{n=0}^{\infty}\frac{2}{\pi}\frac{B}{A}\frac{[(2n-1)!!]^2}{2^{2n}(n+1)\,(n!)^2}\left(\frac{B^2}{A^2}\right)^n$$

(c) Intermodulation product term (See Appendix – II)

$$z_{xq}(t)=\sum_{i:even}^{\infty}A_{xqi}\cos w_ct+\sum_{k:odd}^{\infty}A_{xqk}\cos w_ct$$
$$+\sum_{i:odd}^{\infty}A_{xqi}\,C_1(t)\,C_2(t)\cos w_ct$$
$$+\sum_{k:even}^{\infty}A_{xqk}\,C_1(t)\,C_2(t)\cos w_ct \quad (9)$$

where
$$A_{xqi}=(-1)^i\frac{4}{\pi}\int_0^{\infty}\frac{1}{w}J_{i+1}(Aw)\cdot J_i(Bw)\,dw$$
$$=(-1)^i\frac{2}{\pi}\left(\frac{B}{A}\right)^i\frac{1}{2^{i-1}}\sum_{n=0}^{\infty}\frac{(2n+2i-1)!!\,(2n-3)!!}{(-2^{2n})\,(n+i)!\,n!}\left(\frac{B^2}{A^2}\right)^n,$$

$$A_{xqk}=(-1)^k\frac{4}{\pi}\int_0^{\infty}\frac{1}{w}J_k(Aw)\cdot J_{k+1}(Bw)\,dw$$
$$=(-1)^k\frac{2}{\pi}\left(\frac{B}{A}\right)^{k+1}$$

$$\sum_{n=0}^{\infty} \frac{(2n+2k-1)!!\ (2n-1)!!}{2^{2n+k}\, n!\ (n+k+1)!} \cdot \left(\frac{B^2}{A^2}\right)^n .$$

(d) Downlink noise term

$$n_d(t) = N\cos(w_c t + \xi) \tag{10}$$

where $\text{pdf}(\xi) = \frac{1}{2\pi}$.

$$z_{nd}(t) = n_d(t)\,C_1(t) = NC_1(t)\cos(w_c t + \xi) = N'\cos(w_c t + \xi) \tag{11}$$

where the variance of z_{nd}(t) can be represented as

$$\sigma_{nd}^2 = \sigma^2 / P \tag{12}$$

σ^2 : variance of downlink white Gaussian noise

P : process gain of PN code $C_1(t)$

$\left(= \frac{\text{Bandwidth of } BP_1(\text{ or } BP_o)}{\text{Bandwidth of } BP_Z}\right)$.

The p.d.f (probability density function) of z_{nd}'(t) is given by

$$\text{pdf}[z_{nd}(t)] = \frac{1}{\sqrt{2\pi}\,(\sigma/\sqrt{P})} \cdot \exp(-x^2/2(\sigma^2/P)). \tag{13}$$

The input signal of the receiver thus can be written as

$$z(t) = z_x(t) + z_q(t) + z_{xq}(t) + z_{nd}(t). \tag{14}$$

3. PROBABILITY OF ERROR

Input signal of the level comparator (output of Integrated & Dump filter) may be writen as

$$z_k = \frac{1}{2}A_x + A'_q + R_1 + R_2 + A'_{xqi} + A'_{xqk} + n_G \tag{15}$$

where

$$A'_q = \frac{1}{T}\int_0^T A_q C_1(t)\, C_2(t) \cos^2 w_c t\, dt,$$

$$A'_{xqi} = \sum_{i:odd}^{\infty} \frac{1}{T}\int_0^T A_{xqi} C_1(t)\, C_2(t) \cdot \cos^2 w_c t\, dt,$$

$$A'_{xqk} = \sum_{k:even}^{\infty} \frac{1}{T}\int_0^T A_{xqk} C_1(t)\, C_2(t) \cdot \cos^2 w_c t\, dt,$$

$$R_1 = \frac{1}{2}\sum_{i:even}^{\infty} A_{xqi}, \qquad R_2 = \frac{1}{2}\sum_{k:odd}^{\infty} A_{xqk},$$

$$n_G = \frac{N'}{2}\cos\xi \text{ (Gaussian random variable).}$$

In the analysis of spread spectrum system, when the process gain is high the effect of the interference containing the term of a multiplication of C_1(t) and C_2(t) can be neglected. The information data stream assumed in this paper has only two possible states; +1 and −1, and in most practical digital transmission systems, data-stream randomizers assure that the transmitted signals have equal probabilities.[8] Thus we can find the probability of error by considering only one case of +1 or -1 for equiprobable randomized data streams. In this analysis, we consider the case of +1.

Then the probability of error is

$$P_e = \text{prob.}\left[\frac{1}{2}A_x + R_1 + R_2 + n_G < 0\right], \tag{16}$$

and p.d.f. can be represented as

$$p(n_G) = \frac{1}{\sqrt{2\pi}\,\sigma_{nd}} \exp(-n_G^2/2\sigma_{nd}) \tag{17}$$

where σ_{nd}^2 : power of $z_{nd}(t)$.

Finally, the probability of detection error is given by

$$P_e = \int_{-\infty}^{-\frac{A_x}{2}-R_1-R_2} \frac{1}{\sqrt{2\pi}\,\sigma_{nd}} \exp\left(-n_G^2 / 2\sigma_{nd}^2\right) dn_G$$

$$= \frac{1}{2}\,\mathrm{erfc}\left[\frac{1}{\sqrt{2}\,\sigma_{nd}}\left(\frac{1}{2}A_x + R_1 + R_2\right)\right] \qquad (18)$$

where $\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} \exp(-t^2)\, dt$.

4. NUMERICAL RESULTS AND DISCUSSIONS

The results derived in the previous section have been used to evaluate the performance of the system. Numerical results are presented in Fig. 2 - Fig. 7 in terms of CIR (carrier to interference power ratio), downlink SNR (downlink signal to noise power ratio) and the process gain P to illustrate the effects of a hard limiter. The results here are similar to that of Ref. (3), but we should notice that the value of downlink SNR in Ref. (3) is two times larger than that of this paper.

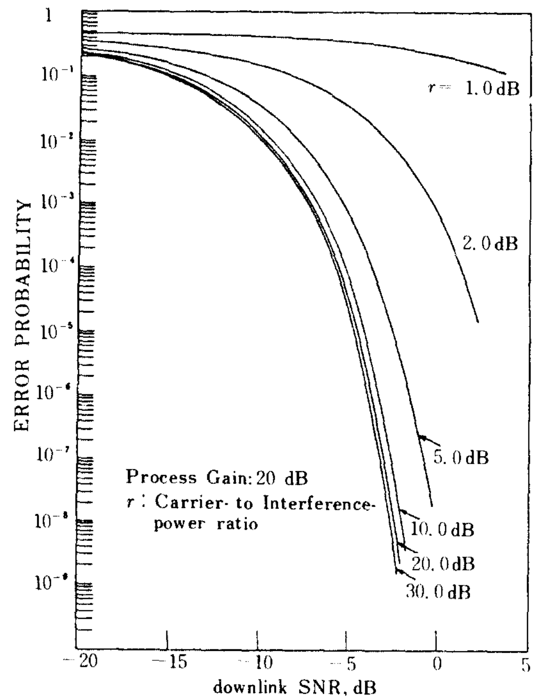

Fig. 3 Error rate performance of DS-BPSK signal with 20 dB of the process gain.

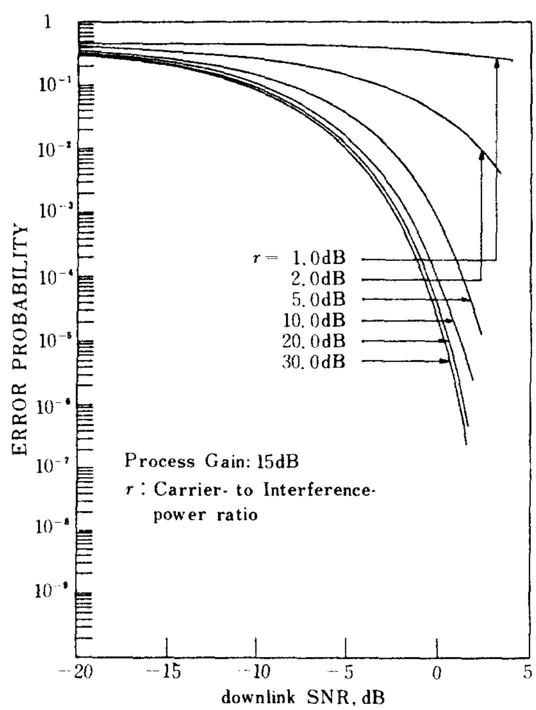

Fig. 2 Error rate performance of DS-BPSK signal with 15 dB of the process gain.

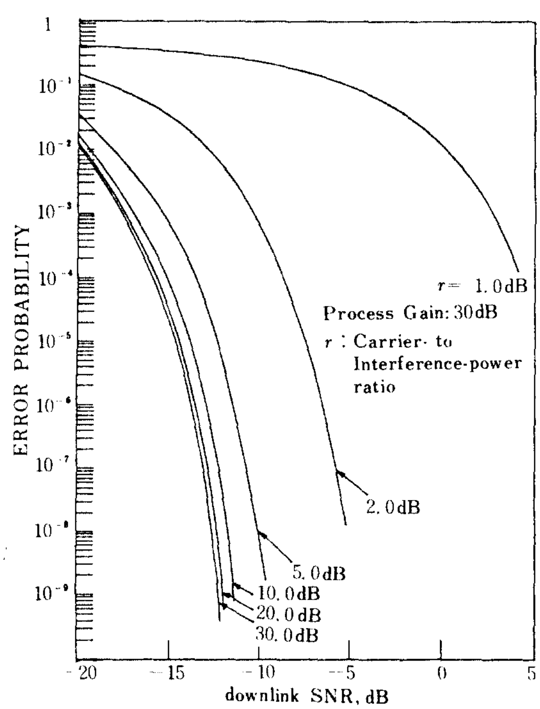

Fig. 4 Error rate performance of DS-BPSK signal with 30 dB of the process gain.

Thus in comparison of downlink SNR's between two papers, the value of SNR's in Ref. (3) is 3dB-smaller than the SNR's in this paper.

From the numerical results, we have known that;

(1) Fig. 2 shows the effect on the error rate performance of DS-BPSK signal produced by the hard limiter in satellite transponder for a fixed process gain of 15dB with the parameter of CIR and the function of downlink SNR. It can be seen from Fig. 2 that when the CIR is less than 10dB the error rate performance is improved remarkably, however, when the CIR is above 30dB the performance cannot be improved any more by increasing the values of CIR.

(2) For a fixed process gain of 20dB, the improvements on the error rate performance is presented in Fig. 3 with the parameter of CIR and the function of downlink SNR. In this figure, we have also known that as the CIR increases the amount of performance improvements becomes less and less.

(3) In Fig. 4, the value of a fixed process gain is 30dB. Since the value of a fixed process gain is high, it shows that the amount of performance improvement between r = 1.0dB and 2.0dB, r = 2.0dB and 5.0dB is very large compared with Fig. 2 and Fig. 3.

(4) Fig. 5 is plotted to show the effect of process gain in combating interference for a fixed CIR of 5dB as the parameter of process gain and the function of downlink SNR. As the process gain varies from 10 to 100 the curve gives the about 10dB gain in downlink SNR to maintain a fixed error probability $P_e = 10^{-3}$.

(5) Fig. 6 is the performance curve with a fixed CIR of 10dB with process gain as a parameter and downlink SNR as a function.

(6) Also the effect of process gain in combating interference is illustrated in Fig. 7 for a fixed CIR of 20dB.

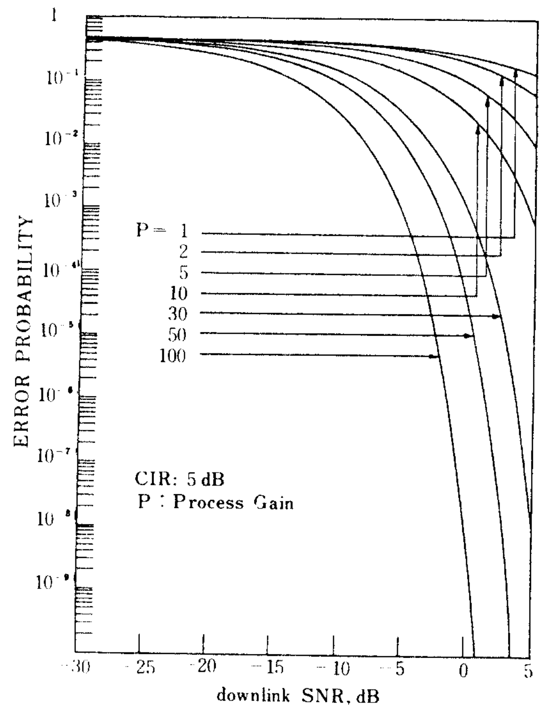

Fig. 5 Effects of process gains in DS-BPSK satellite system(CIR = 5 dB).

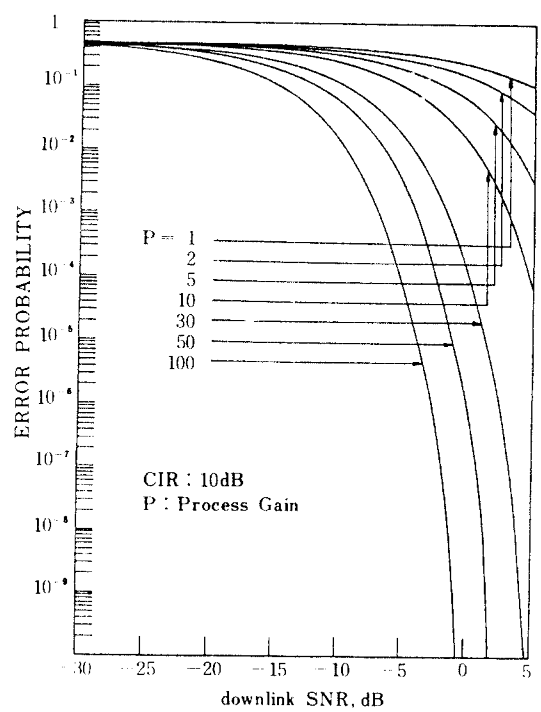

Fig. 6 Effects of process gains in DS-BPSK staellite system(CIR = 10dB).

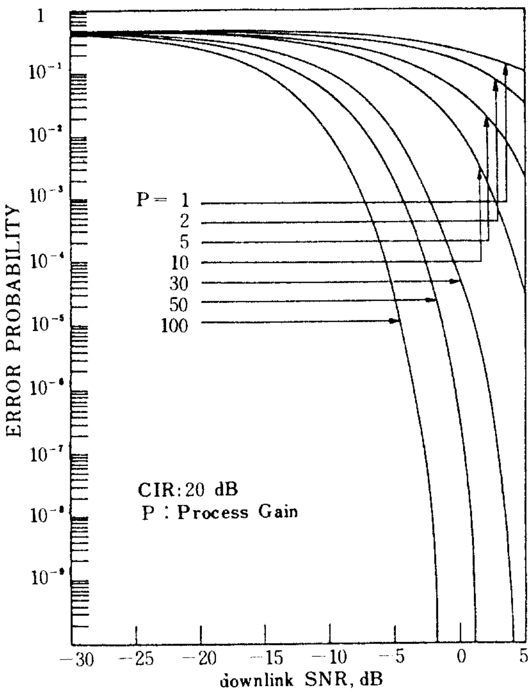

Fig. 7 Effects of process gains in DS-BPSK satellite system (CIR=20dB).

5. CONCLUSIONS

This paper presented the error rate performance of the DS-BPSK signal transmitted over the nonlinear satellite transponder in the presence of cochannel interference and downlink Gaussian noise. We have known that a hard limiter in the satellite transponder leads to the generation of strong narrow band intermodulation products at the output of a receiver. Since the power of this intermodulation product is independent of the process gain, in the worst case, the antijamming capability may be lost completely even with large "process gain" when the power of intermodulation products is equal to that of the information signal.

Our numerical results have been computed and shown that the error probability becomes less according to the increase of the process gain and CIR. Especially the error rate performance can be improved as the process gain increases when the CIR is less than 20dB.

APPENDIX

I. The exponential functions may be written

$$\exp\{jwAC_1(t)\cos w_ct\} = \cos\{Aw\,C_1(t)\cos w_ct\} + j\sin\{AwC_1(t)\cos w_ct\}. \qquad (\text{I}-1)$$

Since the PN code signal $C_1(t)$ attains only the values +1 or -1, (I-1) simplifies to

$$\exp\{jw\,AC_1(t)\cos w_c t\} = \cos\{Aw\cos w_ct\} + j\,C_1(t)\sin\{Aw\cos w_ct\}. \qquad (\text{I}-2)$$

And the cosine and sine terms in (I-2) may be written

$$\cos\{Aw\cos w_ct\} = J_0(Aw) + 2\sum_{k=1}^{\infty}(-1)^k \cdot J_{2k}(Aw)\cos 2kw_ct,$$

$$\sin\{Aw\cos w_ct\} = 2\sum_{k=1}^{\infty}(-1)^k J_{2k+1}(Aw)\cos(2k+1)w_ct. \qquad (\text{I}-3)$$

The exponential function in (I-2) thus may be written as

$$\exp\{jw\,AC_1(t)\cos w_ct\} = J_0(Aw) + 2\sum_{k=1}^{\infty}(j)^k C_1^k(t) \cdot J_k(Aw)\cos k\,w_ct. \qquad (\text{I}-4)$$

II. In the expression of (6), it has been assumed that the signals $C_1(t)$ and $C_2(t)$ attain only the values +1 and -1. The result of the product $C_1(t)\cdot C_2(t)$ raised to an even power thus becomes 1. Consequently, the terms in the last two summations of (6) do not depend upon $C_1(t)$ or

$C_2(t)$ when i is even and k is odd. Thus the last two terms of (6) may be represented as

$$
\begin{aligned}
z_{xq}(t) &= \frac{2}{\pi}\sum_{i=1}^{\infty}(-1)^i\, C_1^i(t)\, C_2^i(t)\cos w_c t \\
&\quad \cdot \int_{-\infty}^{\infty}\frac{1}{w} J_{i+1}(Aw)\, J_i(Bw)\, dw \\
&+ \frac{2}{\pi}\sum_{k=1}^{\infty}(-1)^k\, C_1^{k+1}(t)\, C_2^{k+1}(t)\cos w_c t \\
&\quad \cdot \int_{-\infty}^{\infty}\frac{1}{w} J_k(Aw)\, J_{k+1}(Bw)\, dw \\
&= \sum_{i=1}^{\infty} A_{xqi}\, C_1^i(t)\, C_2^i(t)\cos w_c t \\
&\quad + \sum_{k=1}^{\infty} A_{xqk}\, C_1^{k+1}(t)\, C_2^{k+1}(t)\cos w_c t \\
&= \sum_{i:\,even}^{\infty} A_{xqi}\cos w_c t + \sum_{k:\,odd}^{\infty} A_{xqk}\cos w_c t \\
&\quad + \sum_{i:\,odd}^{\infty} A_{xqi}\, C_1(t)\, C_2(t)\cos w_c t \\
&\quad + \sum_{k:\,even}^{\infty} A_{xqk}\, C_1(t)\, C_2(t)\cos w_c t .
\end{aligned}
$$

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