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Coding for Spread Spectrum Packet Radios

Jim K. Omura

October 15, 1980

National Aeronautics and
Space Administration

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PREFACE

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ABSTRACT

Packet radios are often expected to operate in a radio communication network environment where there tends to be man-made interference signals. To combat such interference spread spectrum waveforms are being considered for some applications [1]. In this report we examine the use of convolutional coding with Viterbi decoding to further improve the performance of spread spectrum packet radios. At 10^{-5} bit error rates improvements in performance of 4 dB to 5 dB can easily be achieved with such coding without any change in data rate nor spread spectrum bandwidth. This coding gain is more dramatic in an interference environment.

I. INTRODUCTION

We derive expressions for the bit error probability, P_b, as a function of energy-per-bit to noise ratio, E_b/N_o, and interference-to-signal power ratio, I/S. Here interference can be due to multipath, intersymbol interference, and other man-made signals. The basic modulation we consider is QPSK where the inphase and quadrature bits consist of orthogonal bit sequences of length N. For our examples we pick N = 16 or 32*. The modulation system is shown in Fig. 1. Here we also show the use of convolutional codes with constraint length K = 4 for $N \geq 16$ and K = 5 for $N \geq 32$.

In this analysis we assume that any interference signal of power I will appear as a Gaussian noise term after passing through the matched filters. Thus when N_o is the single sided noise spectral density when there is additive white Gaussian noise alone, then with interference the new equivalent noise spectral density is

$$\hat{N}_o = N_o + \frac{2IT_b}{N} \tag{1}$$

where IT_b is the interference signal energy during a bit time T_b. This is used to obtain an equivalent energy-per-bit to noise ratio

$$\frac{E_b}{\hat{N}_o} = \frac{E_b}{N_o + \frac{2IT_b}{N}}$$

$$= \frac{\frac{E_b}{N_o}}{1 + \frac{2}{N}\left(\frac{ST_b}{N_o}\right)\left(\frac{I}{S}\right)}$$

*This is based on Collin's packet radio [1].

$$= \frac{\frac{E_b}{N_o}}{1 + \frac{2}{N}\left(\frac{E_b}{N_o}\right)\left(\frac{I}{S}\right)} \qquad (2)$$

The Gaussian assumption is based on the fact that the matched filter essentially provides N samples of the interference using binary, {-1,1}, weighting which is approximated as a Gaussian random variable when N is moderate in size. Thus the channel is assumed to be an additive white Gaussian noise channel of spectral density N_o with no interference and when we want to include interference we use the signal to noise ratio,

$$\frac{E_b}{\hat{N}_o} = \frac{\frac{E_b}{N_o}}{1 + \alpha \frac{E_b}{N_o}} \quad , \quad \alpha = \frac{2}{N}\left(\frac{I}{S}\right) \qquad (3)$$

In this report we first examine coherent receivers both with and without convolutional coding [2]. This is followed by an examination of non-coherent receivers using basically the same modulation/coding transmitters. We shall refer to inphase and quadrature orthogonal binary sequences of length N which will be modulated on QPSK carriers to form the spread spectrum signals as inphase and quadrature chip sequences. These orthogonal chip sequences can be generated using rows of $2^k \times 2^k$ matrices denoted H_k generated in the following recursive manner.

$$H_{k+1} = \begin{bmatrix} H_k & H_k \\ H_k & -H_k \end{bmatrix} \qquad k = 1, 2, \ldots \qquad (4)$$

$$H_1 = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix}$$

The two rows of H_1 form two chip sequences, (1 1) and (1 -1), of length 2 that are orthogonal. Next for k = 2 we have the $2^k = 4$ rows of H_2,

$$H_2 = \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 \end{bmatrix} \tag{5}$$

which are orthogonal binary chip sequences of length $2^k = 4$. In general there are $N = 2^k$ orthogonal binary chip sequences of length $N = 2^k$.

In the following we shall denote, as shown in Fig. 1 and Fig. 2, inphase and quadrature chip sequences as $\underline{C}_I$ and $\underline{C}_Q$ respectively. Each of these N chip sequences can be one of N orthogonal binary sequences as shown above. In practice we may want to find sequences with good partial correlation properties such as Gold codes or BCH codes [3].

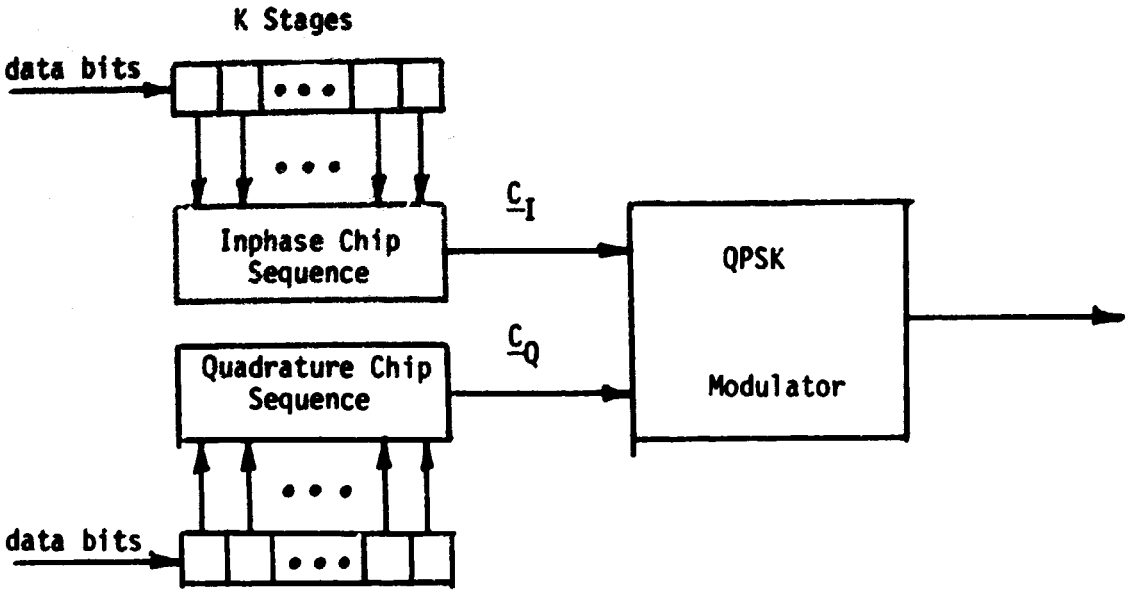


Figure 1. Encoder/Modulator

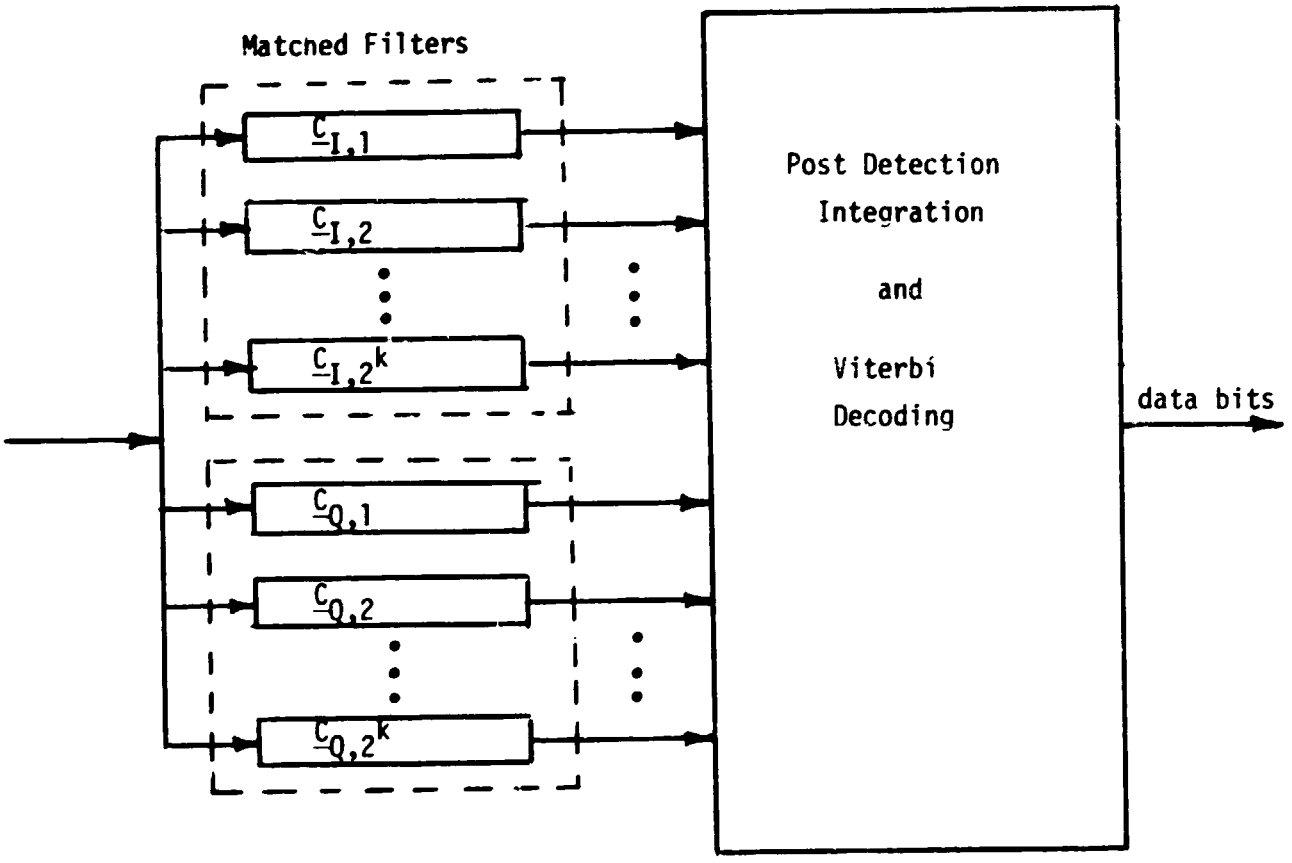


Figure 2. Demodulator/Decoder

II. COHERENT SYSTEM – NO CODING

We first examine a coherent receiver that tracks a reference phase for the transmitted packet radio signal. With no coding consider the inphase and quadrature data bit and chip sequence relations

Inphase		Quadrature	
0	$\rightarrow \underline{C}_I$	0	$\rightarrow \underline{C}_Q$
1	$\rightarrow -\underline{C}_I$	1	$\rightarrow -\underline{C}_Q$

where $\underline{C}_I$ and $\underline{C}_Q$ are any two chip sequences. In a white Gaussian noise channel the bit error probability is simply that of binary antipodal signals [2],

$$P_b = Q\left(\sqrt{\frac{2E_b}{N_o}}\right) < \frac{1}{2} e^{-\frac{E_b}{N_o}} \tag{6}$$

where

$$Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-\frac{t^2}{2}} dt. \tag{7}$$

If multipath causes chip sequences to overlap in time one approach to overcoming this intersymbol interference is to use sequences with small partial correlation properties [3]. Another option is to consider alternating orthogonal chip sequences as follows

	Inphase		Quadrature	
0 0	$\rightarrow \underline{C}_{I,1}$	$\underline{C}_{I,2}$	$\rightarrow \underline{C}_{Q,1}$	$\underline{C}_{Q,2}$
0 1	$\rightarrow \underline{C}_{I,1}$	$-\underline{C}_{I,2}$	$\rightarrow \underline{C}_{Q,1}$	$-\underline{C}_{Q,2}$
1 0	$\rightarrow -\underline{C}_{I,1}$	$\underline{C}_{I,2}$	$\rightarrow -\underline{C}_{Q,1}$	$\underline{C}_{Q,2}$
1 1	$\rightarrow -\underline{C}_{I,1}$	$-\underline{C}_{I,2}$	$\rightarrow -\underline{C}_{Q,1}$	$-\underline{C}_{Q,2}$

where $\underline{C}_{I,1}$ and $\underline{C}_{I,2}$ are orthogonal and $\underline{C}_{Q,1}$ and $\underline{C}_{Q,2}$ are orthogonal with small partial cross-correlation properties. The receiver is assumed to sample the outputs of the matched filters on alternate bit intervals. Thus, multipath interference of one data bit does not add much interference at the sample time of the following chip sequence matched filter output, which corresponds to the next data bit. If the multipath delays are longer than 4 bit time, $4T_b$, then alternating two orthogonal chip sequences can be extended to many orthogonal chip sequences*. Assuming ideal data bit synchronization as well as ideal phase synchronization we achieve the same uncoded bit error probability given by (6) except with interference due to various partial correlation terms of each multipath component.

Figure 3 shows the bit error probability as a function of the energy-per-bit to noise ratio, E_b/N_o. Here we have interference parameterized by

$$\alpha = \frac{1}{N}\left(\frac{I}{S}\right) \tag{8}$$

where I/S is the interference to signal power ratio and N is the number of chips per data bit. Hence for

$$\alpha = .05$$

$$N = 32$$

we have

$$\frac{I}{S} = 1.6 = 2.04 \text{ dB}$$

*$4T_b$ is equal to two symbol times on each of the I and Q channels.

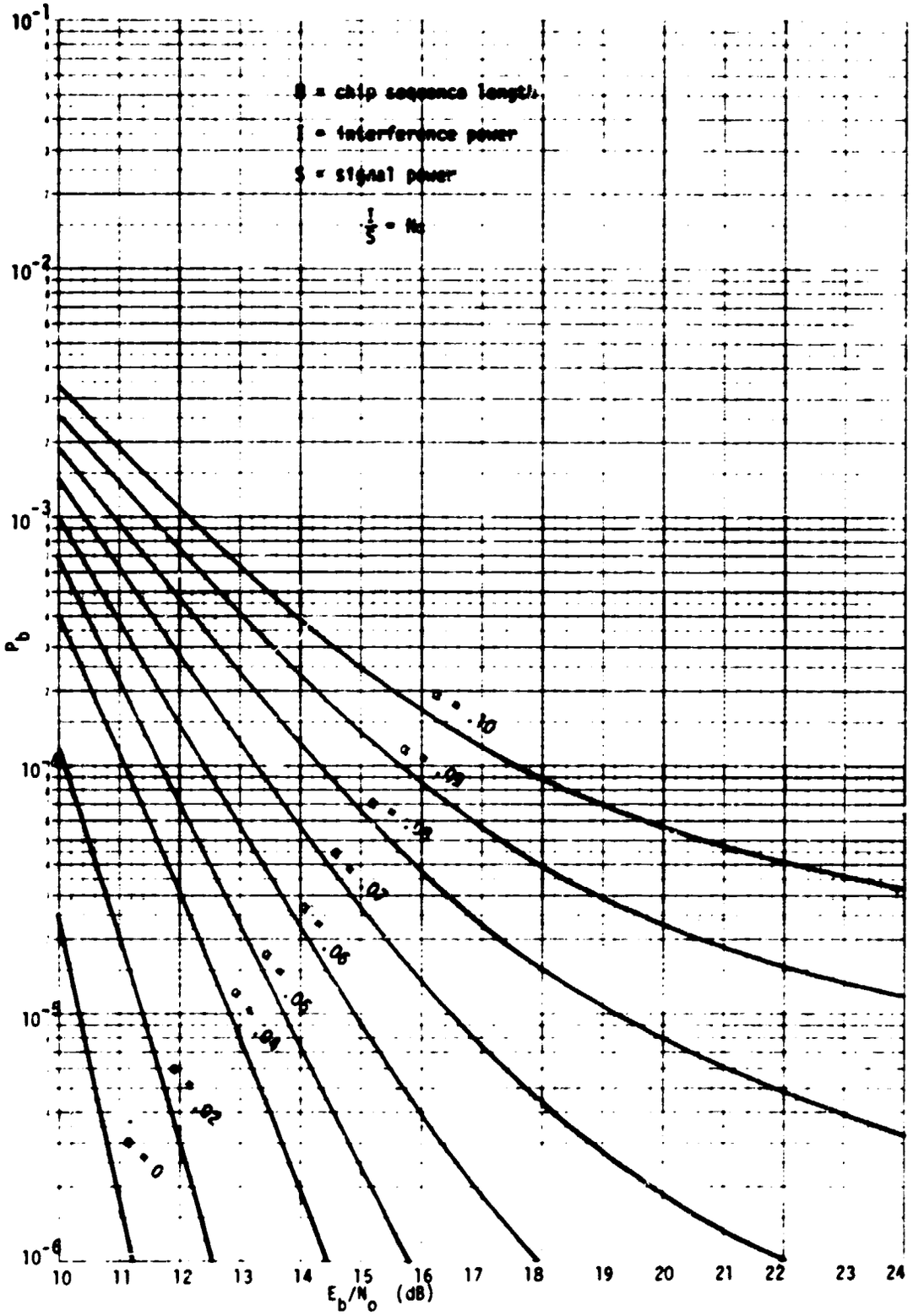


Figure 3. Coherent BPSK, QPSK (Uncoded)

III. COHERENT SYSTEM – CODING

We illustrate the use of convolutional coding for the case where the constraint length is K = 5. Also we treat only the inphase data bit and the inphase chip sequence since the two components of the QPSK modulation can be considered as separate channels.

A Simplex* convolutional code of constraint length K = 5 consists of a 5 bit shift register where the 5 bits in the shift register are used to select one of 32 possible binary sequences of 32 bits length. Sixteen of these sequences are orthogonal to each other while the other 16 are sign reversals of the first 16 sequences. This is shown in Fig. 4.

Consider a K = 5 constraint length convolutional code of rate

$$r = \frac{1}{32} = 2^{-5}$$

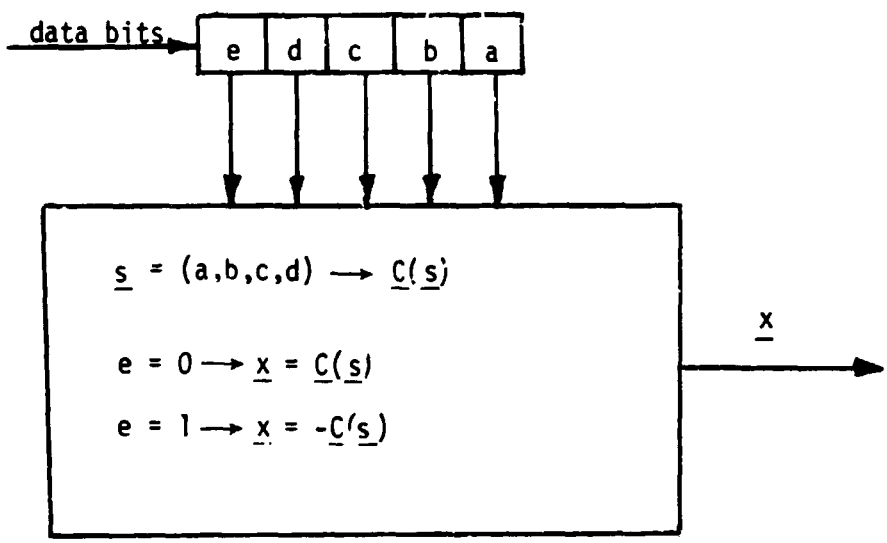


Figure 4. Simplex Convolutional Encoder

*These convolutional codes are analogous to the Simplex block codes [2] since any two diverging and later remerging sequences have cross-correlation that is almost zero and negative. These codes were independently discovered by James Massey [4].

For each data bit that shifts in we generate a 32 bit sequence denoted $\underline{x}$ as follows:

(a) Use data bits $\underline{s} = (a,b,c,d)$ in the register to pick one of 16 orthogonal binary sequences of length 32. Since we have at least 32 such sequences any subset of 16 will serve our purpose. Denote this 32 bit binary sequence as $\underline{C}(\underline{s})$. Hence each of the 16 possible shift register "state" $\underline{s} = (a,b,c,d)$ has a unique orthogonal binary sequence of length 32 associated with it.

(b) The transmitted 32 bit sequence is then given by

$$\underline{x} = \begin{cases} \underline{C}(\underline{s}) & \text{if} \quad e = 0 \\ -\underline{C}(\underline{s}) & \text{if} \quad e = 1 \end{cases} \tag{9}$$

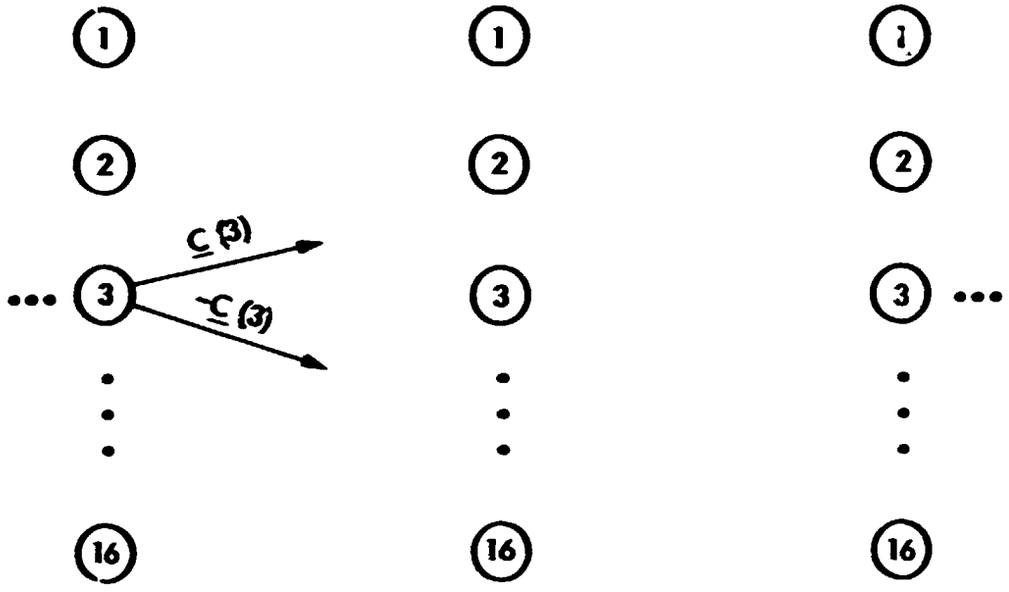
Repeating this procedure each time a data bit enters the shift register results in a 32 bit expansion of the data rate and the desired spread spectrum signal of 15 dB processing gain. Note that this procedure has not forced any change in data rate nor any change in the signal spread bandwidth.

The 16 state trellis diagram has the property that the 32 bit sequences on branches leaving the same state are of opposite sign whereas they are orthogonal to all other 32 bit sequences leaving other states.

Bit Error Bound

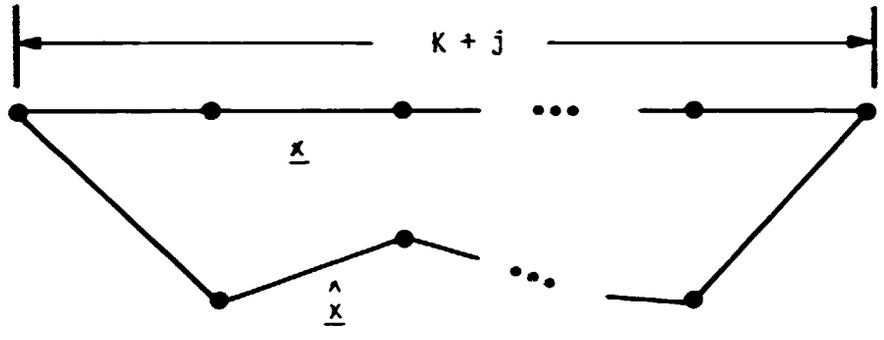
We assume an additive white Gaussian noise channel with spectral density $N_0/2$ (double sided). Each of the bits in the 32 bit code sequence is called a "chip" and is transformed into a BPSK carrier with energy E_c. The energy per data bit is

$$E_b = 32\, E_c \tag{10}$$



TRELLIS DIAGRAM

Consider now the usual coding bound starting with two chip sequences $\underline{x}$ and $\hat{\underline{x}}$ that diverge and remerge over a span of K + j data bits or "branches."



Error Event of Length K + j

The pairwise error probability is

$$P_j = \Pr\{\underline{x} \to \hat{\underline{x}}\} \leq \frac{1}{2} e^{-\frac{\| \underline{x} - \hat{\underline{x}} \|^2}{4N_o}} \tag{11}$$

where

$$\| \underline{x} - \hat{\underline{x}} \|^2 = \| \underline{x} \|^2 + \| \hat{\underline{x}} \|^2 - 2(\underline{x}, \hat{\underline{x}}). \tag{12}$$

For the Simplex convolutional code we get

$$\| \hat{\underline{x}} \|^2 = \| \underline{x} \|^2 = (K + j)\ 32\ E_c$$

$$= (K + j)\ E_b \tag{13}$$

and

$$(\underline{x}, \hat{\underline{x}}) = -32\ E_c = -E_b \tag{14}$$

Hence

$$\| \underline{x} - \hat{\underline{x}} \|^2 = 2(K + j)\ E_b + 2\ E_b$$

$$= 2(K + 1)\ E_b + 2j\ E_b \tag{15}$$

and

$$P_j \leq \frac{1}{2}\ e^{-(K + 1)\frac{E_b}{2N_o}}\ e^{-j\frac{E_b}{2N_o}} \tag{16}$$

There are $M_j \leq 2^j$ possible sequences $\hat{\underline{x}}$ that diverge from $\underline{x}$ during this span of K + j branches (K=5) each causing up to possibly j + 1 bit errors if chosen over $\underline{x}$, the assumed transmitted sequence. Thus the bit error P_b is bounded by

$$P_b \leq \sum_{j=0}^{\infty} (j + 1)\ M_j\ P_j$$

$$\leq \sum_{j=0}^{\infty} (j + 1)\ 2^j \frac{1}{2}\ e^{-(K + 1)\frac{E_b}{2N_o}}\ e^{-j\frac{E_b}{2N_o}}$$

or

$$P_b \leq \frac{\frac{1}{2} e^{-(K+1)\frac{E_b}{2N_o}}}{\left(1 - e^{-\left[\frac{E_b}{2N_o} - \ln 2\right]}\right)^2} \qquad (17)$$

For K = 5 we then have

$$P_b \leq \frac{\frac{1}{2} e^{-3\frac{E_b}{N_o}}}{\left(1 - e^{-\left[\frac{E_b}{2N_o} - \ln 2\right]}\right)^2} \qquad (18)$$

For the uncoded case the bit error is given by (6).

Fig. 5 shows plots of the uncoded error probability and the Simplex convolutional coded error bounds for K = 3,4,5,6. Note that at 10^{-5} bit error probability with constraint length K = 5 the coding gain is approximately 4.5 dB It is almost 3 dB for a simple K = 3 Simplex convolutional code with Viterbi decoding. Fig. 6 and 7 show the K = 4 and K = 5 Simplex convolutional codes' performance with the added impact of interference signals. Comparing with the uncoded case of Fig. 1 we see that the potential coding gain is much greater with interference in the channel. In anti-jamming applications it is known [5] that coding gains against jamming can be much greater than expected from the usual white Gaussian noise channel.

It should be noted that orthogonal convolutional codes have the same performance shown in (17) except with K + 1 replaced by K. Hence the Simplex convolutional code achieves performance equivalent to one constraint

length longer orthogonal convolutional codes. We conjecture that these convolutional codes which are analogous to Simplex block codes are optimum for the additive white Gaussian noise channel. Also note that since only 2^{K-1} orthogonal sequences are required for this convolutional code it can use less bandwidth than the orthogonal convolutional codes. Again this is analogous to the relationship between Simplex and orthogonal block codes.

In general we require the chip length N and constraint length K satisfy

$$N \geq 2^{K-1} \tag{19}$$

As long as this is satisfied we see that for fixed N, there is no reduction in data rate nor change spread spectrum bandwidth to achieve these coding gains.

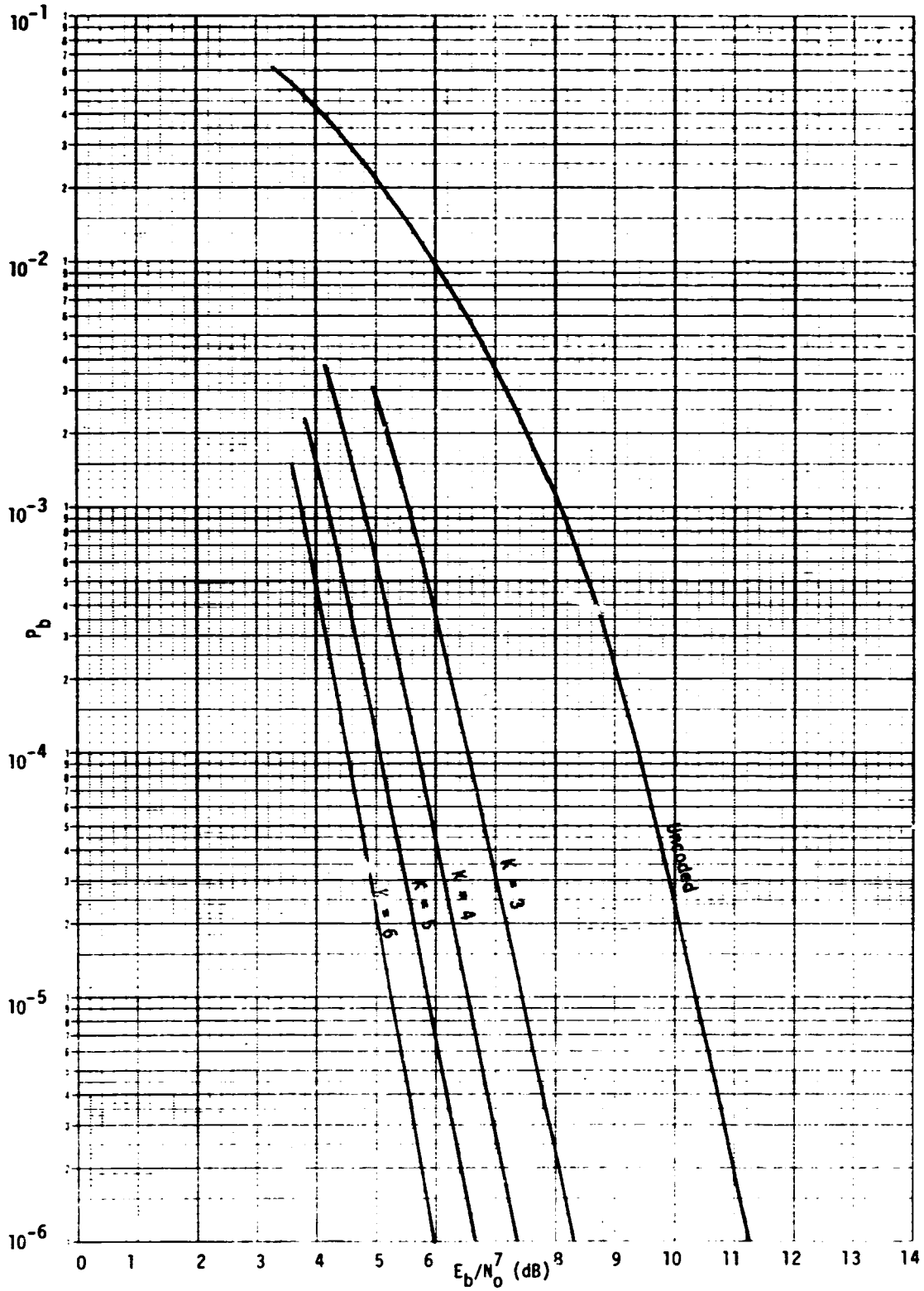


Figure 5. Coherent BPSK, QPSK (I = 0)

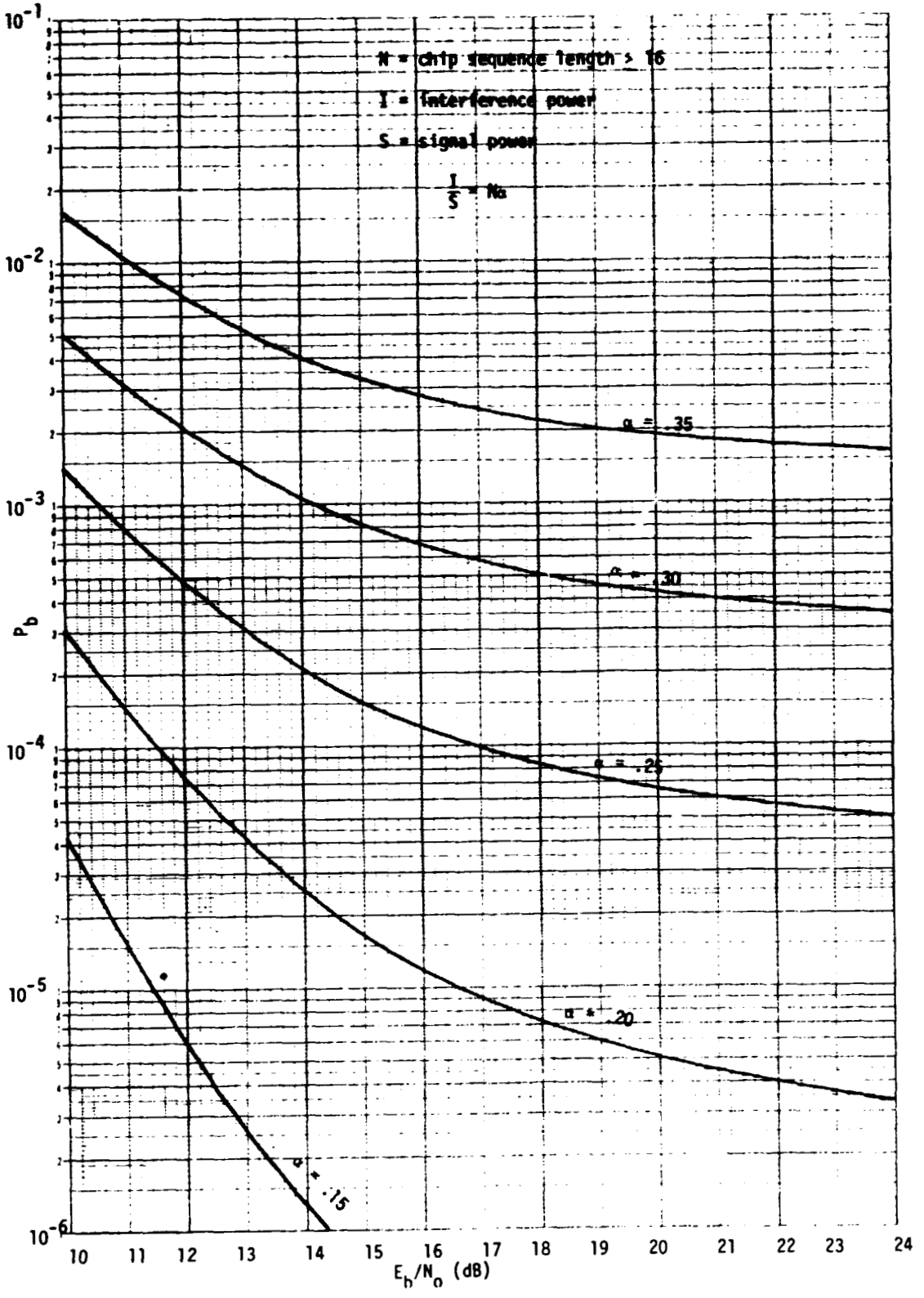


Figure 6. Coherent BPSK, QPSK (Coded K = 4)

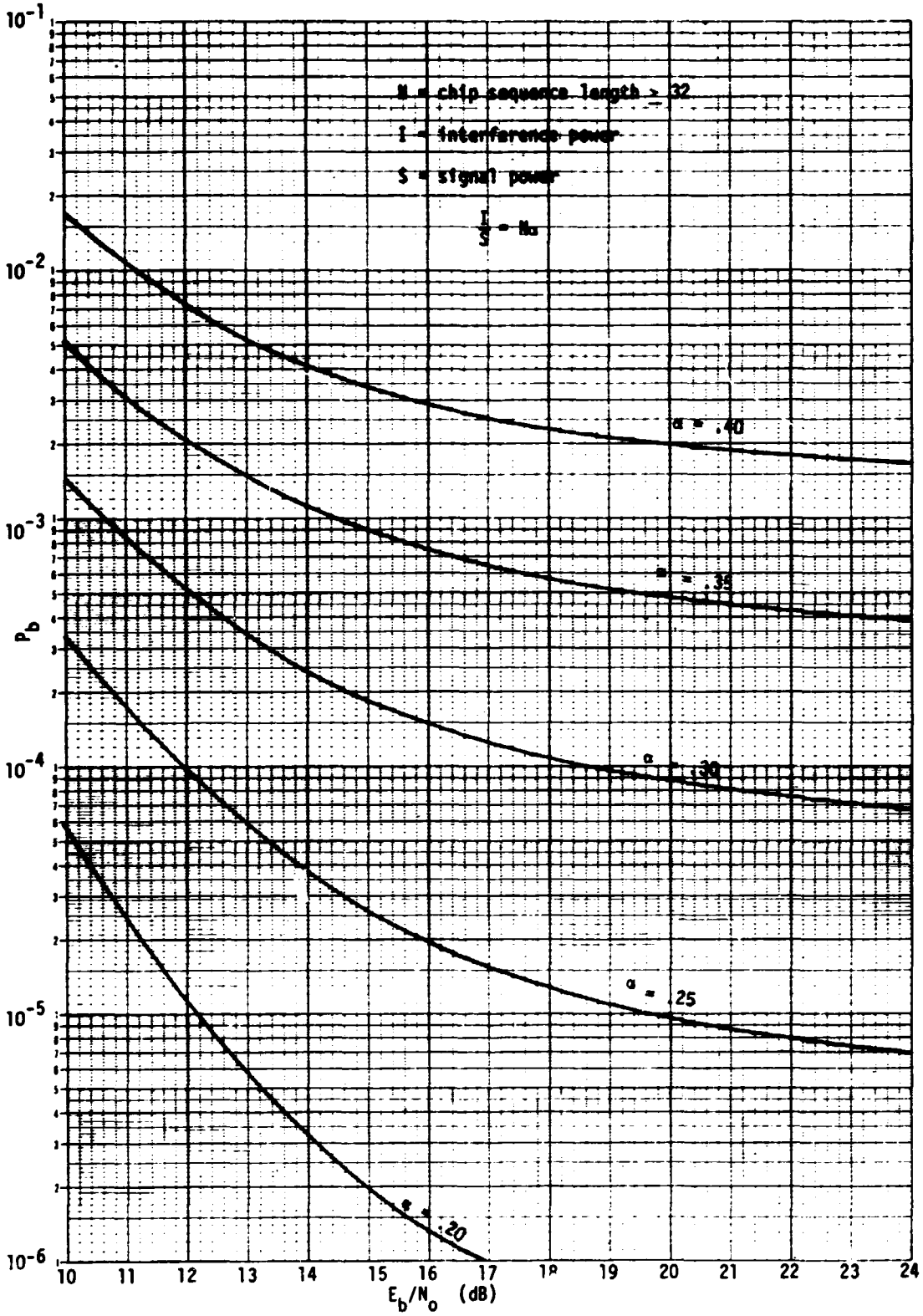


Figure 7. Coherent BPSK, QPSK (Coded K = 5)

IV. NONCOHERENT SYSTEM – NO CODING

When the inphase chip sequence $\underline{C}_I$ and the quadrature chip sequence $\underline{C}_Q$ are orthogonal, noncoherent detection is possible. In the uncoded case we assume the following data bit and chip sequence relation

Inphase	Quadrature
$0 \rightarrow \underline{C}_{I,1}$	$0 \rightarrow \underline{C}_{Q,1}$
$1 \rightarrow \underline{C}_{I,2}$	$1 \rightarrow \underline{C}_{Q,2}$

where $\underline{C}_{I,1}$, $\underline{C}_{I,2}$, $\underline{C}_{Q,1}$, and $\underline{C}_{Q,2}$ are orthogonal to each other.

The noncoherent receiver for this modulation consists of the four matched filters followed by envelope detectors whose outputs are sampled at the symbol time $T = 2T_b$. The samples at the matched filter envelope detector outputs for the two inphase chip sequences are compared and an inphase data bit decision is made. The same procedure is followed for the quadrature matched filter envelope detector outputs to make the quadrature data bit decision. Again the inphase and quadrature channels are treated separately. The performance is the same as binary FSK signals with noncoherent detection [2]. Hence we have

$$P_b = \frac{1}{2} e^{-\frac{E_b}{2N_o}} \tag{20}$$

This is shown in Fig. 8 for various interferences.

Note that this noncoherent system results in 3.3 dB degradation compared to the coherent system. This is primarily due to using orthogonal signals rather than antipodal signals. The noncoherent system, however, does not require phase tracking which can be a problem in a multipath and interference environment. For this reason this noncoherent system may require fewer preamble symbols for synchronization associated with each packet of bits in the packet radio application.

To minimize interference due to long multipath delays we can apply the alternating orthogonal chip sequence technique described in Section II. Here we can also use post detection integration techniques. To illustrate this consider multipath where we have m paths to the receiver with average energy $E_1, E_2, E_3, \ldots, E_m$. The energy detector* output at the m path sample times are denoted $Z_1, Z_2, Z_3, \ldots, Z_m$. We assume these sample times are known and the decision is based on

$$\sum_{k=1}^{m} \lambda_k Z_k \tag{21}$$

where $\lambda_1, \ldots, \lambda_m$ is some weighting of the m multipath samples. Using a Chernoff bound [2] we have the bit error probability bound

$$\begin{aligned} P_b &= P\left(\sum_{k=1}^{m} \lambda_k Z_k < \sum_{k=1}^{m} \lambda_k \hat{Z}_k \right) \\ &= P\left(\sum_{k=1}^{m} \lambda_k \left[\hat{Z}_k - Z_k \right] > 0 \right) \\ &\leq E\left\{ \exp \frac{1}{\sigma^2} \sum_{k=1}^{m} \lambda_k \left[\hat{Z}_k - Z_k \right] \right\} \\ &= \prod_{k=1}^{m} E\left\{ \exp \frac{\lambda_k}{\sigma^2} \left[\hat{Z}_k - Z_k \right] \right\} \end{aligned} \tag{22}$$

*We can set a threshold and sample those outputs of the energy detector exceeding this threshold.

where $\{Z_k\}$ are squared samples out of the matched filter/envelope detector of the transmitted chip sequence and $\{\hat{Z}_k\}$ are the corresponding samples from the alternative chip sequence. Assuming multipath delays are confined to the chip sequence time duration we have [6]

$$E\left\{e^{\frac{\lambda_k \hat{Z}_k}{\sigma 2}}\right\} = \frac{1}{1-\lambda_k} \tag{23}$$

$$E\left\{e^{-\frac{\lambda_k Z_k}{\sigma 2}}\right\} = \frac{1}{1+\lambda_k^2} \quad e^{-\left(\frac{\lambda_k}{1+\lambda_k}\right)\frac{E_k}{N_o}} \tag{24}$$

Thus,

$$P_b \leq \prod_{k=1}^{m} \left\{\frac{1}{1-\lambda_k^2} \quad e^{-\left(\frac{\lambda_k}{1+\lambda_k}\right)\frac{E_k}{N_o}}\right\} \tag{25}$$

This bound can be minimized with respect to $0 \leq \lambda_k \leq 1 \qquad k = 1, 2, \ldots, m.$

By just choosing $\lambda_k = \frac{1}{2}$ we have

$$P_b \leq \left(\frac{4}{3}\right)^m \quad e^{-\frac{1}{3N_o}\sum_{k=1}^{m} E_k}$$

$$= e^{-\frac{1}{3}\left(\frac{E_T}{N_o}\right) + m \ln\frac{4}{3}} \tag{26}$$

where

$$E_T = \sum_{k=1}^{m} E_k \tag{27}$$

This bound is plotted in Fig. 9 for the example with

$$m = 4$$

and

$$E_k = \left(\frac{1}{2}\right)^{k-1} E_b \qquad k = 1,2,3,4$$

Post detection integration offers some improvement in performance. Since Fig. 9 is an upper bound compared to the exact values of Fig. 8 it is not clear how much improvement is actually achieved here.

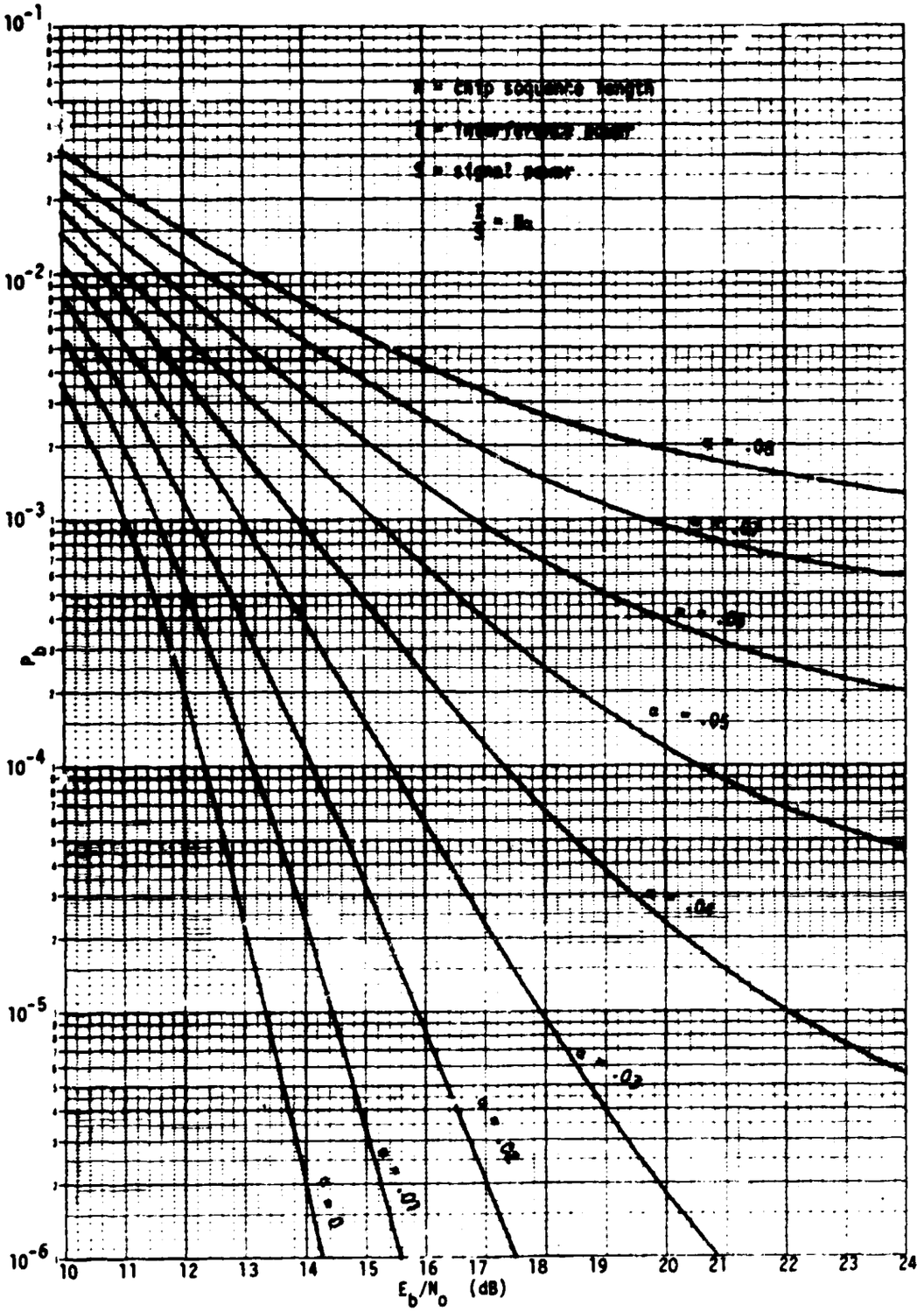


Figure 8. Noncoherent BPSK, QPSK (Uncoded)

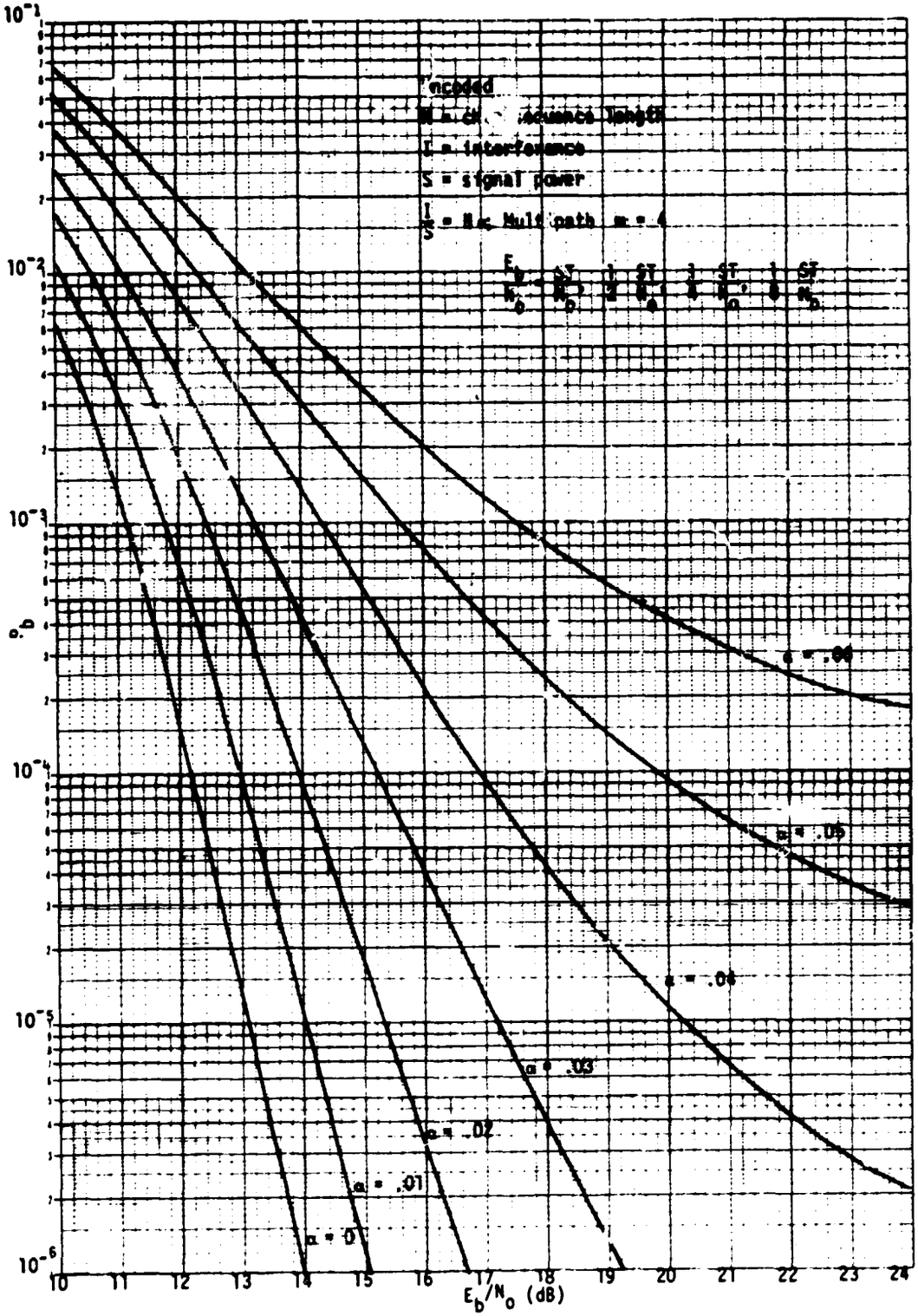


Figure 9. Noncoherent BPSK, QPSK with PDI

V. NONCOHERENT SYSTEM – CODING

We can use orthogonal convolutional codes for the noncoherent receiver system. The bit error bound for such a code of constraint length K is [2]

$$P_b \leq \frac{D^k}{(1 - 2D)^2} \tag{28}$$

where for the PDI detector described above we have

$$D \leq \prod_{k=1}^{m} \left\{ \frac{1}{1 - \lambda_k^2} \; e^{-\left(\frac{\lambda_k}{1 + \lambda_k}\right) \frac{E_k}{N_o}} \right\} \tag{29}$$

For no multipath this becomes

$$D \leq \frac{1}{1 - \lambda^2} \; e^{-\left(\frac{\lambda}{1 + \lambda}\right) \frac{E_b}{N_o}} \tag{30}$$

where choosing $\lambda = \frac{1}{2}$ yields*

$$D \leq \frac{4}{3} \; e^{-\frac{1}{3}\left(\frac{E_b}{N_o}\right)} \tag{31}$$

Fig. 10 shows the bit error bound for this noncoherent case with no multipath. With multipath and the use of PDI the performance improves. Hence these curves can be viewed as loose upper bounds on the bit error probabilities. With interference and orthogonal convolutional codes of constraint lengths K = 4 and K = 5 we have the bit error bounds plotted in Figs. 11 and 12 respectively. Comparing these curves with the uncoded case of Fig. 8 we see again the large coding gain achieved when there are interference signals in the channel.

* $\lambda = 1/2$ is not the optimum choice but provides a simple evaluation of the bound.

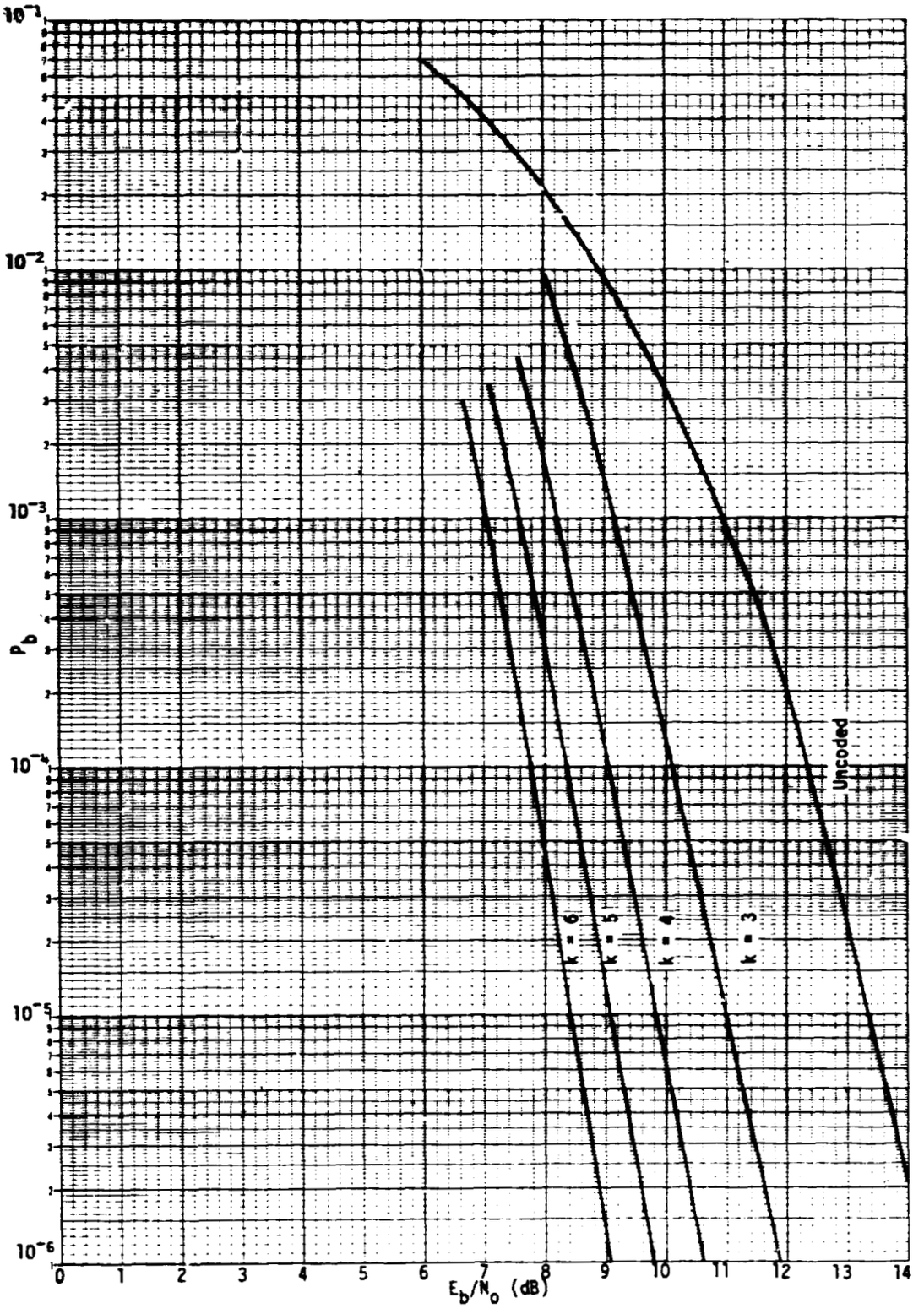


Figure 10. Noncoherent BPSK, QPSK (I = 0)

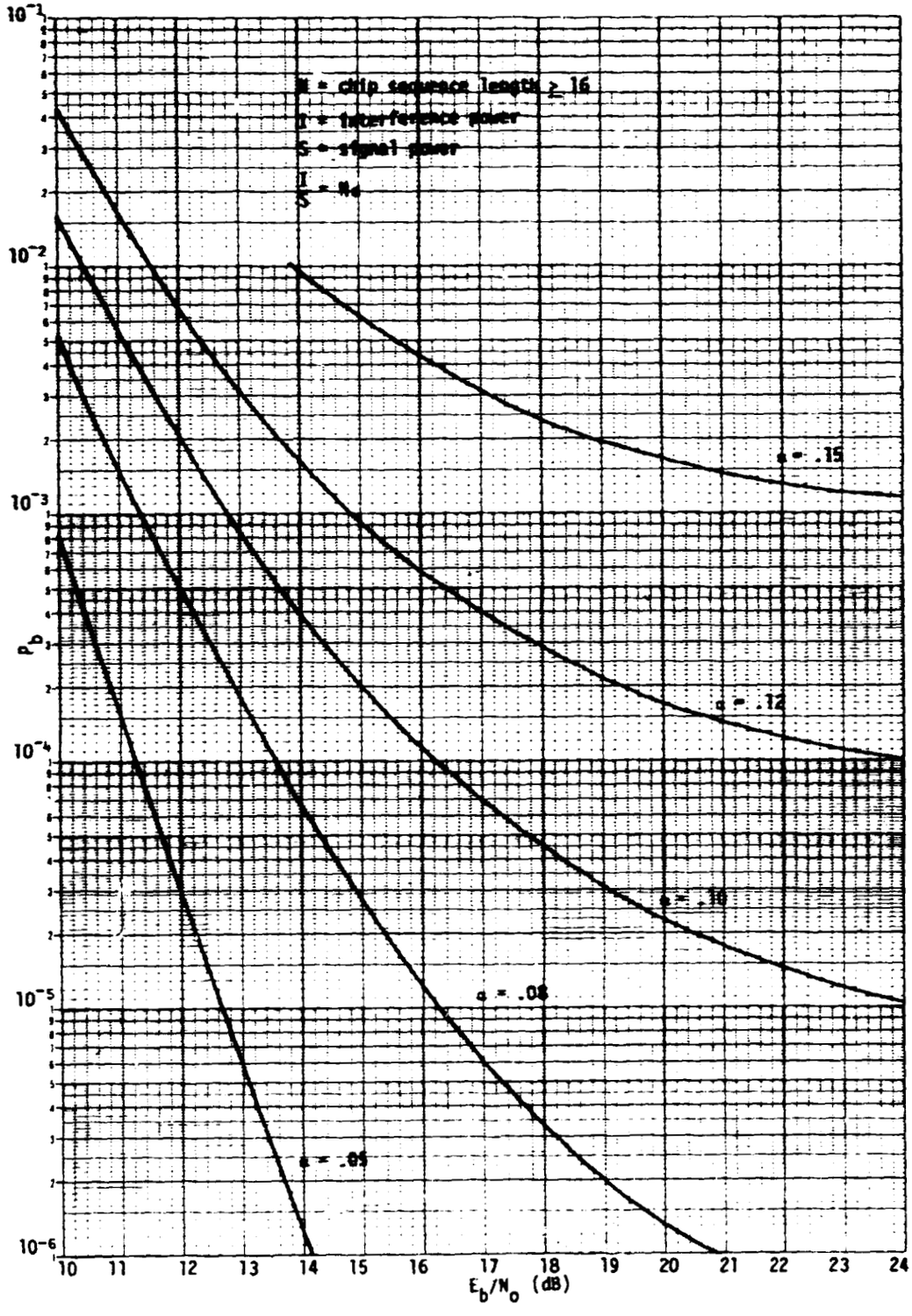


Figure 11. Noncoherent BPSK, QPSK (Coded K = 4)

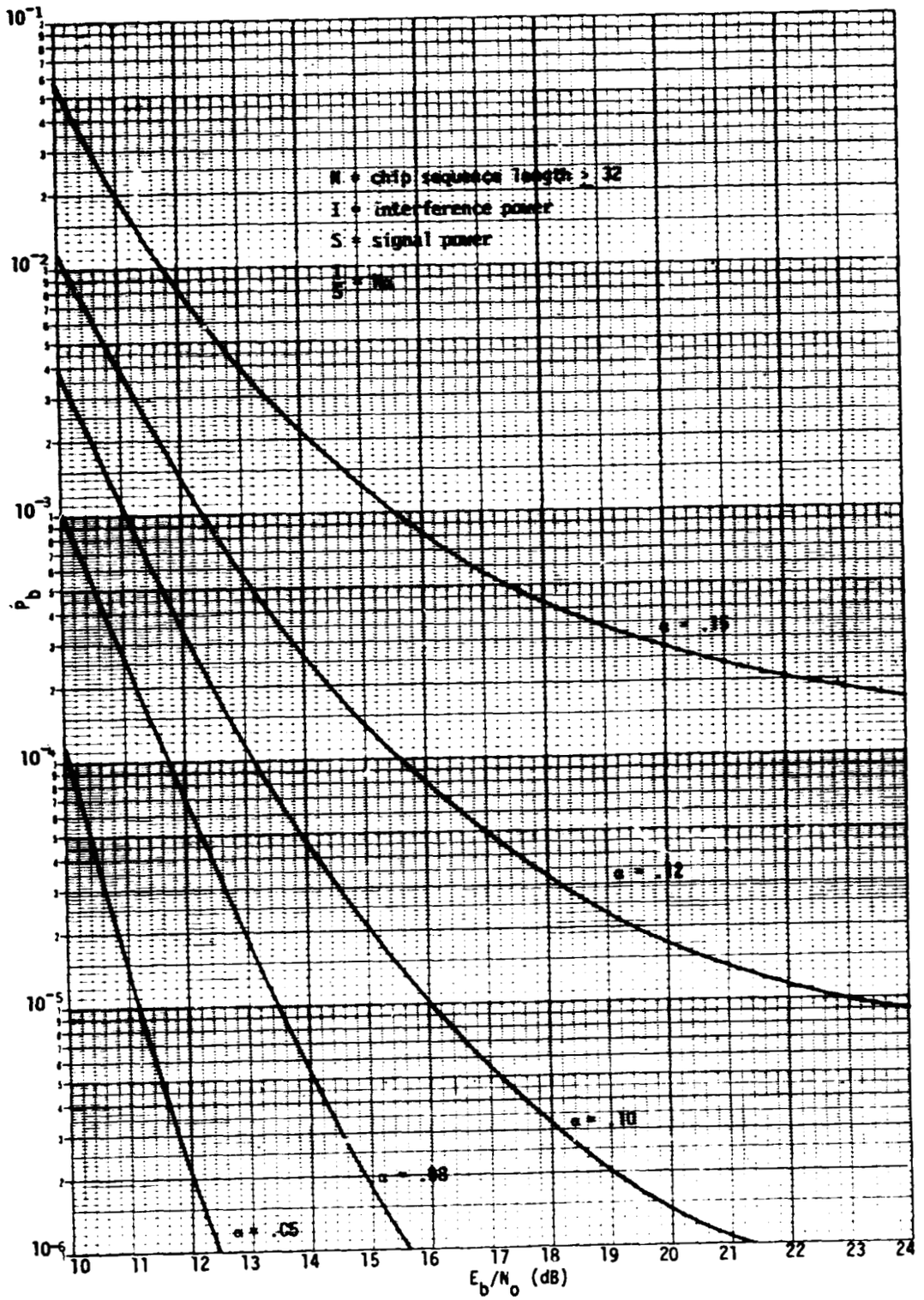


Figure 12. Noncoherent BPSK, QPSK (Coded K = 5)

VI. CONCLUSIONS

Although there is some loss in performance, the noncoherent receiver system has the advantage of a simpler receiver structure and the easy employment of post detection integration (PDI) to collect the multipath energy within a data bit duration. By using orthogonal convolutional codes the noncoherent system certainly performs better than the uncoded coherent system especially in an interference environment. It may also reduce the number of overhead bits required in each packet of data bits.

As an example of a coded noncoherent system we can have N = 32 orthogonal bit sequences for both the inphase and quadrature chip sequences where each chip sequence is selected by an orthogonal convolutional encoder with K = 4. The set $2^K = 16$ orthogonal chip sequences of length N = 32 used by the inphase signal is orthogonal to the quadrature orthogonal chip sequence set. The noncoherent receiver uses N = 32 matched filters followed by envelope detectors (possibly PDI too). The 16 detectors corresponding to the inphase chip sequences are then inputs to a Viterbi decoder with only $2^{K-1} = 8$ states. Another similar Viterbi decoder operates on the quadrature chip sequence detector outputs. The performance is shown in Fig. 10 for the K=4 curve and in Fig. 11 for the case with interference signals.

We compare the various coded and uncoded cases for both coherent and noncoherent receivers in Fig. 13 where we fix the bit error bounds at 10^{-5}. These curves show the locus of required E_b/N_o for various values of the interference parameter α given by (3). The noncoherent cases have the advantage of robustness and easy employment of PDI.

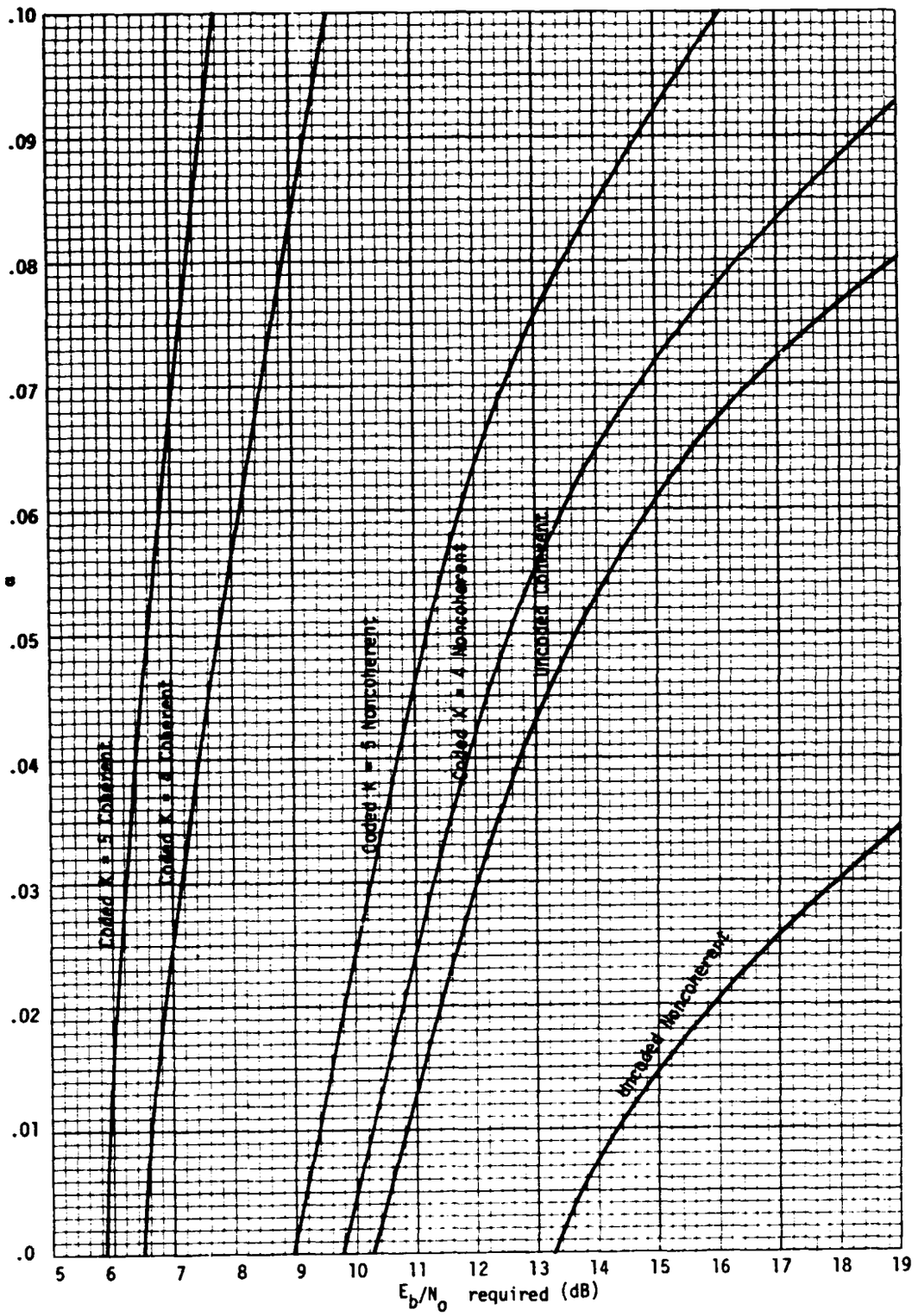


Figure 13. Fixed $P_b \leq 10^{-5}$ Requirements

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