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**Optimum processing of multilevel narrow-band digital FM
utilizing noncoherent FM demodulation**

Bolourchi, Nader, Ph.D.

City University of New York, 1992

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Ann Arbor, MI 48106**

A

**OPTIMUM PROCESSING OF MULTILEVEL NARROW-BAND DIGITAL FM
UTILIZING NONCOHERENT FM DEMODULATION**

BY

NADER BOLOURCHI

A dissertation submitted to the Graduate Faculty in Engineering in partial
fulfillment of the requirement for the degree of Doctor of Philosophy

The City University of New York

1992

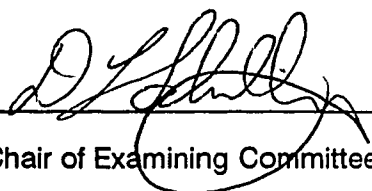
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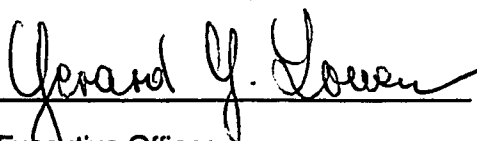
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This manuscript has been read and accepted for the Graduate Faculty in Engineering in satisfaction of the dissertation requirement for the degree of Doctor of Philosophy.

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Chair of Examining Committee

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ABSTRACT**Optimum Processing of Multilevel Narrow-Band Digital FM
Utilizing Noncoherent FM Demodulation**

by

Nader Bolourchi

Adviser: Professor Donald L. Schilling

A new technique for the detection of narrow-band multilevel continuous phase frequency shift keyed (CPMFSK) signals is described to attain spectral efficiencies in excess of 2 bits/s/Hz. This method is a noncoherent demodulation which utilizes a limiter-discriminator-integrator (LDI) followed by a nonlinear postprocessor to combat the intersymbol interference caused by the predemodulation IF filter. The approach taken is to obtain an expression for the frequency output of the demodulator and then to determine the required thresholds as specified by the past history of the signal. This problem has not been addressed in the literature.

The error rate performance is computed for this receiver, both with and without postprocessor as a function of various system parameters, to support the superiority of this algorithm. Quantitative results are presented for the case of additive Gaussian noise. Computer simulation results show significant improvement in the bit error probability (BER) at high SNR as well as at low SNR.

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Special thanks to my loving wife Arezou whose patience and encouragement made it all the more worthwhile.

Many thanks to my family for having faith in me and supporting me through out my entire education

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1 INTRODUCTION

1.1 STATEMENT OF THE PROBLEM

Frequency modulation is used as the modulation of choice whenever the communication channel is nonlinear or randomly time varying, and when the use of coherent detection is undesirable as is the case in dispersive channels. For example, cellular telephony employs FM and the next generation of mobile telephones will employ digital FM . In this new system, voice will be digitized to 16 kb/s . To insure a maximum number of users the FM bandwidth will be made as small as possible. A typical bandwidth, discussed in the literature, is 16 kb/s , i.e., the bandwidth will be equal to the bit rate.

Between modulation methods which have been proposed for future mobile communication services, tamed frequency modulation (TFM) [1] and Gaussian filtered minimum shift keying (GMSK) [2] which fall under digital narrowband continuous phase FM, are desirable scheme fitting the mobile radio communication bandwidth and out-of-band radiation power requirements.

In the rapid fading conditions of mobile radio channels, fast acquisition of the reference carrier phase is one of the crucial technical problems encountered when coherent detection is used [3]. Therefore, the noncoherent demodulation tech-

niques can be applied due to lack of carrier recovery circuit and ease of circuit implementation in expense of decrease of power efficiency [4]-[7]. Among noncoherent schemes, Limiter-Discriminator-Integrator (LDI) detection demonstrates protection opposed to rapid fading environment.

When using frequency modulation to modulate data, the question arises to the effect of pre-modulation (IF) filtering and adjacent channel interference on the probability of error when a FM discriminator is used for the purpose of detection. The problem arises because FM is a nonlinear modulation/demodulation process [8]. Thus, distortion caused by the prefilter and the effect of adjacent channel interference both effect the phase of the FM signal in a nonlinear manner. Such nonlinear distortion cannot be compensated for by the use of a linear equalizer as in an AM system, such as phase shift keying (PSK), quadrature amplitude modulation (QAM), etc.

This problem has been partially solved by Hirono [7] and Pawula [9] who determined the improved probability of error of binary frequency shift keying ($BFSK$) obtained when compensating for the effect of IF filtering by using a nonlinear detector following the FM discriminator. The problem concerned with this research arose when trying to design a narrowband FM system of $B \text{ bits/s/Hz}$ and $B > 2$. In such a system M -ary FSK was used with trellis coding by Schilling [10]. To maintain a narrow bandwidth the spacing between adjacent frequencies was decreased. Although this resulted in an increase in error rate the trellis code

was designed to combat this increase and produce a net error rate decrease. However, the decrease in separation between adjacent frequencies resulted in a large amount of intersymbol interference being introduced when a sharp cutoff narrowband IF filter was employed. On the other hand, if the IF filter was broadened by using a filter which was, for example, Gaussian shaped with a 3 dB cutoff bit rate, an interfering signal present in the adjacent band affected the demodulation process and caused the error rate to increase.

This dissertation is concerned with a study of the distortion produced by the IF filter when M -ary FSK is employed when $M > 2$ and the algorithm needed to compensate for the distortion caused by this filter. This problem has not been addressed in the literature.

1.2 SIGNIFICANCE OF THE PROBLEM

The FSK modem is of lower cost, size and complexity than QAM or M -ary PSK . However, the basic advantage of FSK is the use of noncoherent detection. As a result of its inherently wide bandwidth FSK is often limited to low bit rate channels. The introduction of continuous phase FSK ($CPFSK$) such as tamed FM (TFM) has resulted in a considerable reduction in the FM spectrum but an increase in error rate. Trellis coding of the $CPFSK$ signal [10] has resulted in a lower error rate. However, the trellis coded $CPFSK$ signal is sensitive to IF filtering.

The solution to the problem proposed here will result in a narrowband FSK system operating with low error rates using discriminator demodulation. Such systems could be used as telephone modems in a cellular telephone system, for the transmission of high speed data over cables, in $ISDN$ networks and for microwave, satellite, HF or Meteor Burst channels.

To illustrate one of the above mentioned host of applications consider the next generation of mobile, cellular, telephones. This is the application addressed by Hirono [7]. In this application the voice is digitized using an adaptive delta modulator, such as the $CVSD$, at a 16 kb/s rate. The proposed bandwidth of the system is $BW = 16\text{ kHz}$, so that the channel spacing can be 16 kHz apart. This bandwidth currently employed for cellular telephony. Consider that by using trellis coded FM one can transmit 3 bits/symbol . Using the trellis code will increase the number of frequencies to 16. Decreasing the deviation between frequencies can result in a spectrum which is 40 dB below the maximum at a bandwidth of 8 kHz [10]. The result is an increased number of channels. This improvement is possible using the nonlinear post detector ("equalizer") that is proposed in this work.

2 BACKGROUND

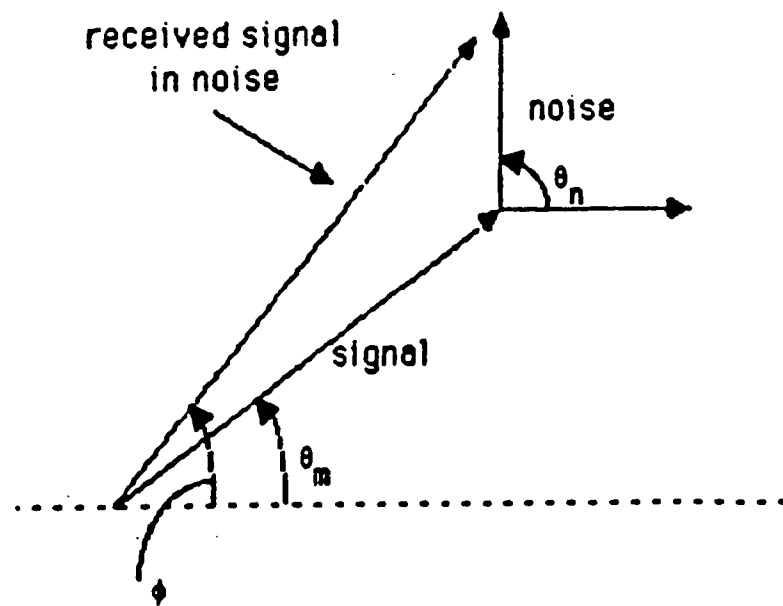
In 1963 Rice published a "theory" of *FM* demodulation which is now considered a classical, standard, approach (as well as the only available analytical approach) and which is discussed in details in many undergraduate texts [8]. Briefly, Rice pointed out that the *FM* signal in noise could be represented as the sum of two phasors, as shown in Fig 2.1. When the signal to noise power ratio is large, typically $P_S / P_N > 100$, the noise is small "most of the time" and the FM problem can be linearized and readily solved. The result is well known:

$$\frac{S_o}{N_o} = 3\beta_{rms}^2 \frac{P_S}{P_N} \quad (2.1)$$

where β_{rms} is the rms value of the phase deviation of the modulating signal from the carrier and S_o / N_o is the output signal power to noise power ratio measured at the output of the discriminator.

Further, and the crux of Rice's hypothesis, is that on rare occasions the noise phasor can grow so that its phasor-length exceeds the length of the *FM* signal phasor (see Fig 2.1) and in addition θ_n can rotate more rapidly than θ_m and cause the resultant phasor ϕ to rotate by 2π radians. Since the output of a FM discriminator is proportional to $\dot{\phi}$, a very sharp spike results with area proportional to 2π . One

FIGURE 2.1: A FM SIGNAL CAN BE REPRESENTED AS A SUM OF PHASORS



can actually hear this noise, which has been reported on in the literature since Armstrong invented *FM*, and it has the sound of static consisting of "pops and/or clicks".

Rice assumed that the spike noise was a Poisson process and determined the mean time between spikes. In addition, Rice showed that if the noise phasor increased but did not cause a 2π phase change in ϕ , the effect of such a phase rotation dose not significantly increase the noise level.

Rice then had two noises: Gaussian noise, when spikes could be neglected (which is more than 99% of the time) and the spike noise, and he assumed these two noises were additive [11].

Rice's theory was verified and extended by Schilling in a series of papers from 1964-1974. Schilling verified that distribution of spikes was Poisson and that a spike had a higher probability of occurring if θ_m was moving in a clockwise manner and θ_n was moving counter clockwise and when $\dot{\theta}_m$ was a maximum [12]. Referring to Fig. 2.1, and noting that the probability of the noise phasor exceeding the signal phasor, is small, one "sees" the reasonableness of a positive spike occurring more often when $\theta_m < 0$, than when $\theta_m > 0$.

2.1 BINARY FSK

Schilling also extended Rice's work into the area of detection of binary frequency shift keyed ($BF SK$) data [13]. At the same time Mazo and Salz [8] independently extended Rice's work for M -ary FSK , i.e., M frequencies, when the discriminator is followed by an integrate and dump filter. Schilling's work also considered the effect of the frequency deviation on the error rate and on the error reduction by detecting and eliminating the spikes. These two papers formed the basis of much of the research in the area of FSK , research which is still continuing 22 years later.

In [13] the authors showed that the worst case sequence from the point of view of error rate is the alternating 1010 sequence, a result also pointed out in [7] and [9]. Further, a comparison is made between the FM discriminator and the matched filter (MF). This result [Ref. 13, Fig. 5] is shown in Fig. 2.2. Note that the performance of the discriminator is comparable to that of the MF until the frequency deviation $\Delta f \approx f_b/2$. At higher frequency deviation the spike noise increases the error rate. The authors also showed [Ref. 13, Fig. 9] that some spikes could be detected and corrected. As seen from Fig. 2.3 such improvement is less than a factor of 10.

In all of these analyses and those performed subsequently by most other investigators, the effect of the pre-discriminator (IF) filter on the output error rate was ignored. In [14]-[17] it was assumed that the receiver filter has a transfer function which is the inverse of the transfer function of the transmitter filter; hence the signal

FIGURE 2.2: COMPARISON OF THE MATCHED FILTER AND FM DISCRIMINATOR

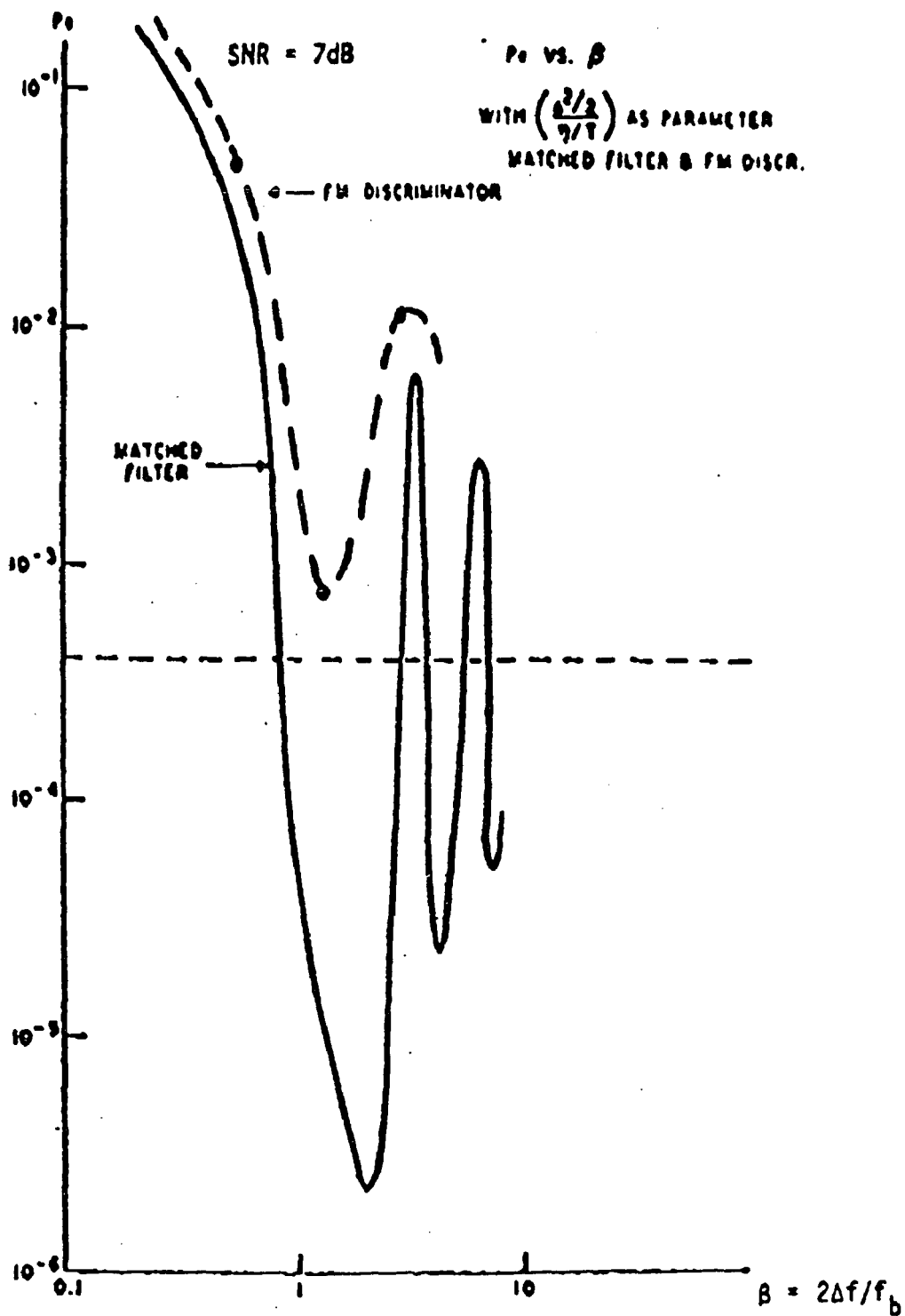
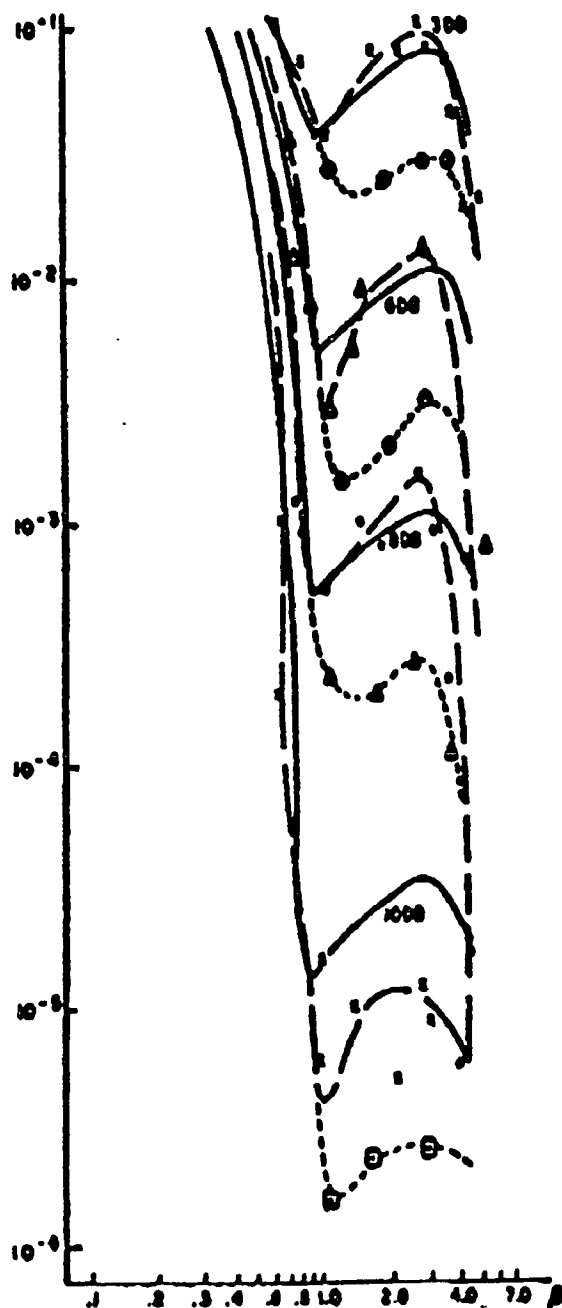


FIGURE 2.3: ERROR PROBABILITY VARIATION WITH MODULATION INDEX
SHOWING THE EFFECT OF SPIKE DETECTION AND CORRECTION

PE VS β WITH CNR AS PARAMETER
(1010 SEQUENCE)

THEORETICAL ———
EXPERIMENTAL WITHOUT CORRECTION ———
EXPERIMENTAL WITH CORRECTION ·····



is not distorted by the system. In [12] the presence of a narrow-band filter in the transmitter and receiver is considered; however, there it was assumed that the filters do not affect the "click" noise, and threshold settings. Experimentalists and practitioners always employed a relatively wide bandwidth filter prior to the discriminator to avoid distorting the FM signal and so they too did not investigate the effect that narrowing this filter had on the error rate. Indeed, everyone appeared content using a bandwidth

$$B = 2(\Delta f + f_b) \quad (2.2)$$

which is similar to the expression derived by Carson for analog FM [8], where $2\Delta f$ is the maximum peak-to-peak frequency deviation of the modulating signal and f_b is the bit rate of the data.

2.2 BFSK WITH NARROWBAND IF FILTERING

In the past few years, spectral congestion has increased dramatically and has focused everyone's attention on the use of bandwidth-efficient modulation techniques. Investigators now turned their attention to the wide bandwidth required of FM and sought to find ways to reduce this bandwidth. Referring to Equation (2.2) we immediately see that B can be reduced by decreasing f_b and/or Δf . However, decreasing f_b implies decreasing the input data rate and usually we want to increase

the input data rate. Decreasing Δf places the frequency separation closer together and an optimum frequency separation was found [13,19] for the case of *BFSK*; decreasing the deviation further increases the error rate.

A major breakthrough was made by Hirono [7], who obtained an optimum nonlinear post-detection processor to undo the nonlinear distortion caused by narrowing the pre-detection filter for binary *FSK*. In their paper the authors show, see Fig. 2.4, that the error rate can be reduced by 10^6 at a *SNR* of 16 dB by setting an adaptive threshold which depends on the past history of the data stream. One of the two aspects of this proposed research is to extend the work of [7] to *M*-ary *FSK*.

The procedure employed by Hirono was to be cognizant of the fact that *BFSK*, if undisturbed by the *IF* filter, produced a broad eye, from which one could separate a "0" from a "1" by using a single threshold placed midway between the two possible. However, when distortion of the *BFSK* signal was generated by a narrowband *IF* filter, the eye depended on the past history of the data. Fig. 2.5 shows that if intersymbol interference is present multiple thresholds are needed to optimize performance.

Hirono employed a post-detection integrator as a filter prior to setting the threshold and sampling. Since the noise at the output of the discriminator is not white and the data is corrupted by intersymbol interference produced in a nonlinear manner,

FIGURE 2.4: COMPARISON OF BER PERFORMANCES FOR THE MULTILEVEL DECISION AND CONVENTIONAL ZERO-THRESHOLD DECISION METHODS. $B_bT = 0.25$, $M = 0.5$, $BT = 1.0$

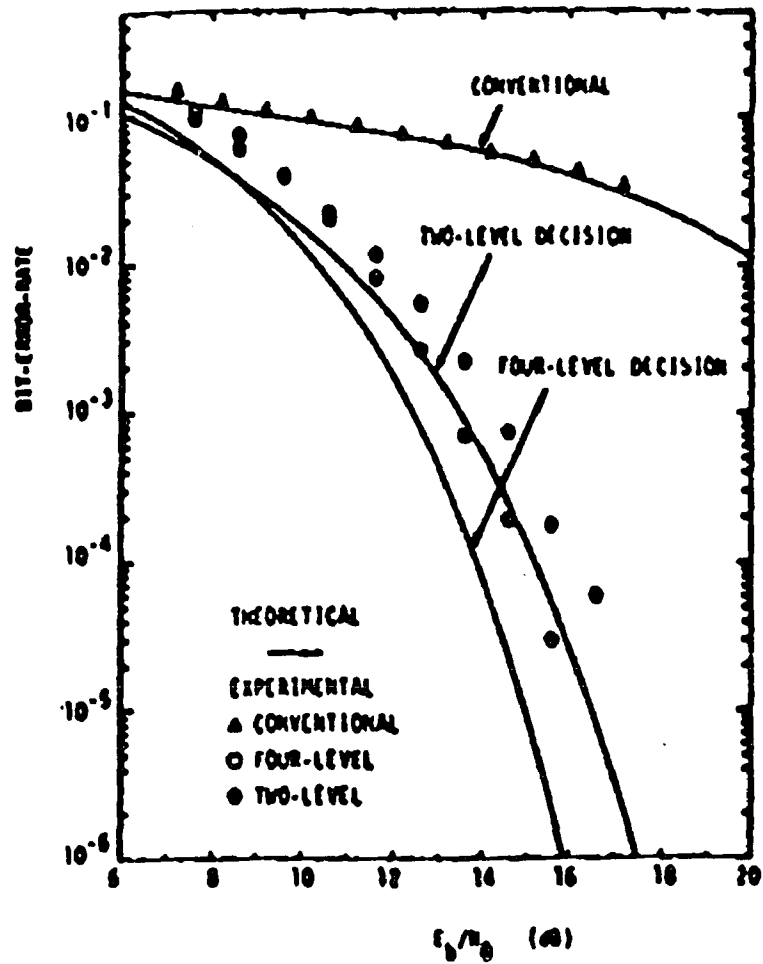
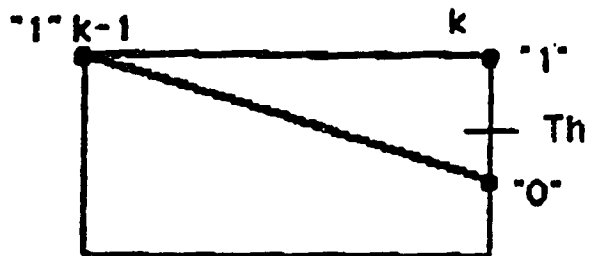
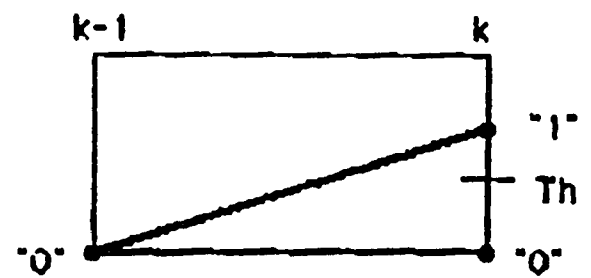


FIGURE 2.5: TWO THRESHOLDS CAN IMPROVE PERFORMANCE IN BFSK



- (a) If the bit at $k-1$ is a "1", set the threshold midway between the two possible outputs at k .



- (b) If the bit at $k-1$ is a "0", set the threshold midway between the two possible outputs at k .

the integrator is not the optimum post-detection filter to be employed.

2.3 ADJACENT CHANNEL INTERFERENCE

Adjacent channel interference is closely related to the increased error rate caused by narrowband IF filtering. Indeed, the purpose of the narrowband IF filter is to permit a smaller separation between channels, i.e, to attenuate to a greater degree the signals in adjacent channels.

To illustrate this point consider two orthogonal FSK signals

$$v_1(t) = \sqrt{2P_1} \cos \omega_1 t \quad \text{"0"} \quad (2.3a)$$

$$v_2(t) = \sqrt{2P_1} \cos \omega_2 t \quad \text{"1"} \quad (2.3b)$$

where $\omega_2 > \omega_1$, and two adjacent signals

$$v_3(t) = \sqrt{2P_3} \cos(\omega_2 t + \Delta t) \quad (2.3c)$$

$$v_4(t) = \sqrt{2P_3} \cos(\omega_1 t - \Delta t) \quad (2.3d)$$

For simplicity we assume that the IF filter does not distort v_1 and v_2 and that v_3 and v_4 are the remainder of the adjacent channel signals after being attenuated by the IF filter. The orthogonal $BFSK$ signals are therefore "surrounded" by two equal amplitude interferers.

Assuming a duration T we have

$$(f_2 - f_1)T = 1 \quad (2.4)$$

and using a Gram-Schmitt representation for the signals we have [8]

$$v_1(t) = \sqrt{P_1 T} \phi_1(t) \quad (2.5a)$$

$$v_2(t) = \sqrt{P_1 T} \phi_2(t) \quad (2.5b)$$

$$v_3(t) = \sqrt{P_3 T} \left(\frac{\sin \Delta T}{\Delta T} \right) \phi_1(t) + \sqrt{P_3 T} \left(\frac{\sin \Delta T}{2\pi + \Delta T} \right) \phi_2(t) + (\dots) \phi_3(t) \quad (2.5c)$$

and

$$v_4(t) = \sqrt{P_3 T} \left(\frac{\sin \Delta T}{2\pi + \Delta T} \right) \phi_1(t) + \sqrt{P_3 T} \left(\frac{\sin \Delta T}{\Delta T} \right) \phi_2(t) + (\dots) \phi_3(t) + (\dots) \phi_4(t) \quad (2.5d)$$

where

$$\phi_1(t) = \sqrt{\frac{2}{T}} \cos \omega_1 t \quad (2.6a)$$

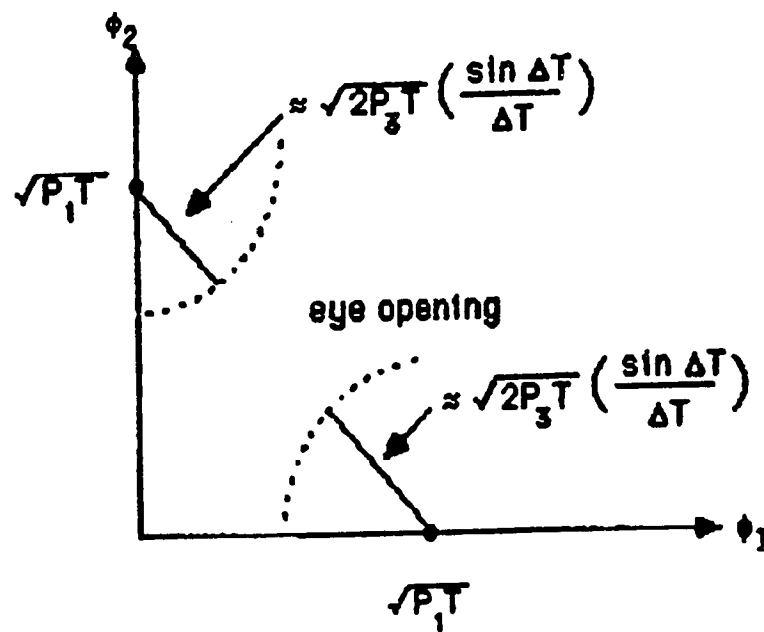
$$\phi_2(t) = \sqrt{\frac{2}{T}} \cos \omega_2 t \quad (2.6a)$$

and ϕ_3 and ϕ_4 are irrelevant since they do not affect the error rate of the *BFSK* signal.

The effect of the interference is to close the "eye" as shown in Fig 2.6. In this figure, the eye when fully opened is $\sqrt{2P_1 T}$ wide. Assuming $\Delta T \ll 2\pi$ we see that the effect of the two adjacent channel signals is to effectively close the eye to an opening E , where

$$E \cong \sqrt{2P_1 T} - 2\sqrt{2P_3 T} \frac{\sin \Delta T}{\Delta T}$$

FIGURE 2.6: ADJACENT CHANNEL INTERFERENCE CLOSES THE EYE



$$= E_0 \left(1 - 2 \frac{\sin \Delta T}{\Delta T} \sqrt{\frac{P_3 T}{P_1}} \right) \quad (2.7.a)$$

and

$$E_0 = \sqrt{2 P_1 T} \quad (2.7.b)$$

The eye opening E decreases as ΔT decreases. If, for example, $\Delta T = \pi/2$

$$E \cong E_0 \left[1 - 1.27 \sqrt{\frac{P_3}{P_1}} \right] \quad (2.8a)$$

A 1 dB decreasing in the eye opening corresponds to an increased error rate of a factor of 10. This corresponds to

$$0.2 \cong 1.27 \sqrt{\frac{P_3}{P_1}} \quad (2.8b)$$

$$P_3 \cong \frac{P_1}{40} \quad (2.8c)$$

Thus, it is seen that even for orthogonal binary FSK interfering signals have to be significantly attenuated; in this case where $\Delta T = \pi/2$, 16 dB of attenuation required to increase the eye opening by 1 dB and decrease the error rate by factor of 10.

It is the purpose of this research to pursue the case of narrowband M -ary FSK where the effect of the interference is even more pronounced.

3 ADAPTIVE THRESHOLD DECISION METHOD

The approach taken is to obtain an expression for the frequency output of the demodulator and then determine the required thresholds as specified by the past history of the signal. Fig. 3.1 illustrates the narrow-band continuous phase digital FM system with a nonlinear post-processor. The received signal $v(t)$ at the input to the receiver IF filter can be written as

$$v(t) = \sqrt{2P_s} \cos[\omega_o t + \theta(t)] + n(t) \quad (3.1)$$

where P_s is the signal power, $\omega_o = 2\pi f_o$ is the center IF filter radian frequency, $\dot{\theta}(t)$ is the modulating signal and $n(t)$ is white Gaussian noise with power spectral density, $\eta/2$. For simplicity we shall assume $\dot{\theta}(t)$ consists of a sequence of frequency hops as shown in Fig. 3.2, each hop lasting for a time T_s . Since the FM signal is continuous phase, each information carrying phase function $\theta(t)$ is continuous at all times for all combinations of data symbols [20]

$$\theta(t) = \omega_{k-1}(t - kT_s) + \sum_{i=1}^k \omega_i T_s \quad kT_s \leq t \leq (k+1)T_s \quad (3.2)$$

FIGURE 3.1.a: NARROW-BAND DIGITAL FM COMMUNICATION SYSTEM

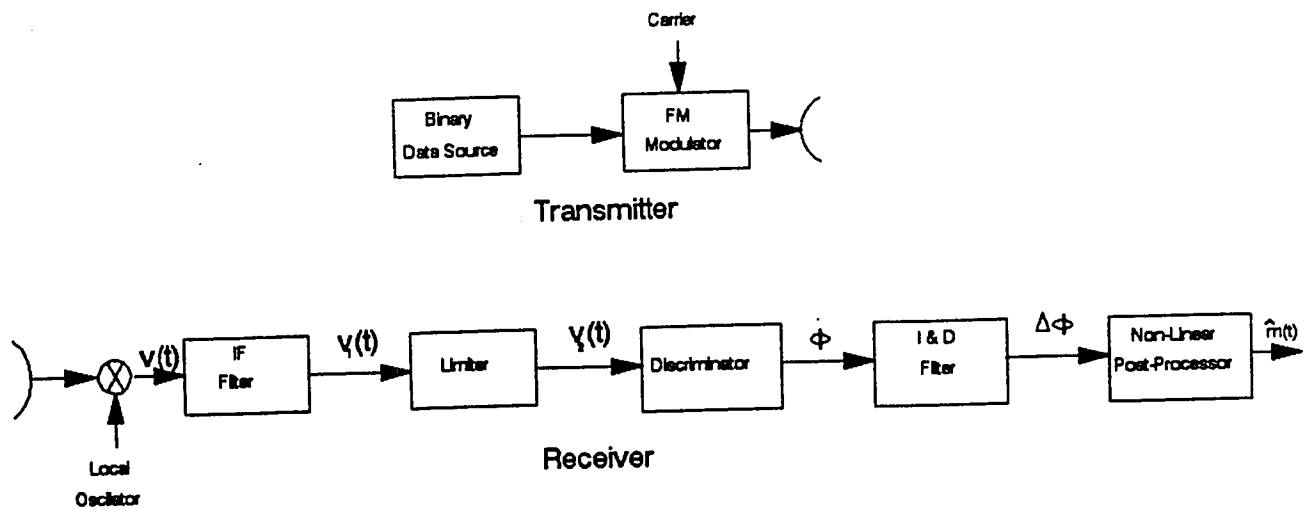


FIGURE 3.1.b: NON-LINEAR POST PROCESSOR

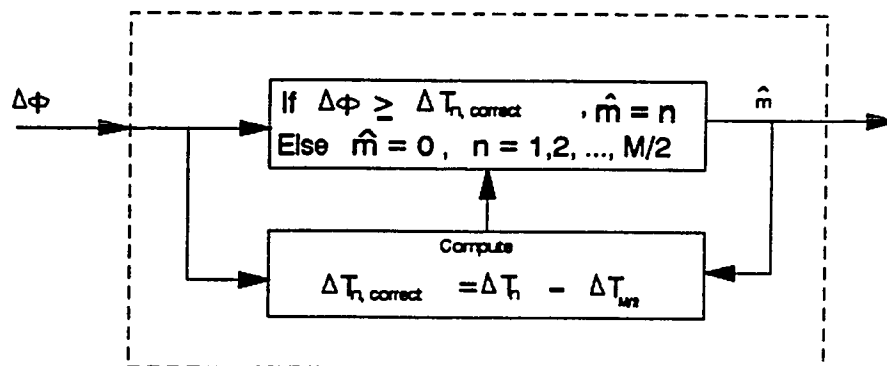
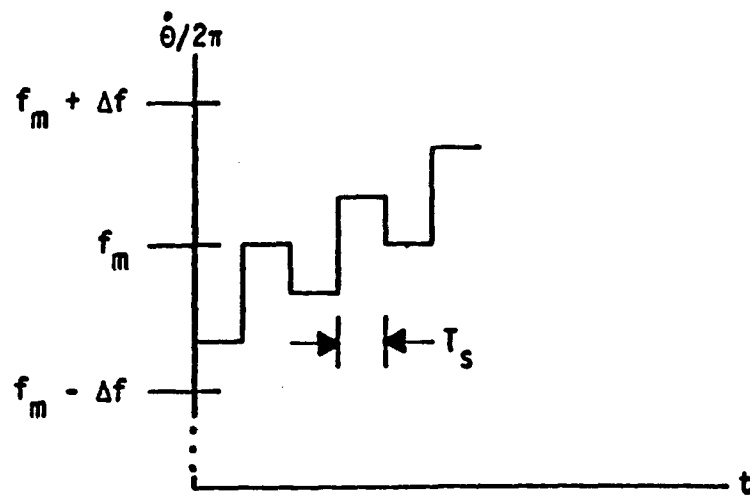


FIGURE 3.2: THE MODULATION IS LIMITED TO THE FREQUENCY BAND $f_m - \Delta f$ to $f_m + \Delta f$



in which $\omega_j = 2\pi f_a d_j$ is the transmitted radian frequency in the time interval $t = jT_s$ where f_a is half of the frequency separation and d_i is the M -ary data symbol which takes values $\pm 1, \pm 2, \dots, \pm(M-1)$. M is normally a power of 2, i.e., $M = 2^n$ ($M = 2$ binary, 4 quaternary, 8 octal system, etc.)

As a result of phase continuity, the waveform $v(t)$ is "smooth" and exhibits no abrupt changes in phase. Therefore, we avoid the intersymbol interference caused by nonlinear amplifiers [8].

The sum of the FM signal and noise enters the FM discriminator which consists of an IF bandpass filter followed by a hard limiter. Harmonics higher than f_m are rejected by a wideband filter. The IF filter distorts the signal and results in a time varying signal envelope, $\rho(t)$, a distorted signal phase, $\phi(t)$, and a signal dependent noise term, $n_o(t)$. At the output of IF filter, (3.1) can be written as [21]

$$v_1(t) = \rho(t) \cos[\omega_o t + \phi(t)] + n_o(t) \quad (3.3)$$

where [9]

$$\phi(t) = \tan^{-1}(N(t)/D(t)) = \tan^{-1} U(t) \quad (3.4.a)$$

and

$$D(t) = \int_{-\infty}^t h(t-\tau) \cos \theta(\tau) d\tau \quad (3.4.b)$$

$$N(t) = \int_{-\infty}^t h(t-\tau) \sin \theta(\tau) d\tau \quad (3.4.c)$$

in which $h(t)$ is the low pass equivalent of the bandlimiting filter. Let us define the following:

$$s(t) = e^{i\theta(t)} \quad (3.5)$$

then

$$D(t) = \Re[s(t)*h(t)] \quad (3.6.a)$$

$$N(t) = \Im[s(t)*h(t)] \quad (3.6.b)$$

hence

$$\phi(t) = \text{phase of } [s(t)*h(t)] \quad (3.7)$$

$n_o(t)$ can be expanded about the modulated carrier frequency using its in-phase and quadrature components:

$$n_o(t) = \zeta(t) \cos[\omega_o + \phi(t)] + \xi(t) \sin[\omega_o + \phi(t)] \quad (3.8)$$

where $\zeta(t)$ and $\xi(t)$ are independent baseband zero mean Gaussian processes with variance $\sigma^2 = \eta B_n$ where B_n is the IF noise bandwidth. Although the band restricting IF filter is considered to be part of the receiver, the results are applicable to the situation where part of the band restriction occurs in a filter at the transmitter. Substituting (3.8) into (3.3) yields

$$v_1(t) = r(t) \cos[\omega_o t + \Phi(t)] \quad (3.9)$$

where the output phase, $\Phi(t)$, is the sum of the signal phase, $\phi(t)$, and an unwanted part which is the output noise of the detection process, $\psi(t)$, i.e.,

$$\Phi(t) = \phi(t) + \psi(t) \quad (3.10.a)$$

$r(t)$ and $\psi(t)$ are given by [21]

$$r(t) = \sqrt{\rho(t) + \zeta^2(t) + \xi^2(t)} \quad (3.10.b)$$

and

$$\psi(t) = \tan^{-1} \mu(t) = \tan^{-1} \left(\frac{\xi(t)}{\rho(t) + \zeta(t)} \right) \quad (3.10.c)$$

The bandpass limiter suppresses the amplitude variation and its output $v_2(t)$ can be expressed as

$$v_2(t) = \cos[\omega_o t + \Phi(t)] \quad (3.11)$$

Then, the discriminator recovers the derivative of the output phase, $\dot{\Phi}(t)$,

$$\dot{\Phi}(t) = \dot{\phi}(t) + \dot{\psi}(t) \quad (3.12.a)$$

where, using (3.4.a) and (3.10.c),

$$\dot{\phi} = \frac{\dot{U}}{1 + U^2} = \frac{D \dot{N} + N \dot{D}}{N^2 + D^2} \quad (3.12.b)$$

and

$$\dot{\psi}(t) = \frac{\dot{\mu}}{1 + \mu} \quad (3.12.c)$$

The postdetection filter which is chosen to be an Integrate-and-Dump ($I \& D$) filter with an integration time of T_s "reintegrates" this derivative producing a phase difference

$$\psi_i(\tau) = \int_{\tau-T_s}^{\tau} \dot{\psi}(t) dt = \psi(\tau) - \psi(\tau - T_s) \quad (3.13)$$

At the end of each interval T_s , the output of the ($I \& D$) filter is sampled. At $\tau = (k+1)T_s$ we obtain

$$\Delta\phi[(k+1)T_s] = \Delta\phi[(k+1)T_s] + \Psi[(k+1)T_s] \quad (3.14)$$

where $\Delta\phi[(k+1)T_s]$ is the filtered signal phase difference between time $(k+1)T_s$ and kT_s obtained using (3.4)

$$\Delta\phi[(k+1)T_s] = \phi[(k+1)T_s] - \phi(kT_s) \quad (3.15.a)$$

where, from (3.4.a), we get

$$\phi[(k+1)T_s] = \tan^{-1} U[(k+1)T_s] \quad (3.15.b)$$

$$\phi(kT_s) = \tan^{-1} U(kT_s) \quad (3.15.c)$$

$$\Delta\phi[(k+1)T_s] = \tan^{-1} \left(\frac{U[(k+1)T_s] - U(kT_s)}{1 + U[(k+1)T_s] \cdot U(kT_s)} \right) \quad (3.15.d)$$

or

$$\Delta\phi[(k+1)T_s] = \tan^{-1} \left(\frac{N[(k+1)T_s] \cdot D(kT_s) - D[(k+1)T_s] \cdot N(kT_s)}{D[(k+1)T_s] \cdot D(kT_s) + N[(k+1)T_s] \cdot N(kT_s)} \right) \quad (3.15.e)$$

and $\Psi[(k+1)T_s]$ is the output noise [7]

$$\Psi[(k+1)T_s] = \psi[(k+1)T_s] - \psi(kT_s) + N_{T_s}, \quad (3.16)$$

where N_{T_s} is the average number of clicks during interval T_s .

To obtain the error rate it is necessary to know the receiver output due to the signal in the absence of noise. We assume the phase change during any time interval T_s is less than or equal to $\pi/2$ to insure narrowband operation.

Since in a M -ary continuous phase FSK communications system the transmitted signal, for the symbol interval T_s is of frequency f_0 or f_1 or ... f_{M-1} , a set of $M-1$ thresholds are placed between every two adjacent possible outputs.

$$T_n[(k+1)T_s] = \frac{\phi_{n+1}[(k+1)T_s] + \phi_n[(k+1)T_s]}{2}$$

$$n = 1, 2, \dots, M-1 \quad (3.17)$$

where $\phi_n[(k+1)T_s]$ is the phase at the $(n+1)T_s$ time interval when the n th symbol is received.

However, the distortion of a *CPMFSK* signal, generated by the narrow-band *IF* filter, shifts the thresholds into improper decision levels. The nonlinear post-processor computes the amount of shift of the threshold levels using (3.15) and (3.17)

$$\begin{aligned} \Delta T_n[(k+1)T_s] &= T_n[(k+1)T_s] - T_n(kT_s) \\ &= \frac{\Delta\phi_{n+1}[(k+1)T_s] + \Delta\phi_n[(k+1)T_s]}{2} \end{aligned} \quad (3.18)$$

To correct the threshold levels, the nonlinear post-processor adds the effect of a shift of the central threshold level $\Delta T_{\frac{M}{2}}$ to all threshold levels and produces $M-1$ new threshold levels

$$\Delta T_{n, \text{correct}}[(k+1)T_s] = \Delta T_n[(k+1)T_s] - \Delta T_{\left(\frac{M}{2}\right)}[(k+1)T_s]$$

$$n = 1, 2, \dots, \left(\frac{M}{2}\right) \quad (3.19)$$

The receiver then decides that the symbol “ n ” was sent if $\Delta\Phi[(k+1)T_s] \geq \Delta T_{n, \text{correct}}[(k+1)T_s]$, $n = 1, 2, \dots, (M-1)$ and “0” otherwise.

3.1 BANDLIMITING FILTER

To illustrate the algorithm we have chosen the class of Butterworth filters as the *IF* filter. The magnitude squared of the frequency response $B(\omega)$ of these filters is of the form

$$|B(\omega)|^2 = \frac{1}{1 + \left(\frac{\omega}{\omega_c}\right)^{2N}} \quad (3.20)$$

where N is number of poles of the filter and ω_c is the 3 - dB bandwidth of the filter.

3.1.1 ONE-POLE BANDPASS IF FILTER

For this example, we use a simple *RC* circuit, which represent a one-pole

Butterworth filter, as the IF filter with $RC = 1/\omega_c$, i.e.,

$$H(\omega) = \frac{1}{1 + i\left(\frac{\omega}{\omega_c}\right)} \quad (3.21)$$

in which its impulse response is

$$h(t) = \alpha e^{-\alpha t} \quad (3.22)$$

where $\alpha = 1/\omega_c$.

Using (3.1) and (3.5), the received signal, in the absence of noise, can be written as

$$s_1(t) = \Re[\sqrt{2P_s} s(t)] \quad (3.23)$$

Then the distorted signal at the output of the bandpass limiter following the IF filter, $s_2(t)$, is found from (3.11):

$$s_2(t) = \Re[s(t) * h(t)] \quad (3.24)$$

Let us define the following:

$$S(t) = s(t) * h(t) = \alpha \int_0^t e^{i\theta(\tau)} e^{-\alpha(t-\tau)} d\tau \quad (3.25.a)$$

Then using (3.6) and (3.7), $S(t)$ can be written as

$$S(t) = e^{i\phi(t)} = D(t) + iN(t) \quad (3.25.b)$$

Using (3.2) and (3.25.a), at the time interval $t = kT_s$ we obtain

$$\begin{aligned} S(kT_s) &= \alpha e^{-\alpha kT_s} \int_0^{kT_s} \sum_{m=0}^{\infty} \beta(\tau - mT_s) e^{i[\theta_m + \omega_m(\tau - mT_s)]} e^{\alpha\tau} d\tau \\ &= \alpha e^{-\alpha kT_s} \left[\int_0^{T_s} \sum_{m=0}^{\infty} \beta(\tau - mT_s) e^{i[\theta_m + \omega_m(\tau - mT_s)]} e^{\alpha\tau} d\tau \right. \\ &\quad \left. + \dots + \int_{(k-1)T_s}^{kT_s} \sum_{m=0}^{\infty} \beta(\tau - mT_s) e^{i[\theta_m + \omega_m(\tau - mT_s)]} e^{\alpha\tau} d\tau \right] \\ &= \alpha e^{-\alpha kT_s} \left[\sum_{n=0}^{k-1} \int_{nT_s}^{(n+1)T_s} \sum_{m=0}^{\infty} \beta(\tau - mT_s) e^{i[\theta_m + \omega_m(\tau - mT_s)]} e^{\alpha\tau} d\tau \right] \end{aligned}$$

$$(3.26.a)$$

i.e., we want to calculate the area under Σ for the $(n + 1)$ th interval, which is equivalent to

$$S(kT_s) = \alpha e^{-\alpha kT_s} \left[\sum_{n=0}^{k-1} \int_{nT_s}^{(n+1)T_s} \beta(\tau - nT_s) e^{i[\theta_n + \omega_n(\tau - nT_s)]} e^{\alpha \tau} d\tau \right] \quad (3.26.b)$$

in which the integrand is the value of Σ within the specified interval. The normalized IF filter output ($\beta(t) = 1$) is

$$\begin{aligned} S(kT_s) &= \alpha e^{-\alpha kT_s} \sum_{n=0}^{k-1} \left[e^{i\theta_n} e^{-i\omega_n(nT_s)} \int_{nT_s}^{(n+1)T_s} e^{(\alpha + i\omega_n)\tau} d\tau \right] \\ &= \alpha e^{-\alpha kT_s} \sum_{n=0}^{k-1} \left[e^{i\theta_n} e^{-i\omega_n(nT_s)} \left(\frac{e^{(\alpha + i\omega_n)nT_s} [e^{(\alpha + i\omega_n)T_s} - 1]}{\alpha + i\omega_n} \right) \right] \\ &= \alpha e^{-\alpha kT_s} \sum_{n=0}^{k-1} \left(\frac{\alpha}{\alpha + i\omega_n} \right) e^{i\theta_n} e^{\alpha nT_s} (e^{(\alpha + i\omega_{n+1})T_s} - 1) \quad (3.26.c) \end{aligned}$$

then,

$$S[(k+1)T_s] = e^{-\alpha k T_s} \left[S(kT_s) + \frac{\alpha}{\alpha + i\omega_k} e^{i\theta_k} (e^{(\alpha + i\omega_{k+1})T_s} - 1) \right] \quad (3.27.a)$$

or

$$\begin{aligned} S[(k+1)T_s] &= \Re\{S[(k+1)T_s]\} + i\Im\{S[(k+1)T_s]\} \\ &= D[(k+1)T_s] + iN[(k+1)T_s] \end{aligned} \quad (3.27.b)$$

Let us define the following:

$$s'[(k+1)T_s] = \frac{\alpha}{\alpha + i\omega_k} e^{i\theta_k} (e^{(\alpha + i\omega_{k+1})T_s} - 1) \quad (3.28)$$

then (3.27.a) can be written as

$$S[(k+1)T_s] = e^{-\alpha T_s} \{S(kT_s) + s'[(k+1)T_s]\} \quad (3.29)$$

where (3.4) and (3.25.b) give

$$N[(k+1)T_s] = e^{-\alpha T_s} \{N(kT_s) + \Im[s'[(k+1)T_s]]\} \quad (3.30.a)$$

$$D[(k+1)T_s] = e^{-\alpha T_s} \{D(kT_s) + \Re[s'[(k+1)T_s]]\} \quad (3.30.b)$$

Using (3.15), (3.17), (3.28), and (3.30), we compute $M - 1$ threshold levels. Then, the proper decision levels can be computed using (3.18) and (3.19).

3.1.2 TWO POLE BANDPASS IF FILTER

Since the transfer function of a filter can be written as

$$H(s) = H_1(s) + H_2(s) \quad (3.31)$$

then the impulse response

$$h(t) = h_1(t) + h_2(t) \quad (3.32)$$

The second-order Butterworth filter's transfer function can then be written as

$$B(s) = \frac{\omega_c^2}{s^2 + \sqrt{2}\omega_c s + \omega_c^2} = \frac{M}{c_1 - i\omega} - \frac{M}{c_2 - i\omega} \quad (3.33)$$

where ω_c is the 3 - dB frequency of the filter. Then

$$b(t) = M[e^{c_1 t} - e^{c_2 t}] \quad (3.34.a)$$

where

$$M = i \frac{\sqrt{2}}{2} \omega_c \quad (3.34.b)$$

$$c_1 = -\frac{\sqrt{2}}{2} \omega_c (1 + i) \quad (3.34.c)$$

$$c_2 = -\frac{\sqrt{2}}{2} \omega_c (1 - i) \quad (3.34.d)$$

$$u(t) = \begin{cases} 1 & t \geq 0 \\ 0 & \text{otherwise} \end{cases} \quad (3.34.e)$$

Assuming $(\sqrt{2}/2)\omega_c = \beta$, (3.34.a) can be written as

$$b(t) = b_1(t) + b_2(t) \quad (3.35.a)$$

where

$$b_1(t) = i\beta e^{-\beta(1+i)t} \quad (3.35.b)$$

$$b_2(t) = i\beta e^{-\beta(1-i)t} \quad (3.35.c)$$

We use a similar approach for two-pole IF filter as used in Sec. 3.1.1 to compute the thresholds. We have

$$S[(k+1)T_s] = S_1[(k+1)T_s] + S_2[(k+1)T_s] \quad (3.36)$$

where

$$\begin{aligned} S_1[(k+1)T_s] = & \frac{\beta}{(\beta + \omega_{k+1}) - i\beta} e^{i\theta_{k+1}} \left(e^{i\omega_{k+1}T_s} - e^{-\beta(1+i)T_s} \right) \\ & + e^{-\beta(1+i)T_s} S_1(kT_s) \end{aligned} \quad (3.37)$$

and

$$\begin{aligned} S_2[(k+1)T_s] = & -\frac{\beta}{(\beta - \omega_{k+1}) + i\beta} e^{i\theta_{k+1}} \left(e^{i\omega_{k+1}T_s} - e^{-\beta(1-i)T_s} \right) \\ & + e^{-\beta(1-i)T_s} S_2(kT_s) \end{aligned} \quad (3.38)$$

The proper threshold level can be computed by (3.15.e), (3.27.b), (3.36), and (3.19).

4 RESULTS

The digital FM system shown in Fig. 3.1 was simulated to determine the eye pattern and probability of error for the continuous phase M -ary FSK signals. The receiver consists of the IF bandlimiting filter, and Limiter-Discriminator-Integrator LDI detector. The nonlinear Post-Processor verifies and corrects all thresholds after each symbol time T_s based on the past history of the signal.

When noncoherent detection is used, the minimum frequency separation of any pair of adjacent frequencies must be $\frac{1}{T_s}$ in order that two adjacent frequencies not to interfere with each other during detection [22]. This requires that the normalized maximum frequency deviation $h = 1$ where $h = \Delta f_{\max} \cdot T_s$ and $\Delta f_{\max} = f_M - f_1$. We have presumed that the transmitter use a voltage controlled oscillator (VCO) which is set to keep $h = 0.25$.

4.1 EYE PATTERNS

In this section we have obtained the eye patterns by computing all possible phase trajectories to show the receiver output due to the signal in the absence of noise.

4.1.1 ONE-POLE BUTTERWORTH FILTER

Figure 4.1 shows the eye patterns for a continuous phase *BFSK* signal filtered by a 1st order band-pass Butterworth filter with one-sided bandwidth of $w = k f_b$ and bandwidth efficiency of $f_b/w = \frac{1}{k}$ bits/s/Hz where $k = 1, 0.5, 0.25, 0.125$. At the time $t = 200$ (normalized by T_b where $t = 200$ is equivalent to the symbol duration) the eye closes as w decreases as shown in Fig. 4.1.a, while it should have only two possible values for the binary case. Figure 4.1.b shows that at $t = 200$ the signal is either positive or negative and the eye is open when the adaptive detector is adopted to correct the threshold levels.

Figure 4.2 shows the eye Patterns for a CP4FSK signal. When adaptive detector is utilized, there are only 4 signals as expected and the eye is open. Fig. 4.3 shows the result for 8FSK. Fig. 4.4 shows that at $BW = 0.125 f_s$ the eye becomes closed when 16FSK is employed. The simulation results presented in Figures 4.1 to 4.4 verifies that the eye is open if

$$f_b/w \leq (64/M) \log_2 M \quad \text{bits/s/Hz} \quad (4.1)$$

When the IF band-pass filter has a constant bandwidth, as M increases, the bandwidth efficiency, f_b/w , increases; however, the eye opening decreases. To keep the eye opening constant as M increases, the bandwidth of IF filter has to

FIGURE 4.1.1: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = f_b$

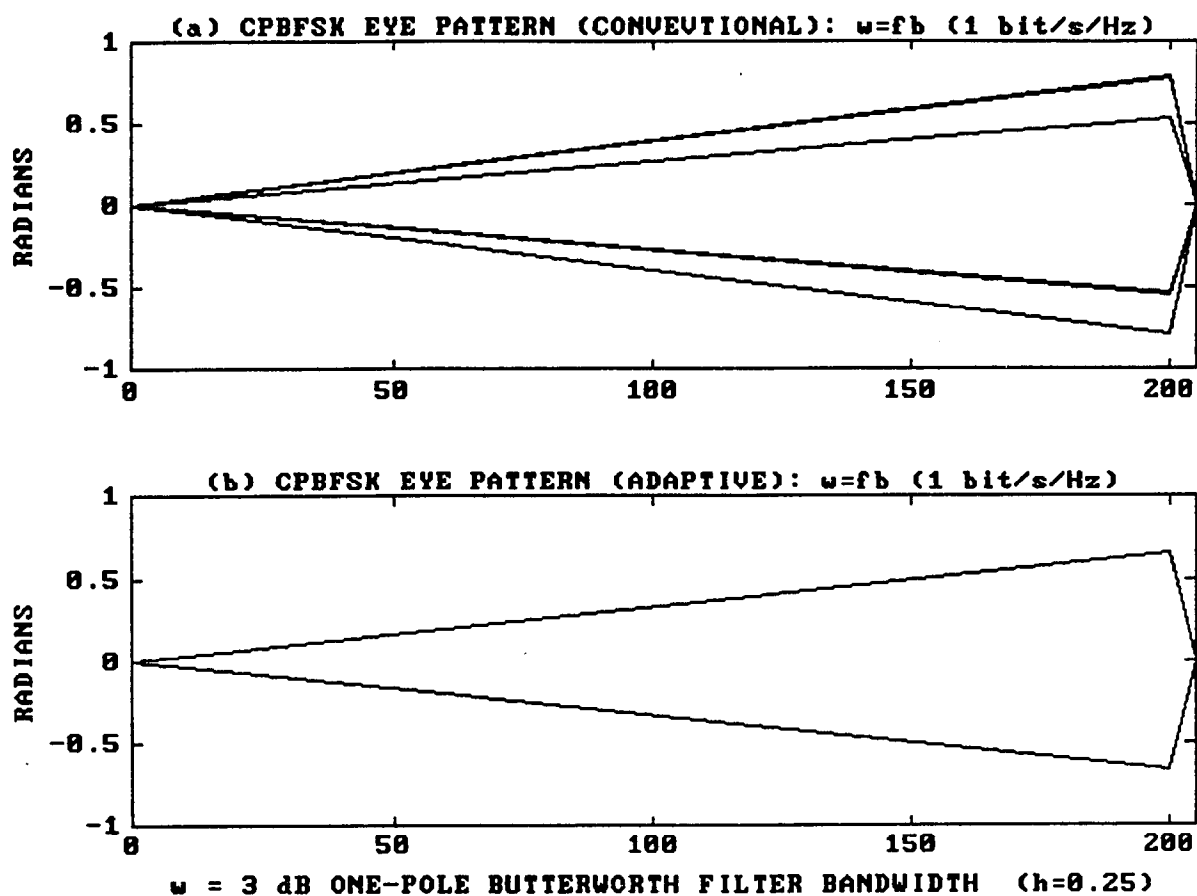


FIGURE 4.1.2: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$

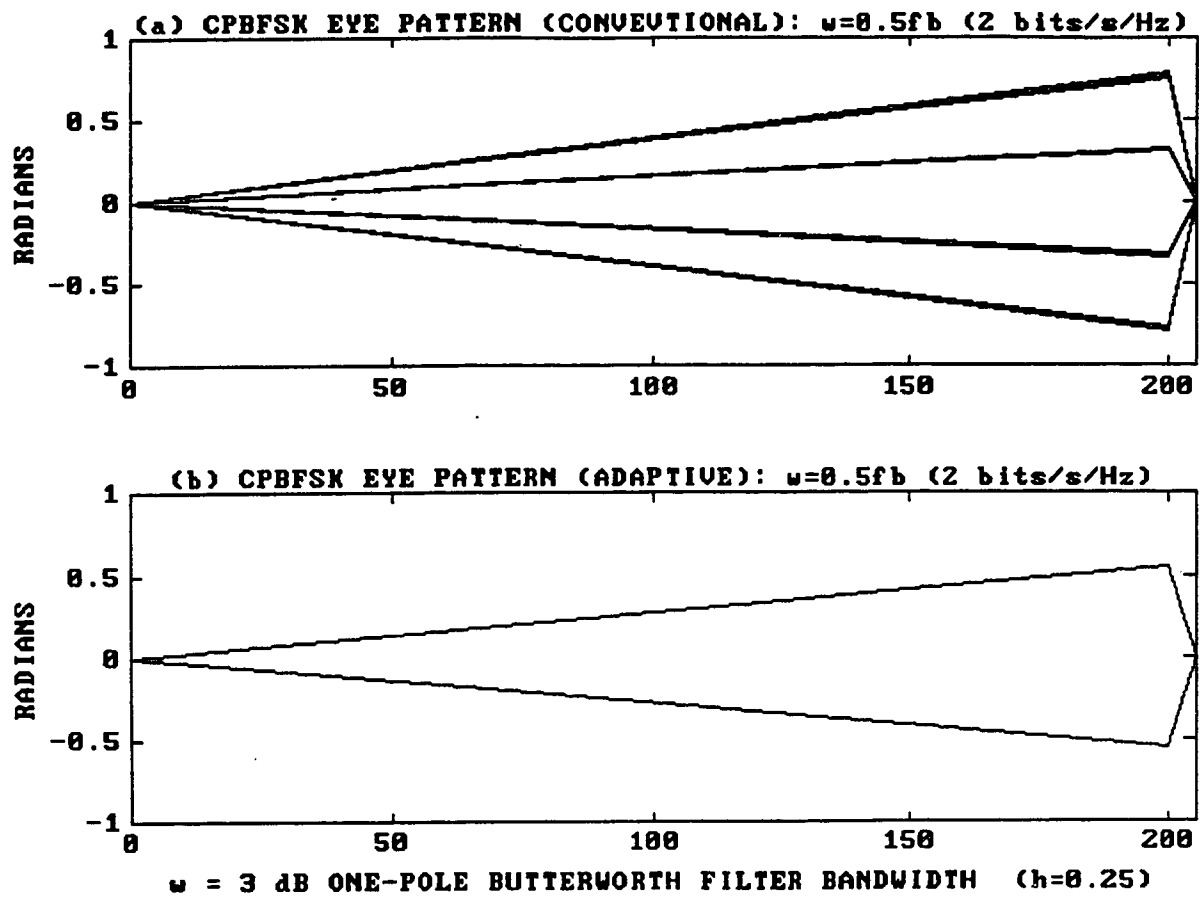


FIGURE 4.1.3: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_b$

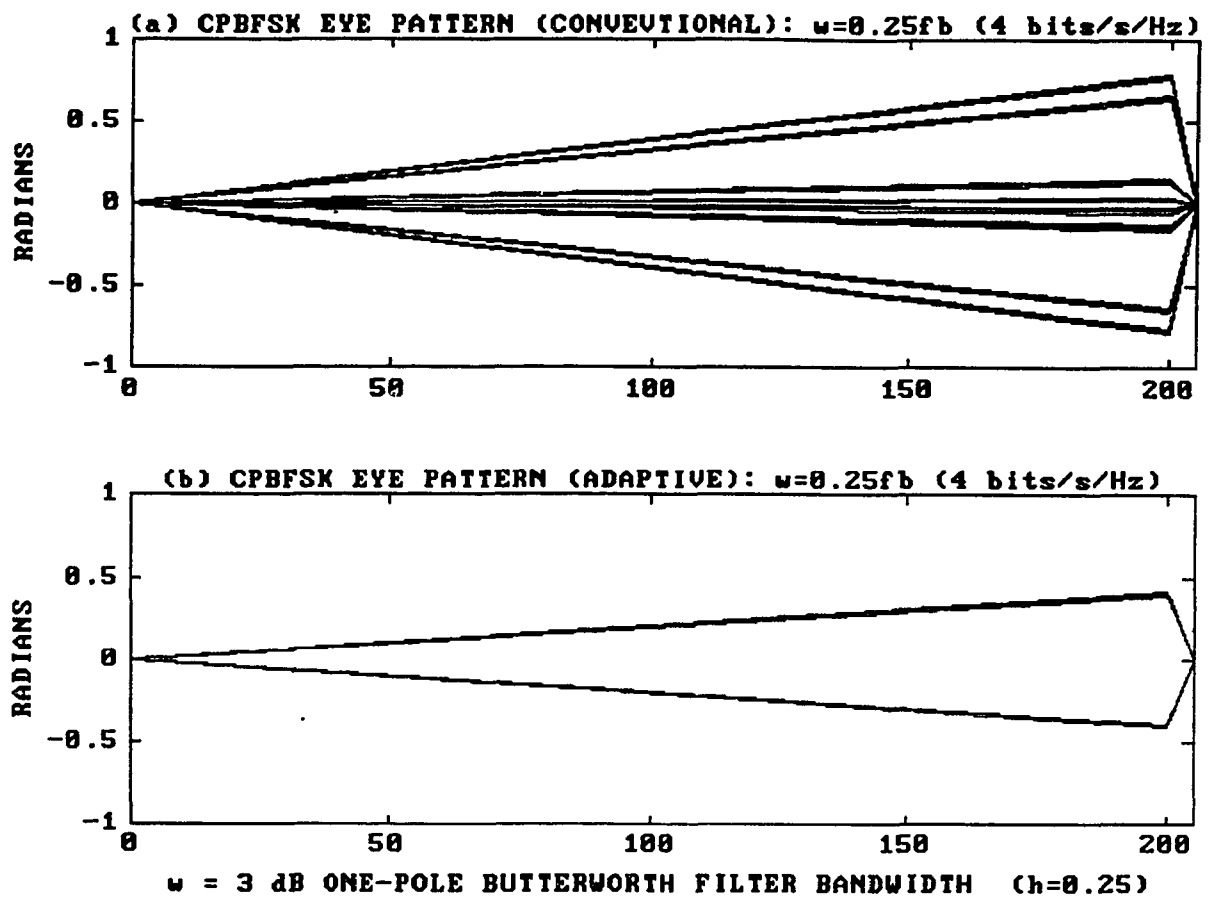


FIGURE 4.1.4: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_b$

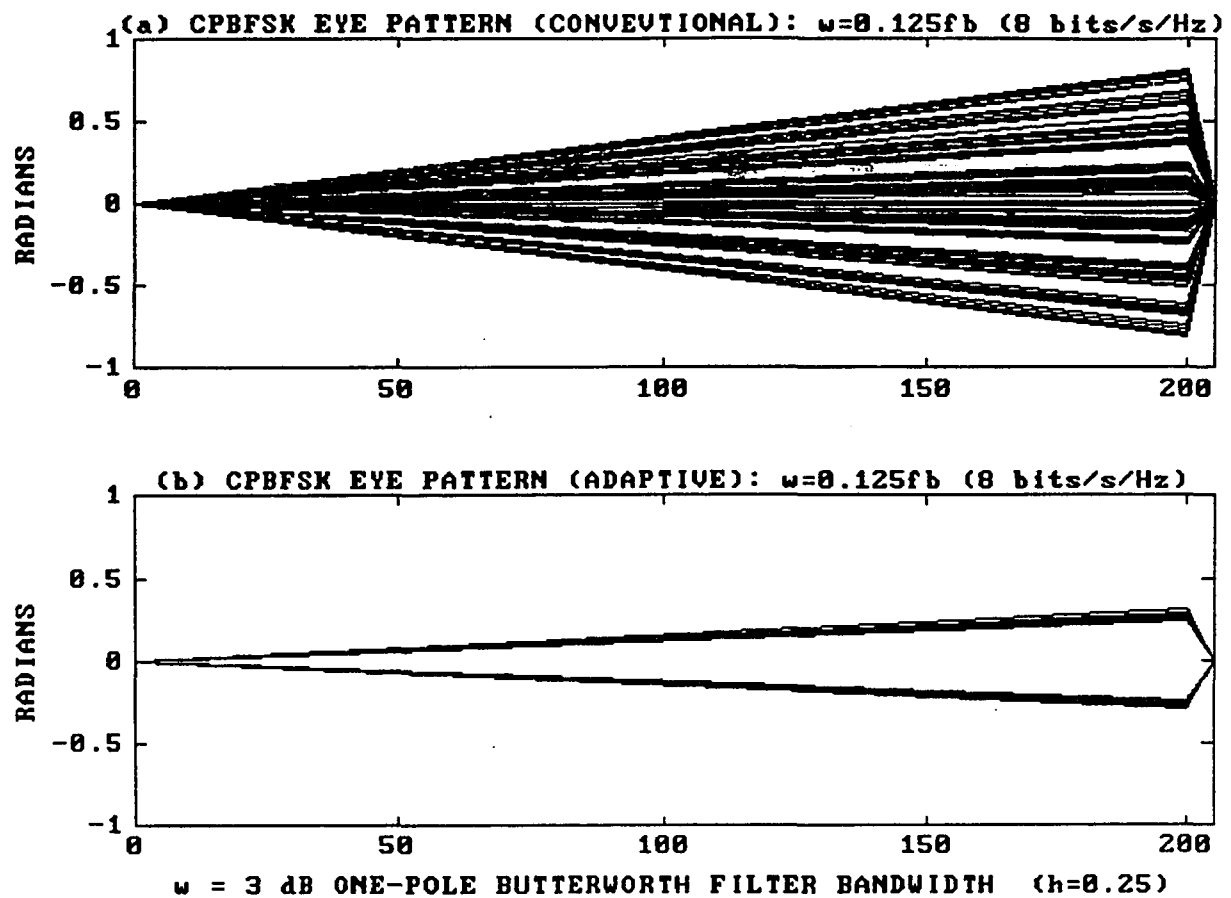


FIGURE 4.2.1: CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = f_s$

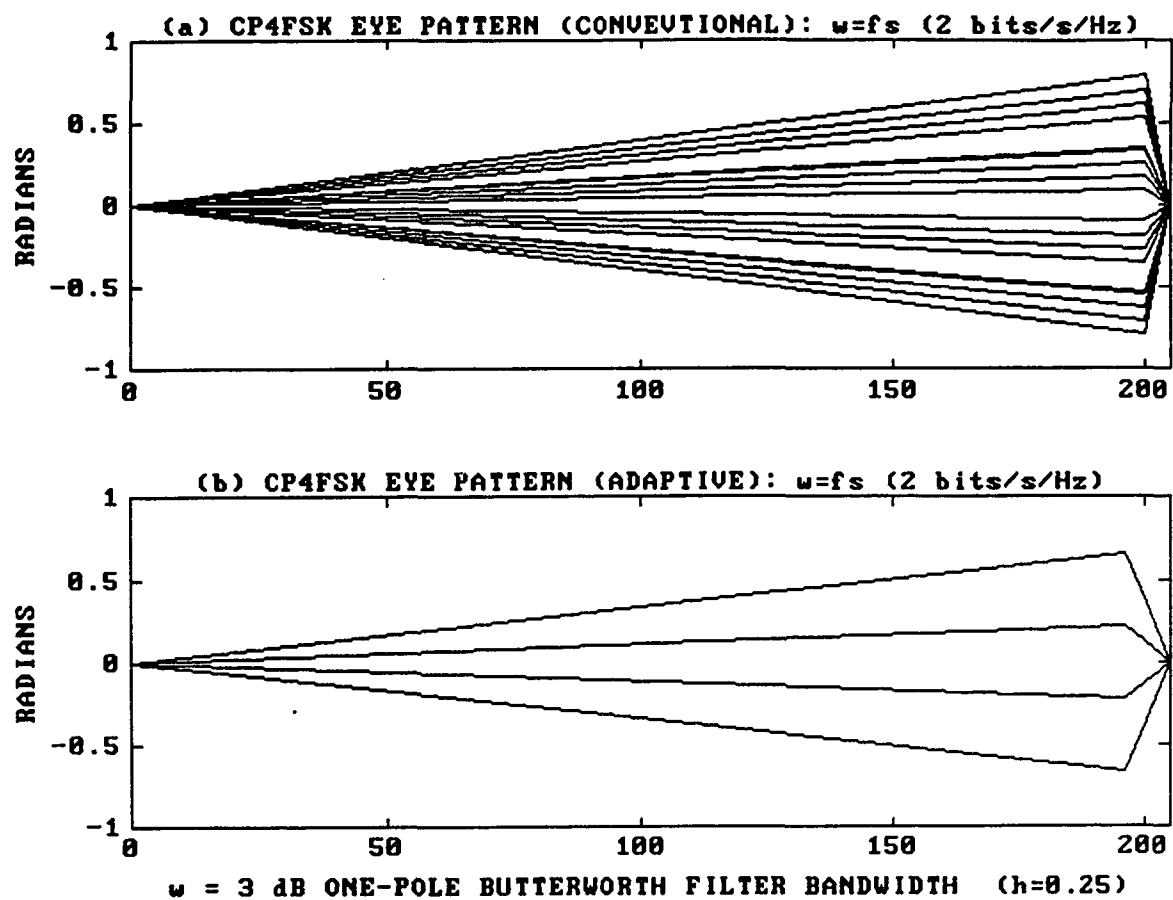


FIGURE 4.2.2: **CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$**

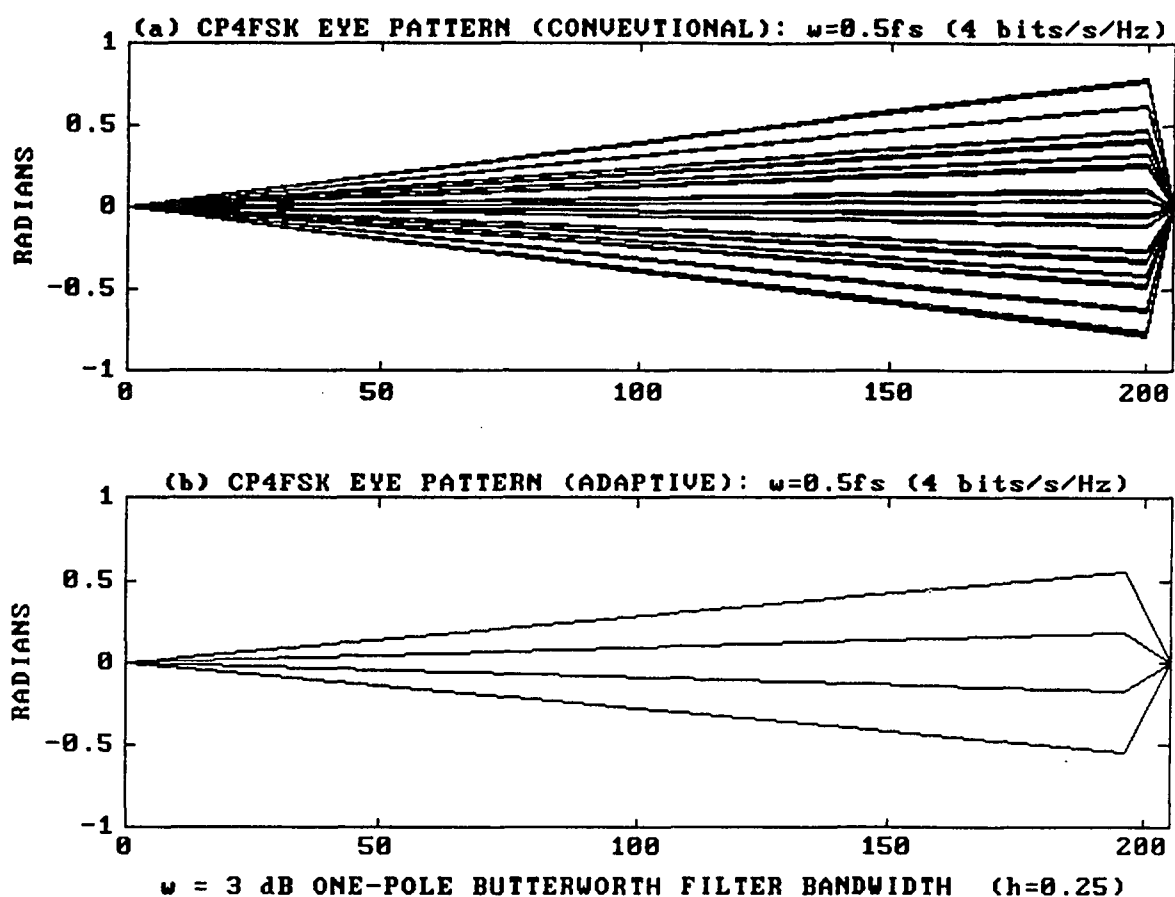


FIGURE 4.2.3: CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

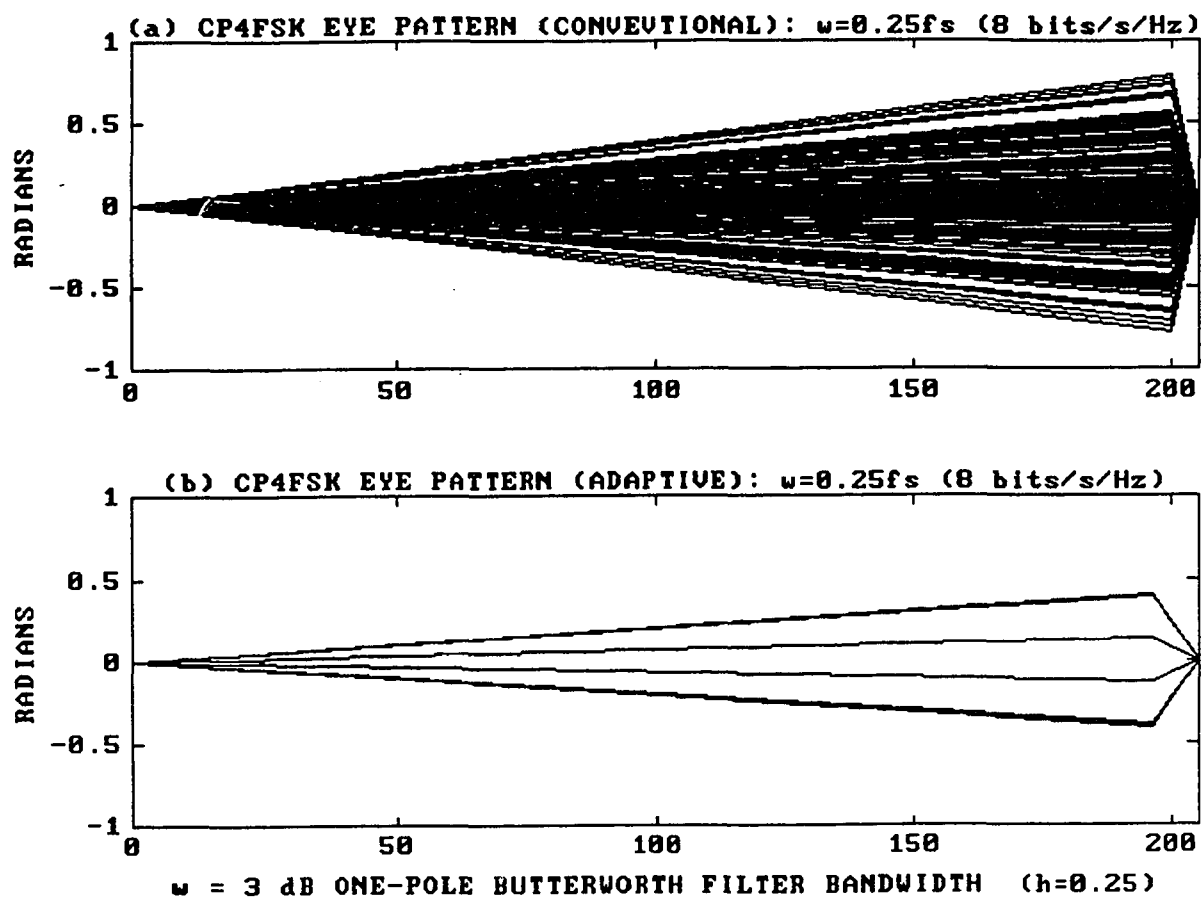


FIGURE 4.2.4: CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_s$

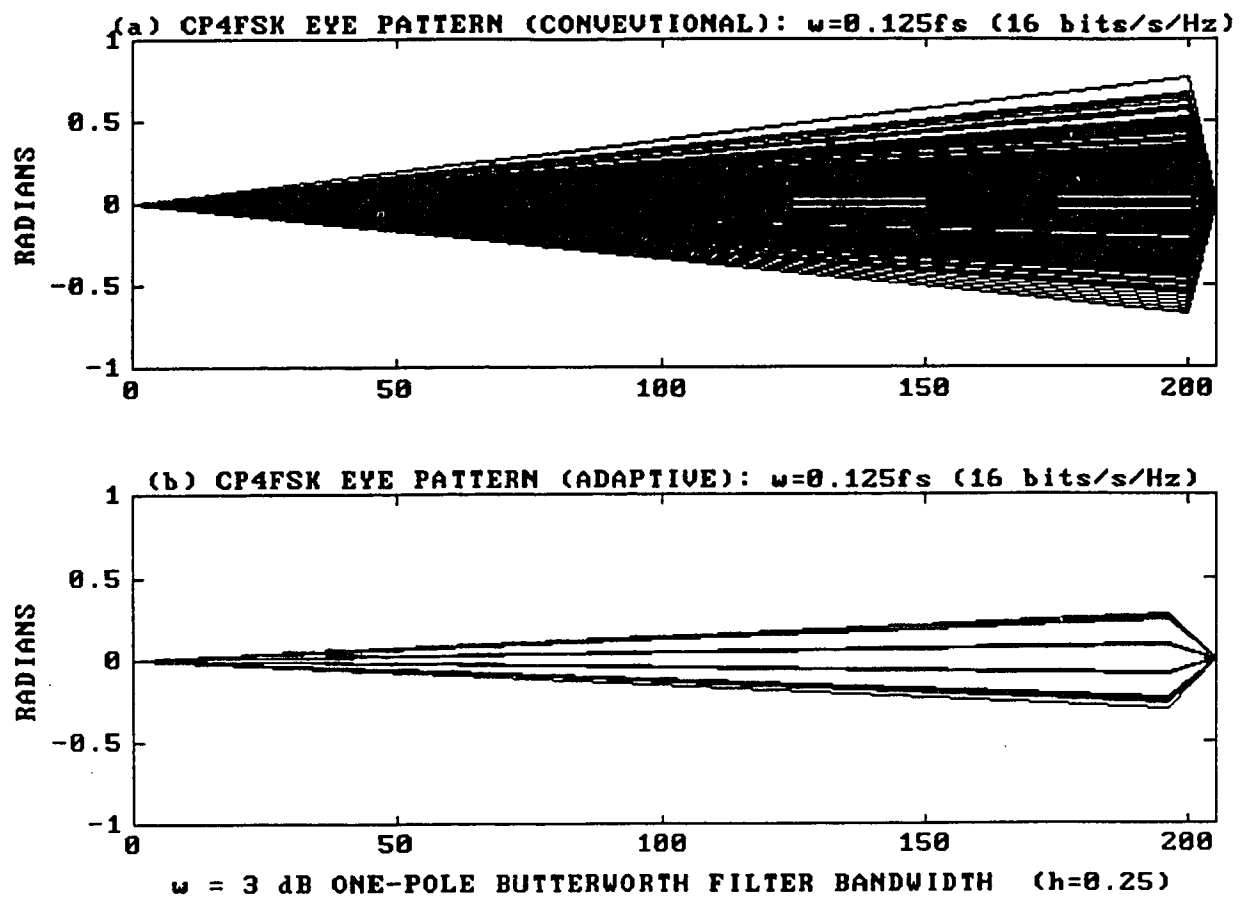


FIGURE 4.3.1: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = f_s$

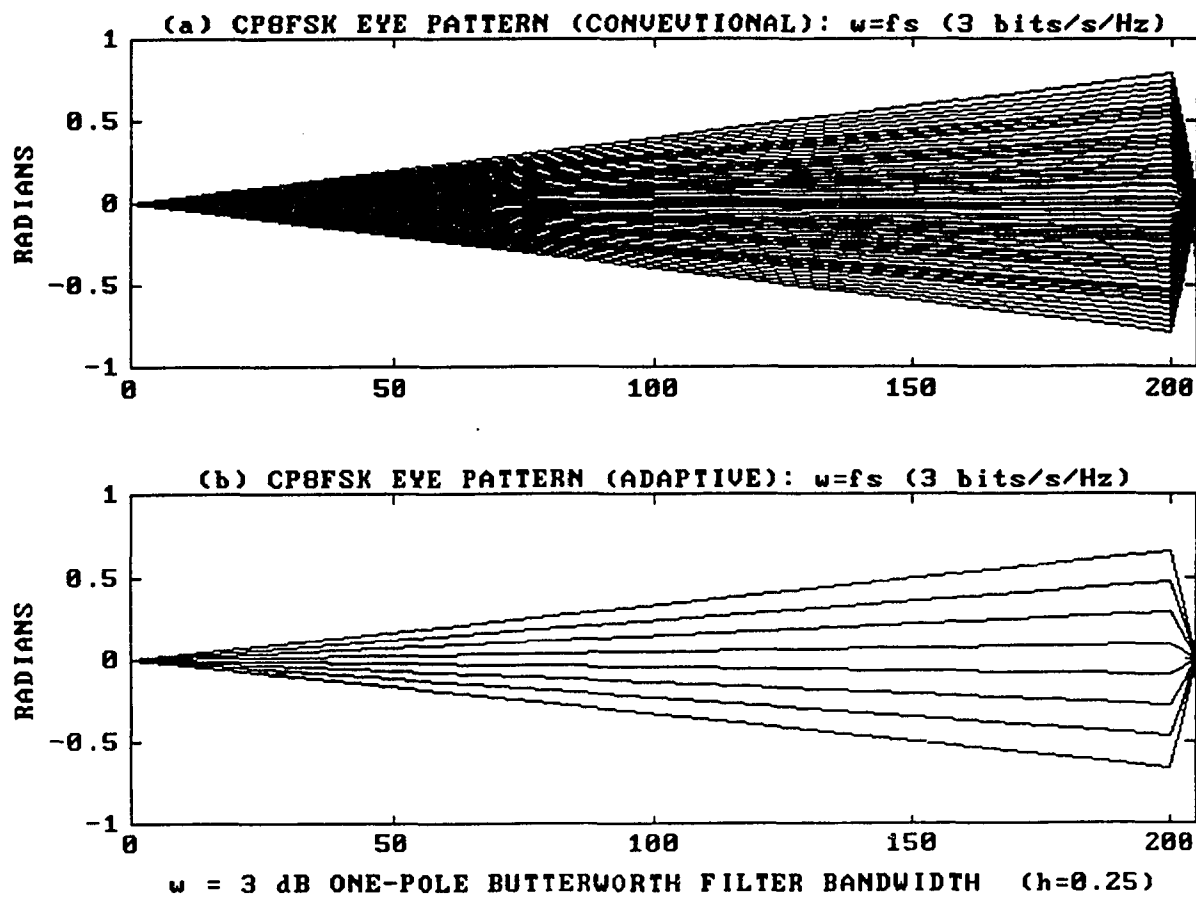


FIGURE 4.3.2: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$

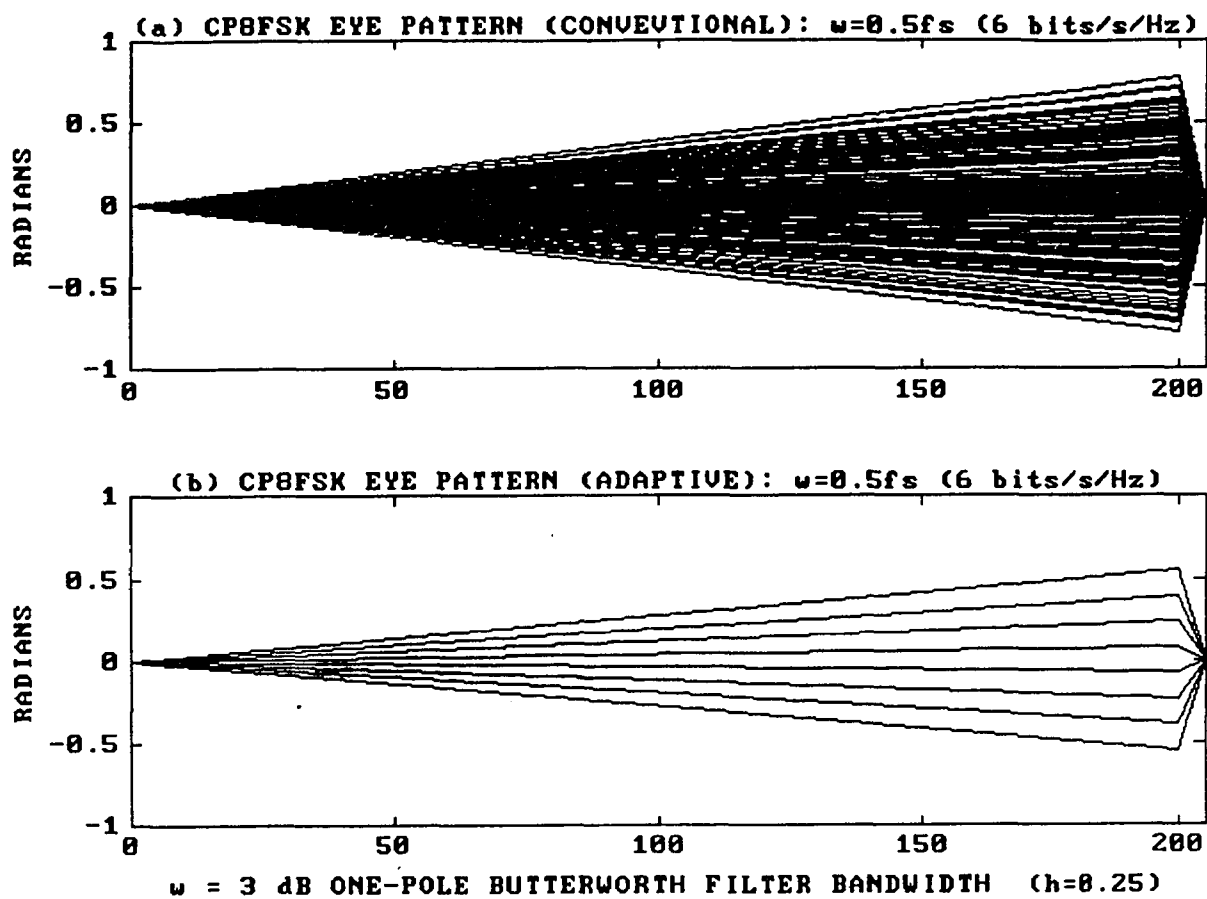


FIGURE 4.3.3: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

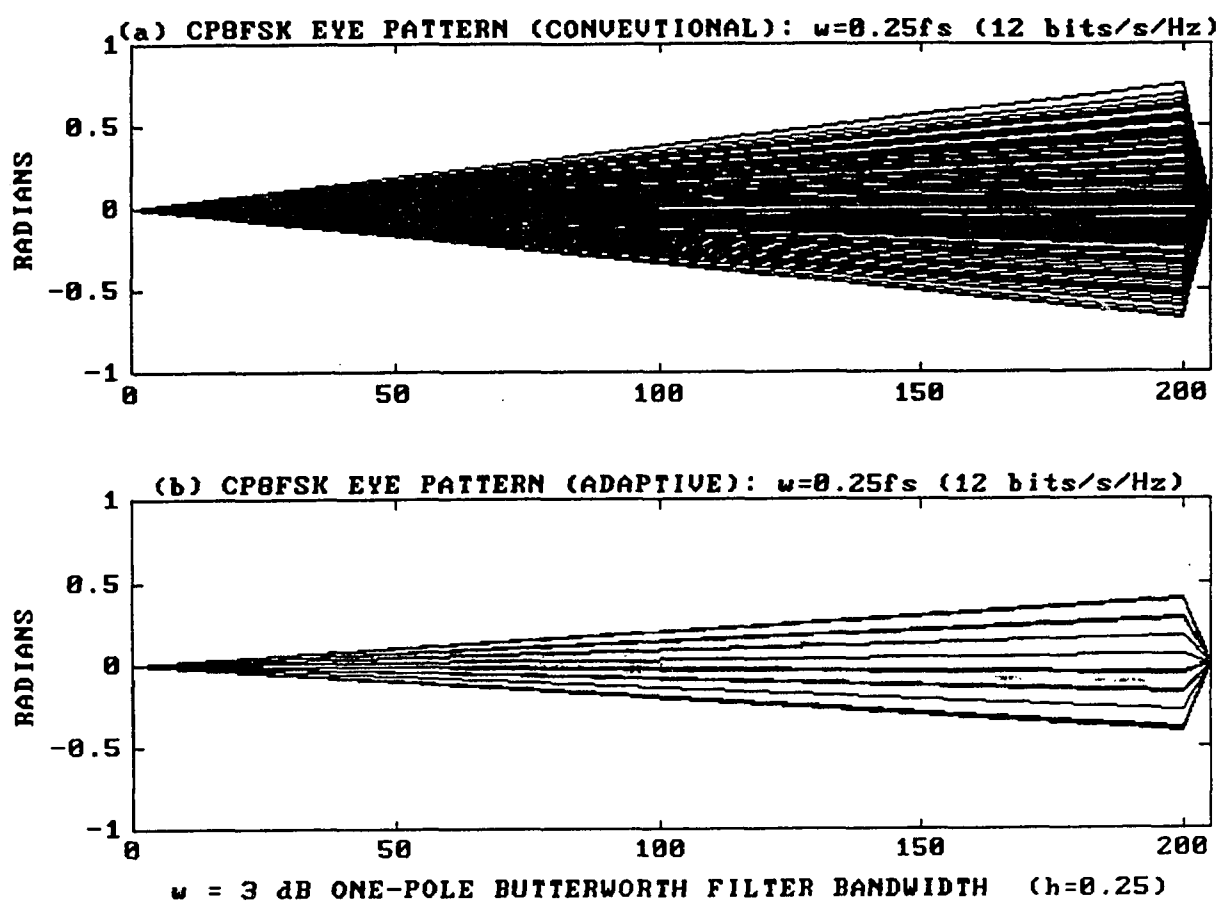


FIGURE 4.3.4: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_s$

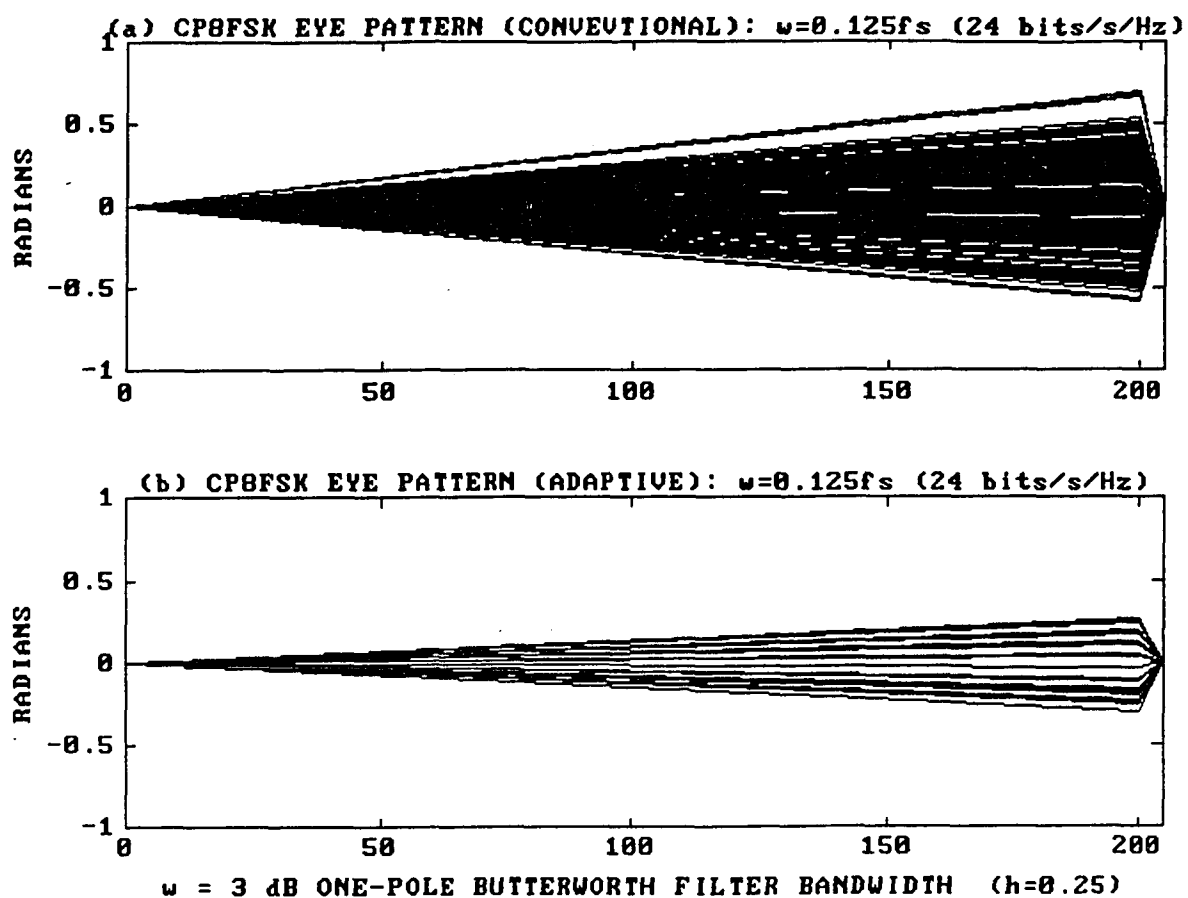


FIGURE 4.4.1: **CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = f_s$**

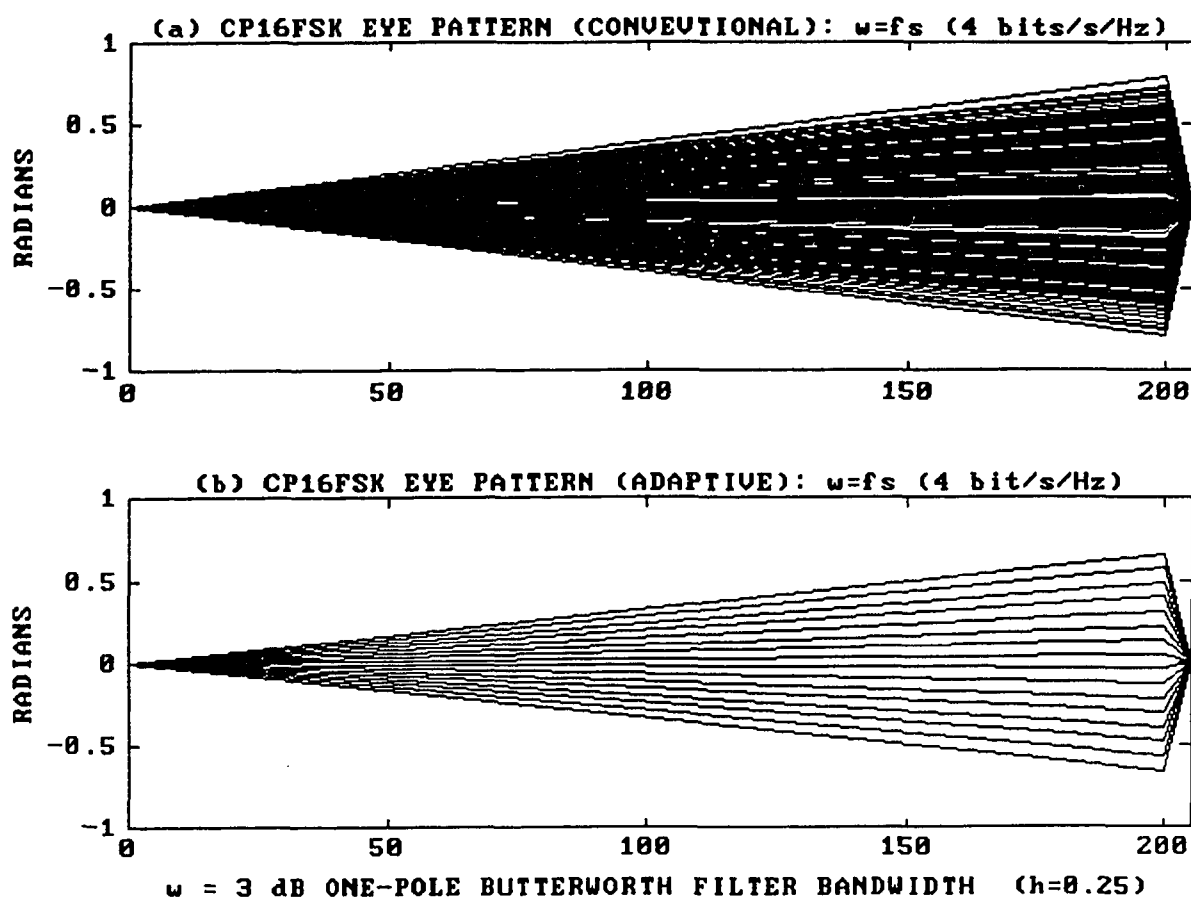


FIGURE 4.4.2: CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$

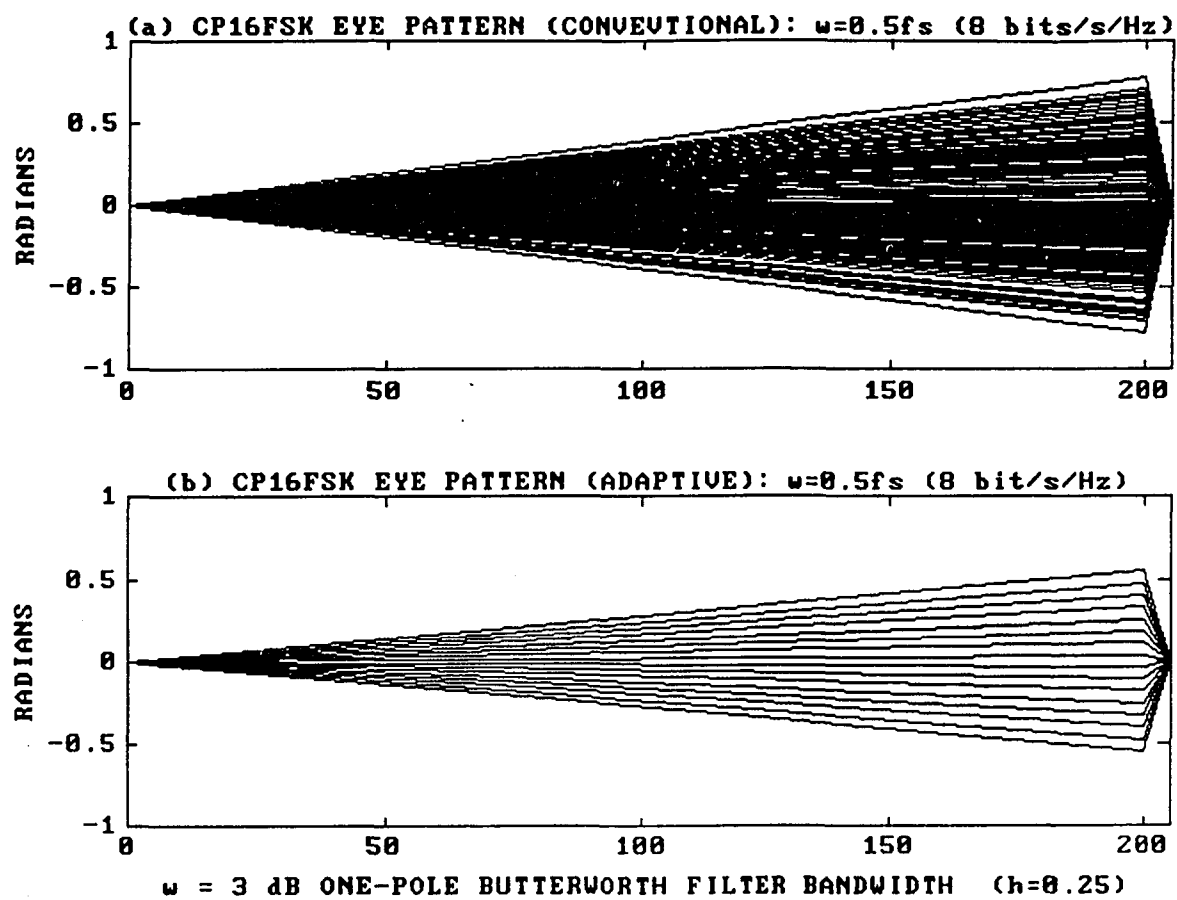


FIGURE 4.4.3: CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

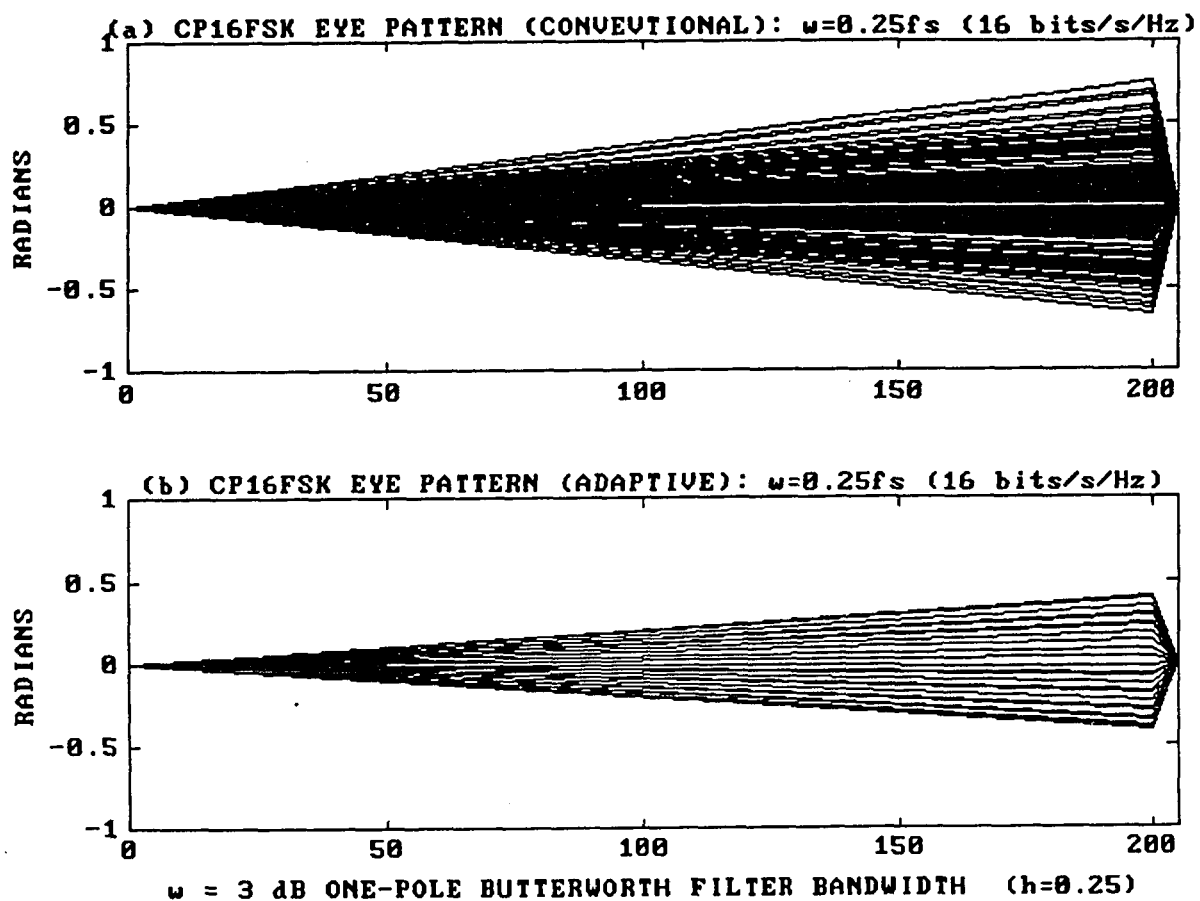
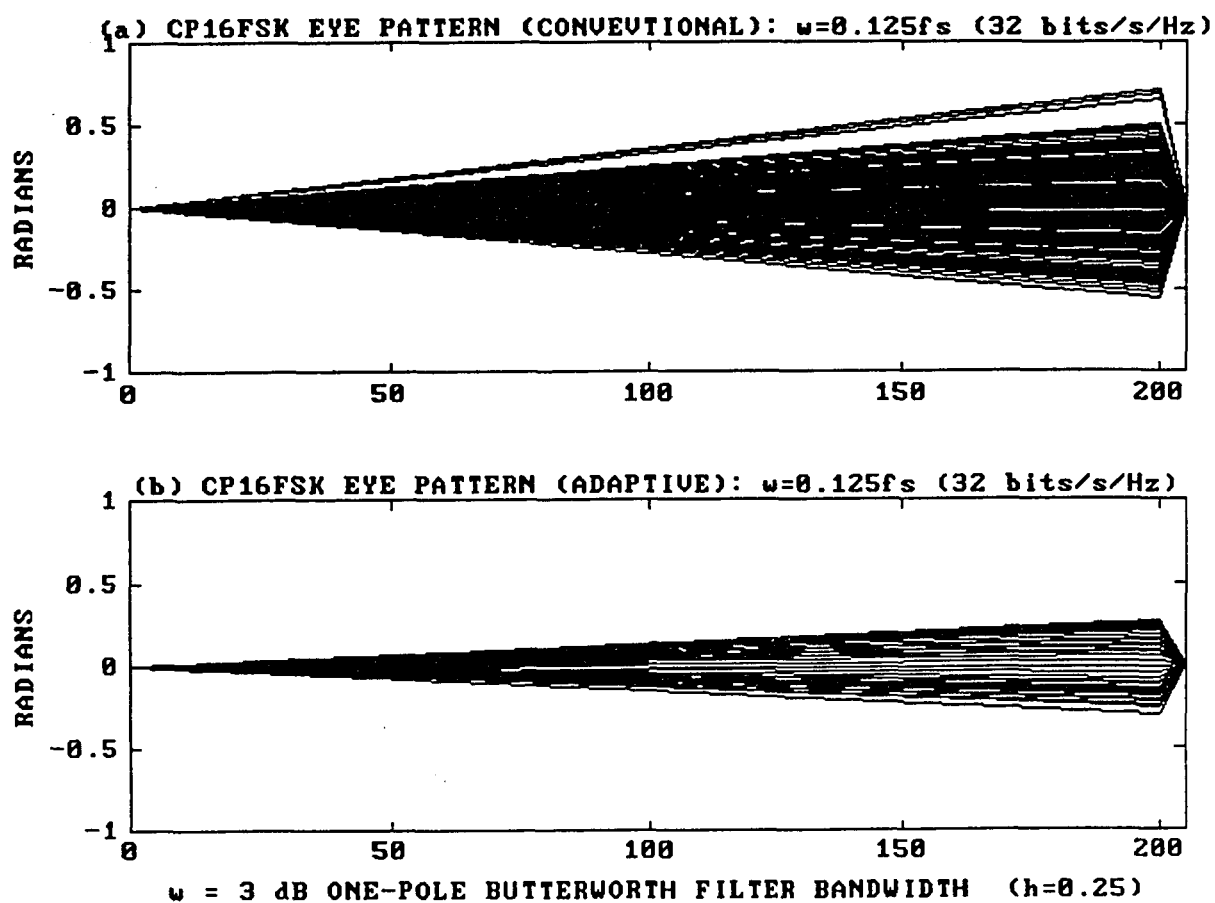


FIGURE 4.4.4: **CP16FSK EYE PATTERNS** FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_s$



be increased.

4.1.2 TWO-POLE BUTTERWORTH FILTER

To see the sensitivity of the proposed system to the bandlimiting filter, the results are also obtained for a 2nd order Butterworth filter. See Fig. 4.5 through Fig. 4.8 When adaptive detector is used, the eye is open if

$$f_b/w \leq (32/M) \log_2 M \quad \text{bits/s/Hz} \quad (4.2)$$

4.2 NOISE EFFECT

In this section performance of system is evaluated when the received signal is corrupted by white Gaussian noise. The effect of noise on the eye pattern of Fig. 4.1.2 is shown in Fig. 4.9 when 1st order Butterworth filter is used to detect *BF SK* signal. The eye is still open with 10 dB noise added.

Figure 4.10 shows a comparison of BER performance for the adaptive decision method and conventional threshold decision method for a CPBFSK when the signal is filtered by a first order Butterworth filter with k bits/s/Hz bandwidth efficiency where $k = 2, 4, 8$. Figure 4.10.1 shows that at The SNR of 12.4 dB where $P_e = 10^{-3}$ dB,

FIGURE 4.5.1: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = f_b$**

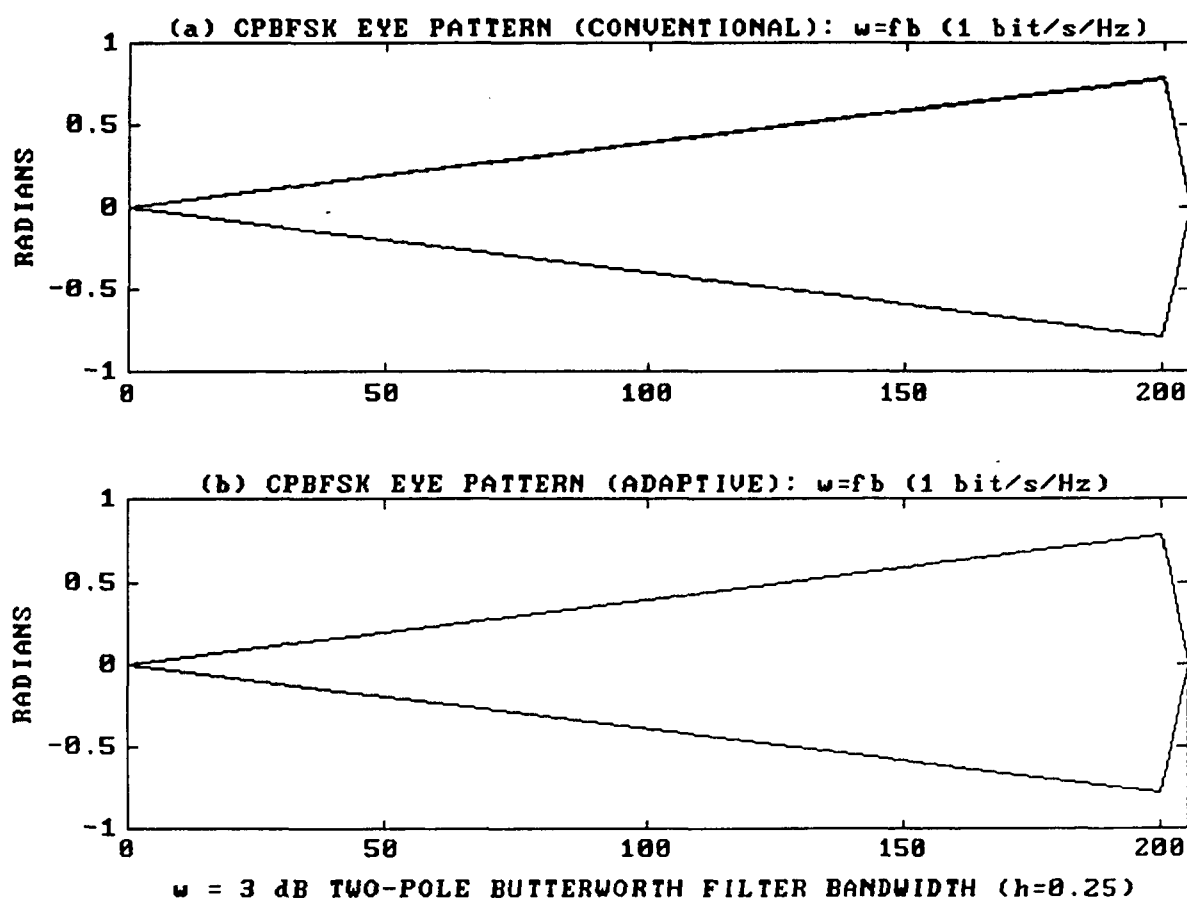


FIGURE 4.5.2: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$

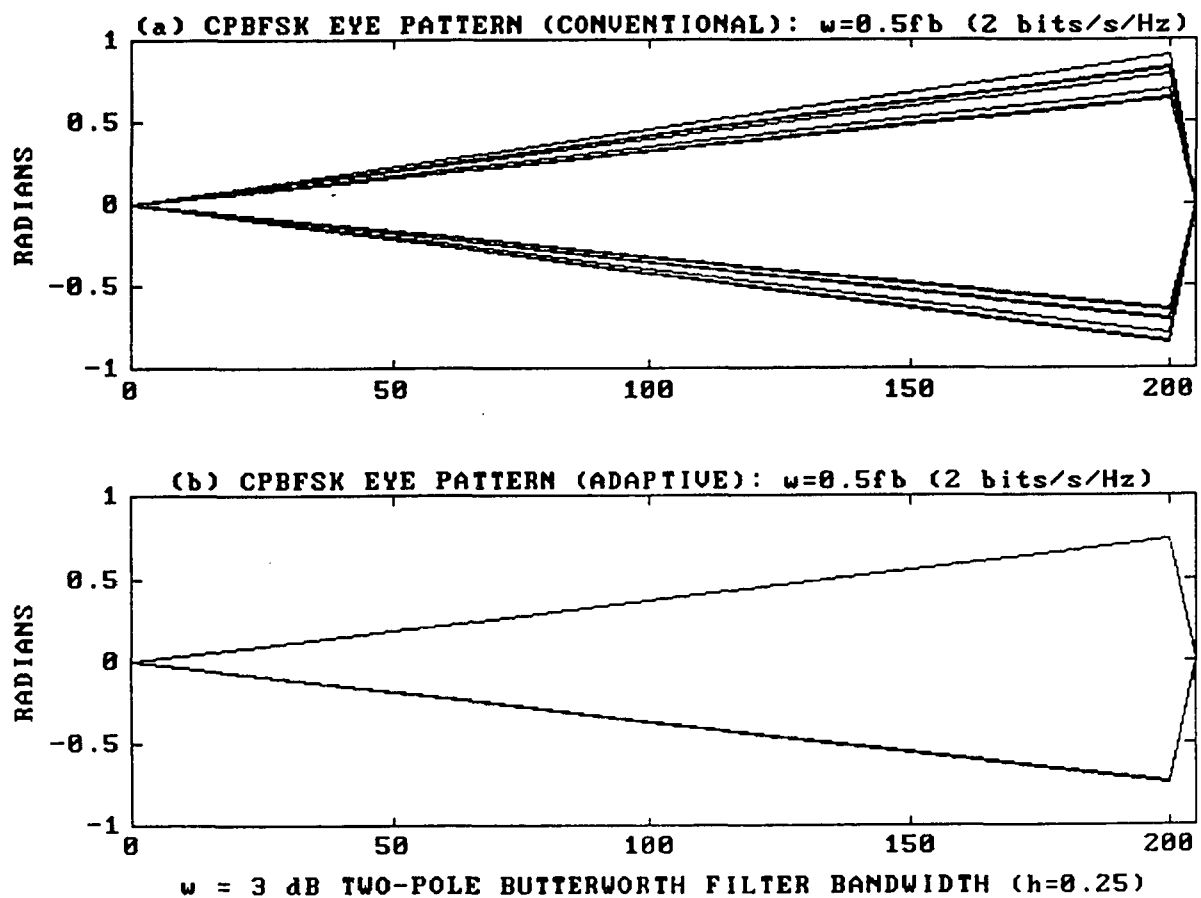


FIGURE 4.5.3: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_b$**

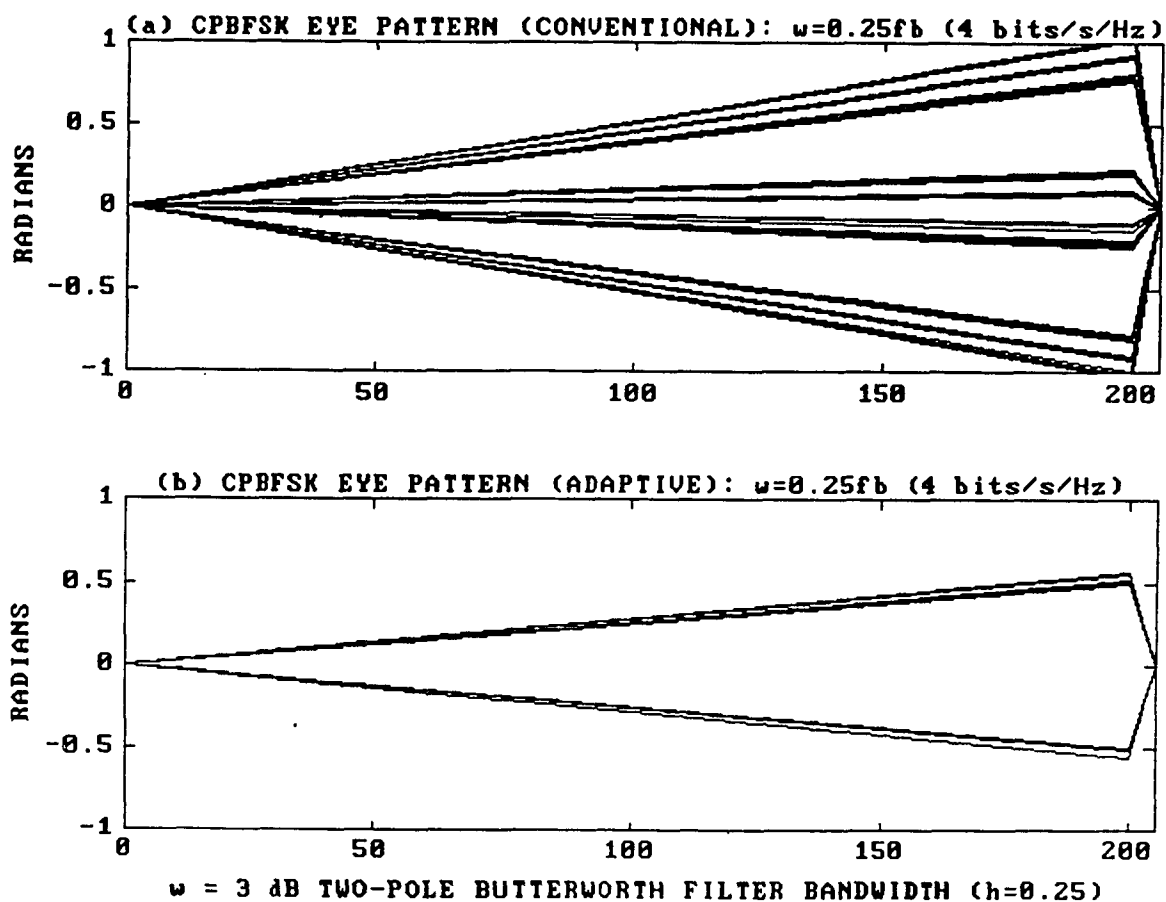


FIGURE 4.5.4: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_b$**

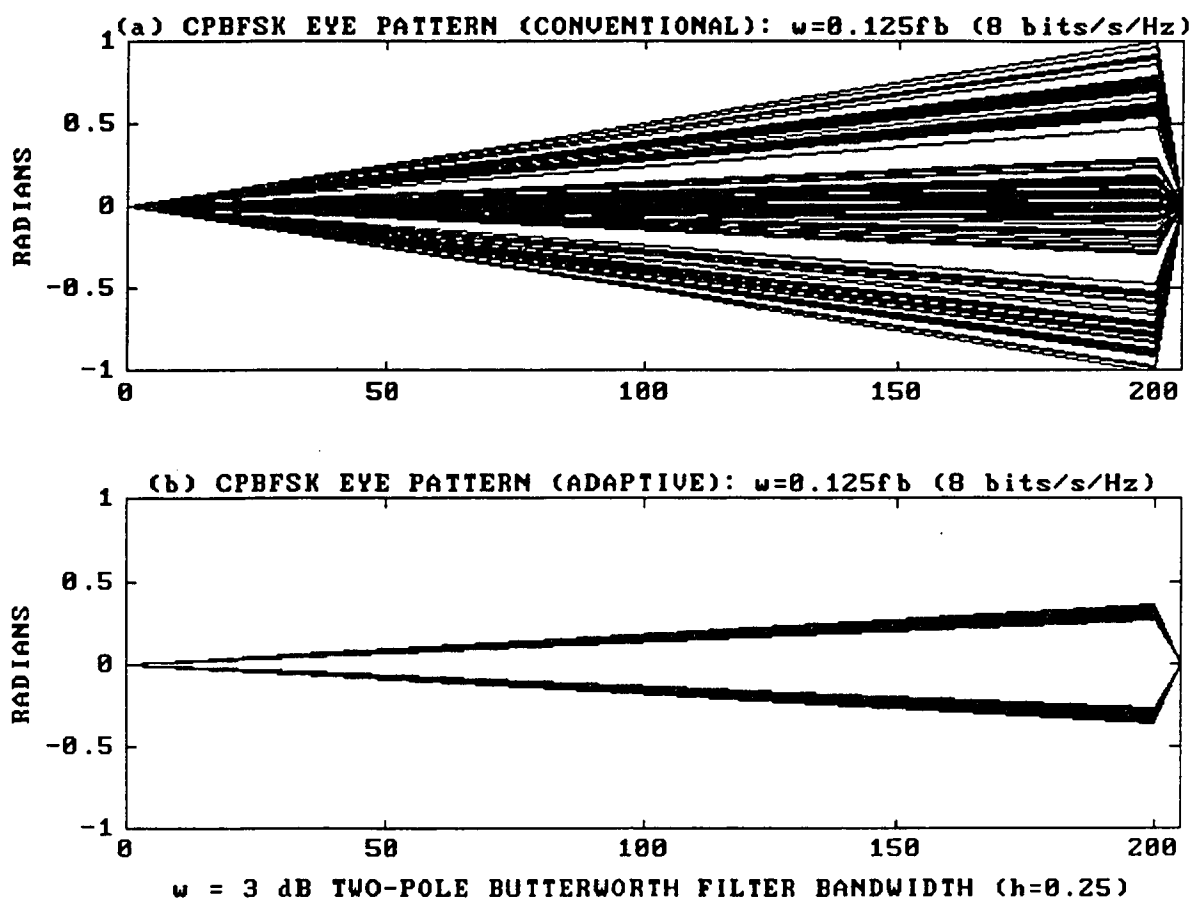


FIGURE 4.6.1: **CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = f_s$**

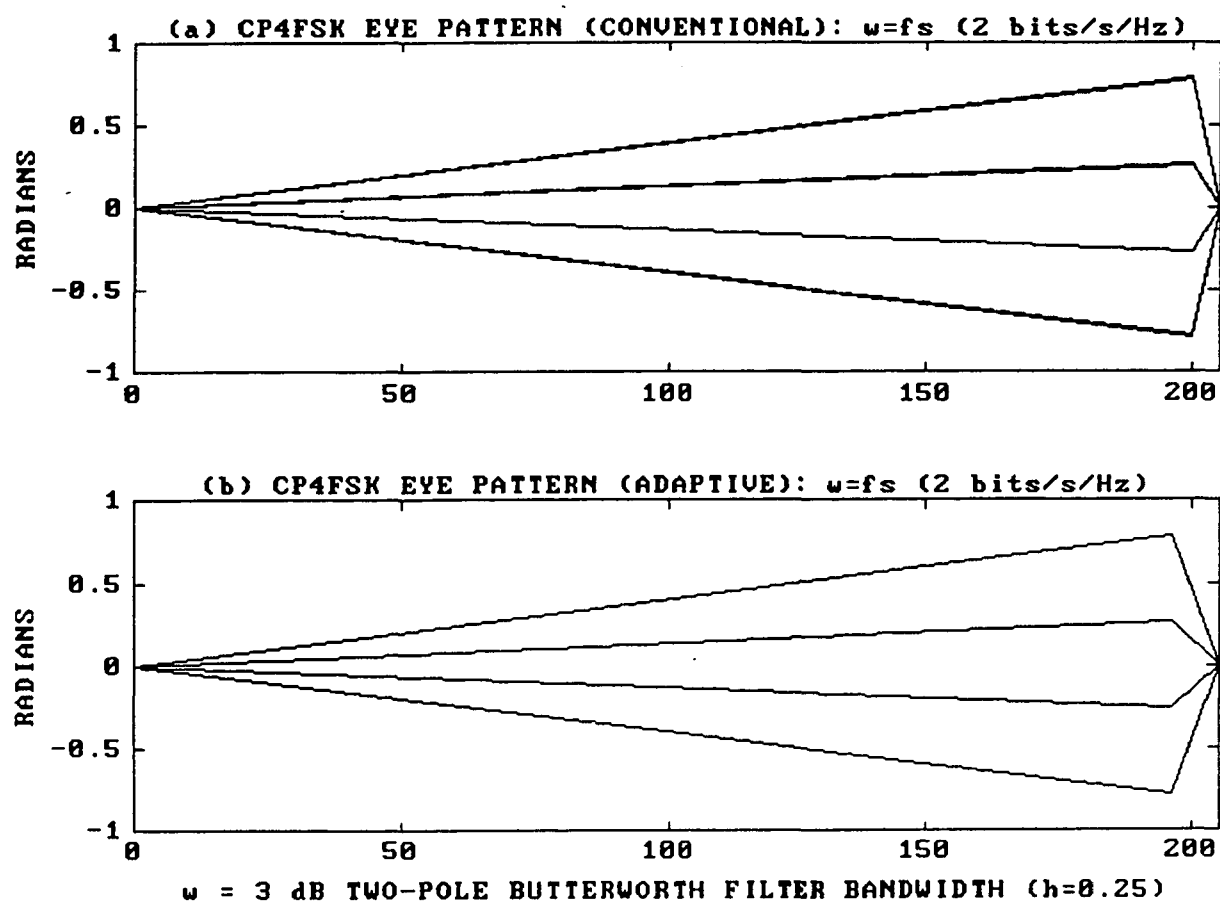


FIGURE 4.6.2: CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$

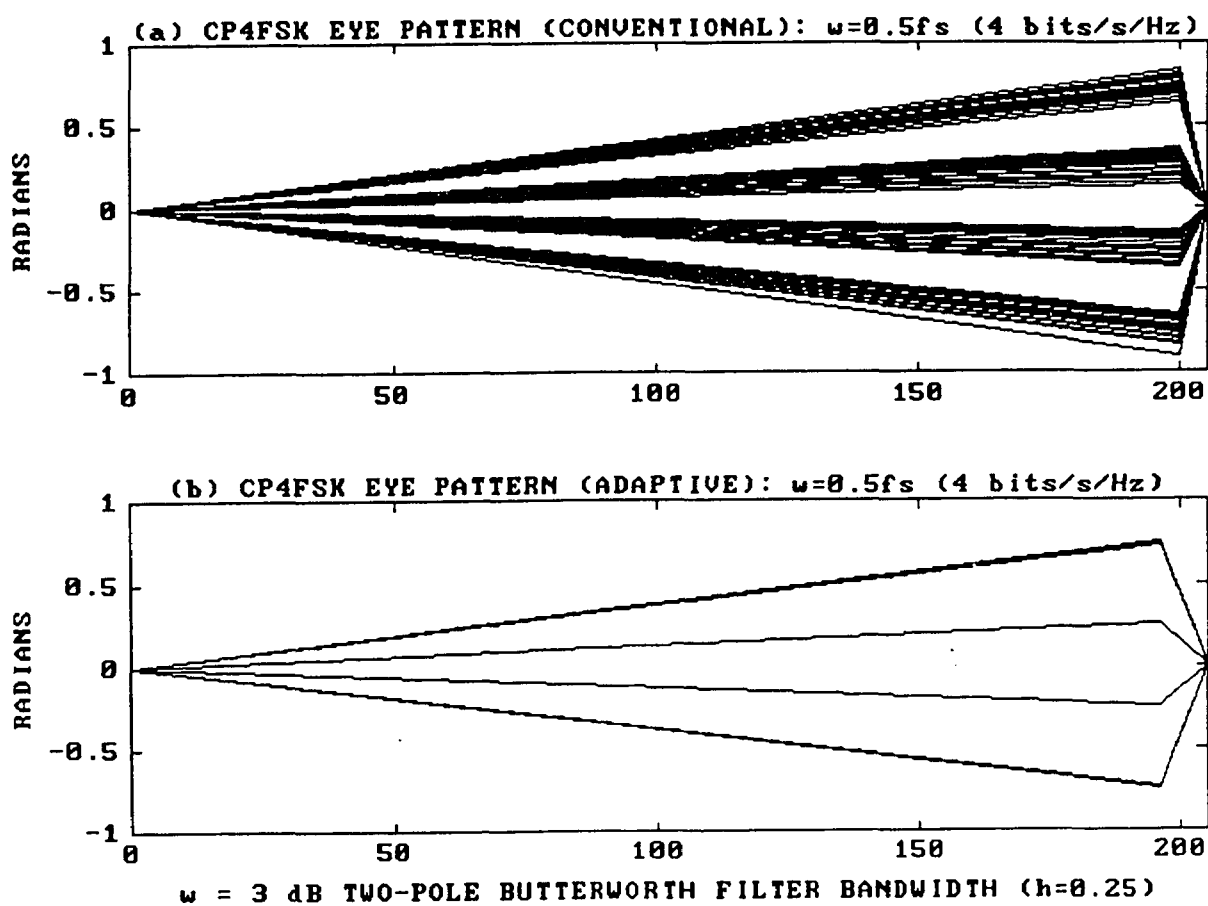


FIGURE 4.6.3: **CP4FSK EYE PATTERNS** FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

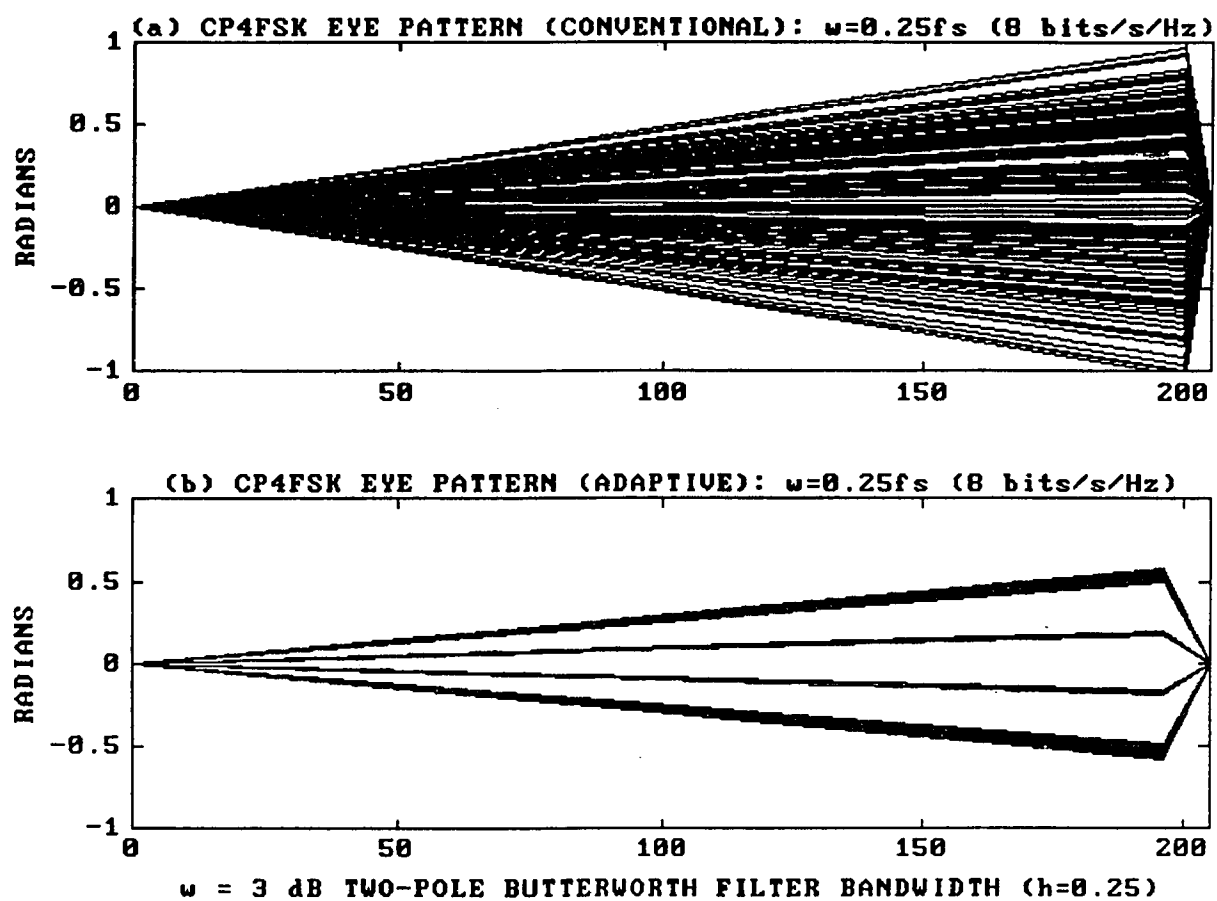


FIGURE 4.6.4: **CP4FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_s$**

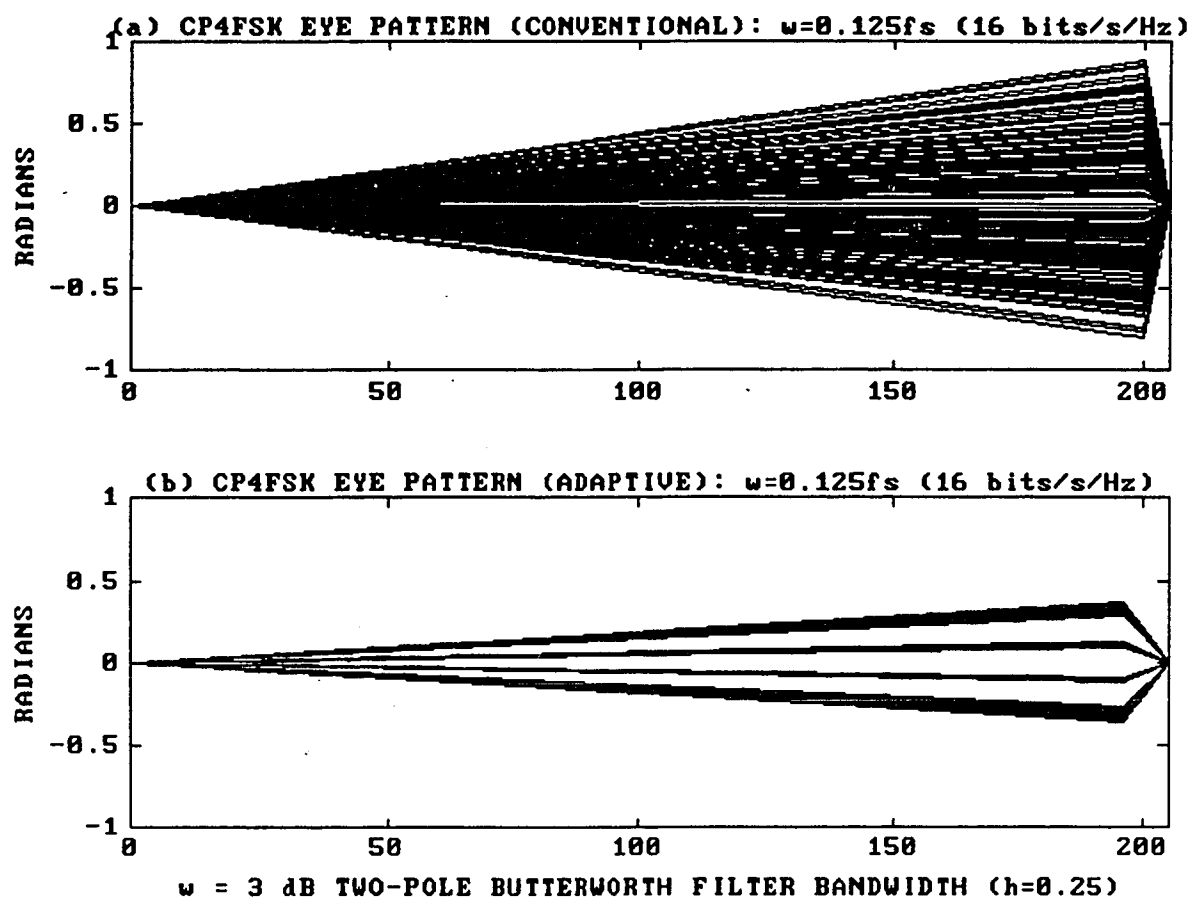


FIGURE 4.7.1: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = f_s$

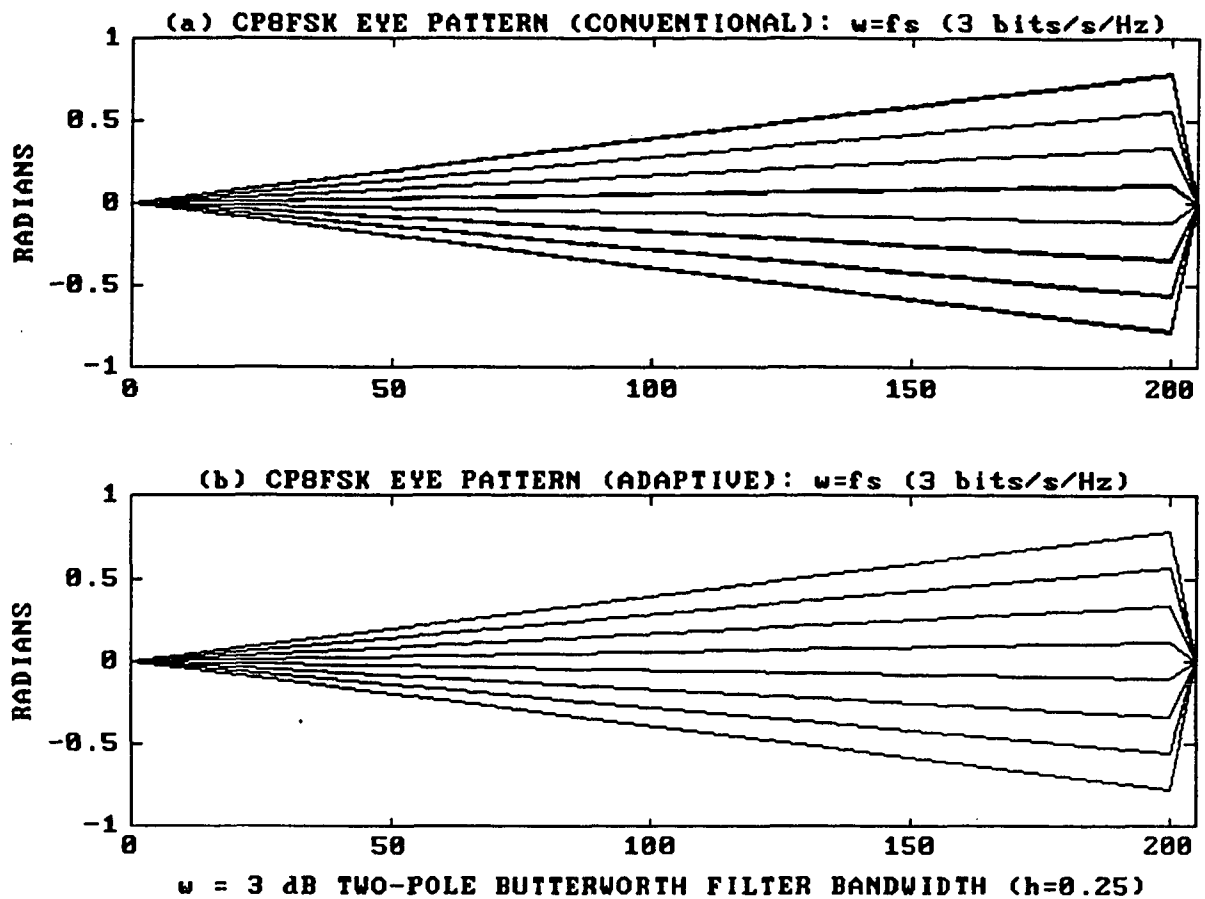


FIGURE 4.7.2: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$

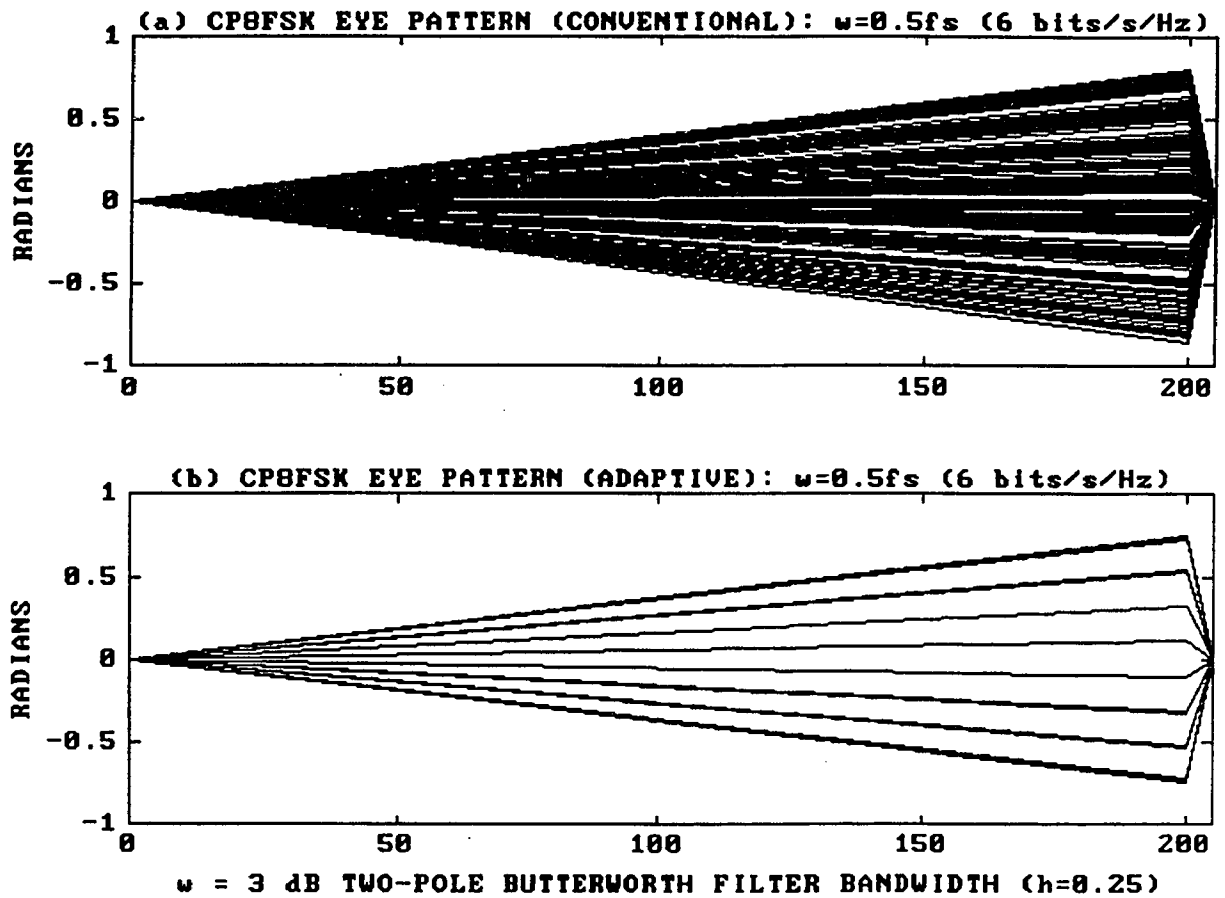


FIGURE 4.7.3: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

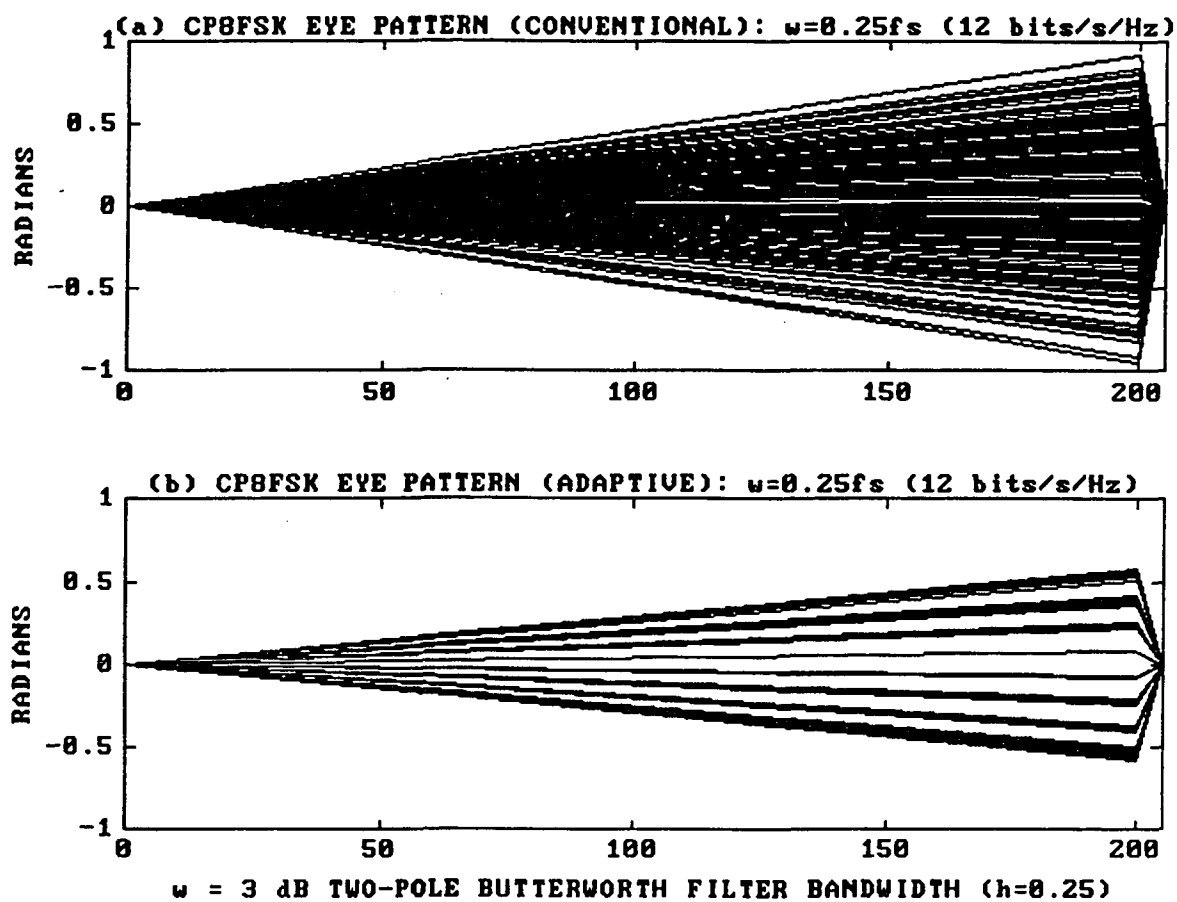


FIGURE 4.7.4: CP8FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.125 f_s$

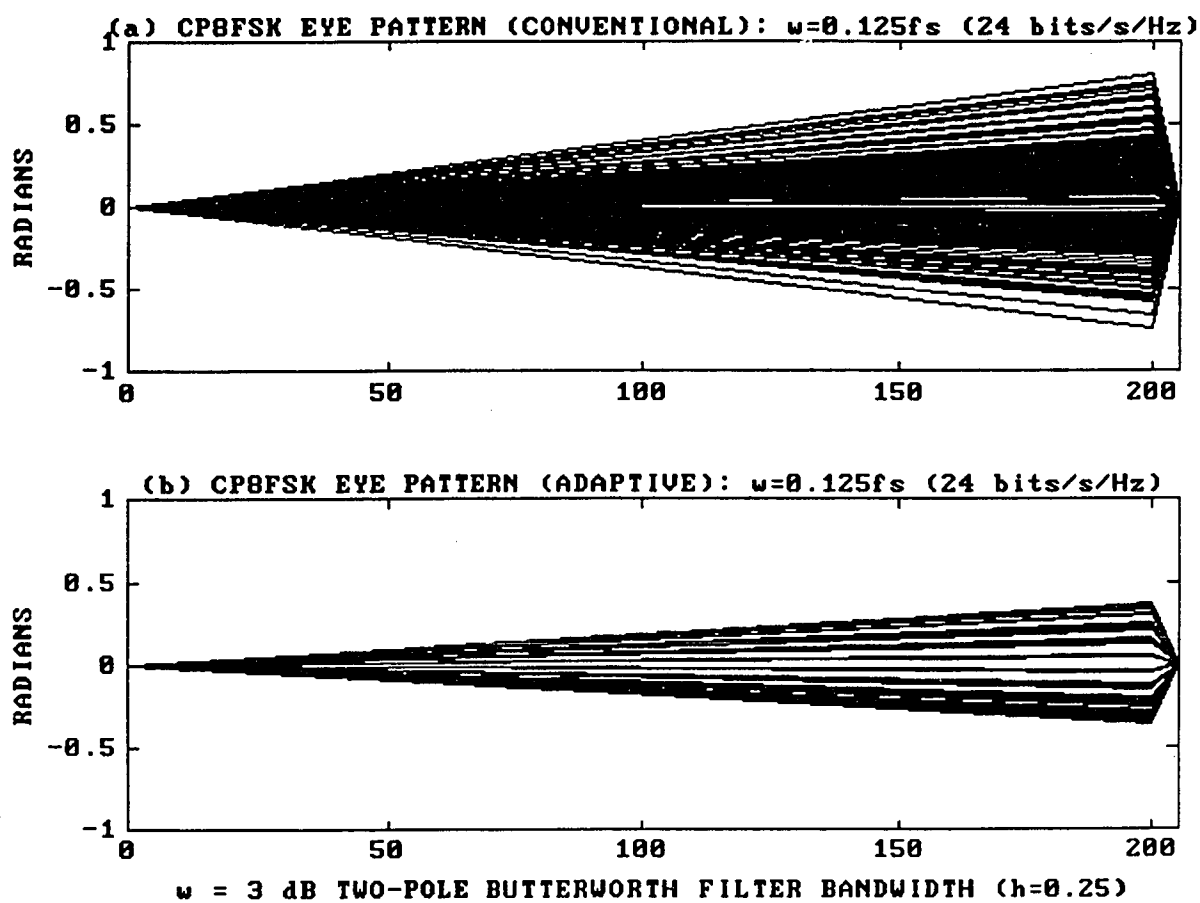


FIGURE 4.8.1: **CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = f_s$**

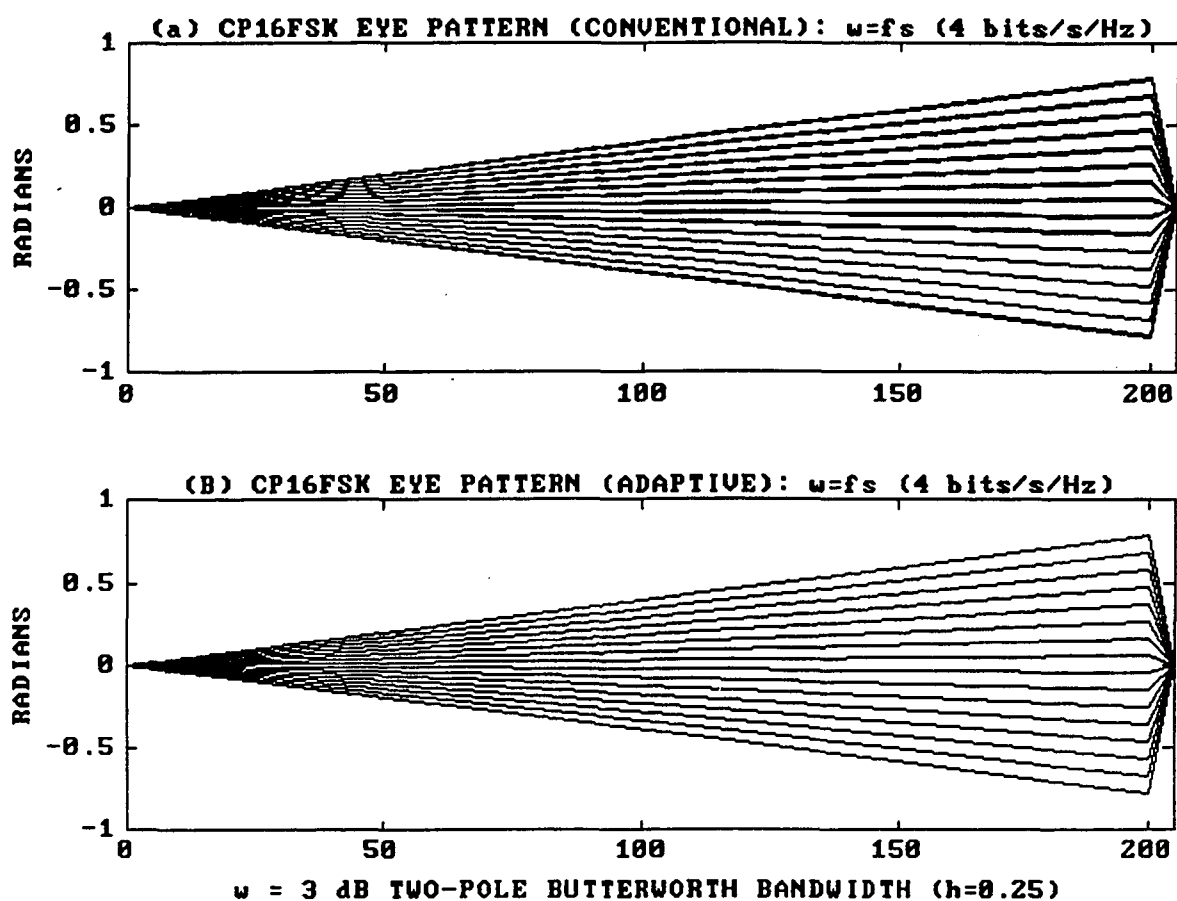


FIGURE 4.8.2: **CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_s$**

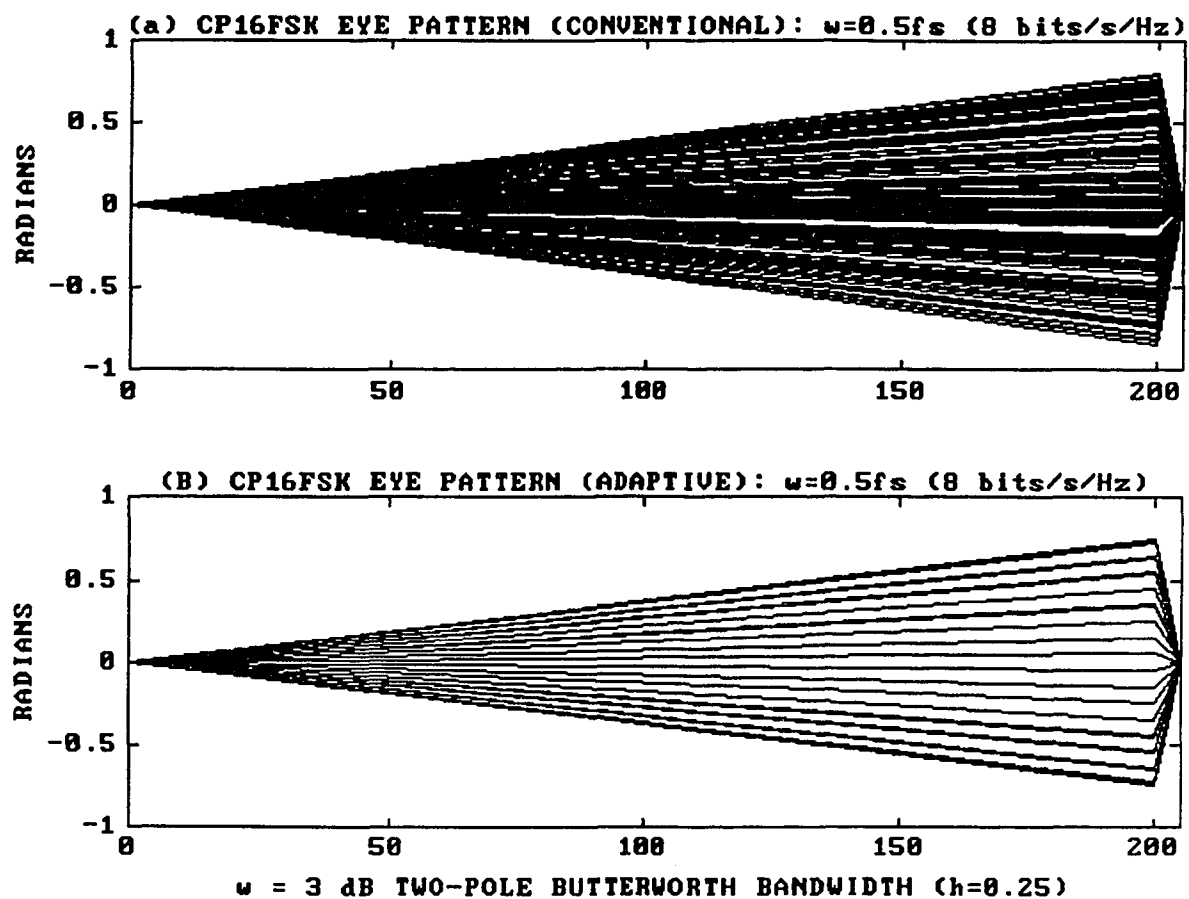


FIGURE 4.8.3: CP16FSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) USING A TWO-POLE BUTTERWORTH FILTER WITH $w = 0.25 f_s$

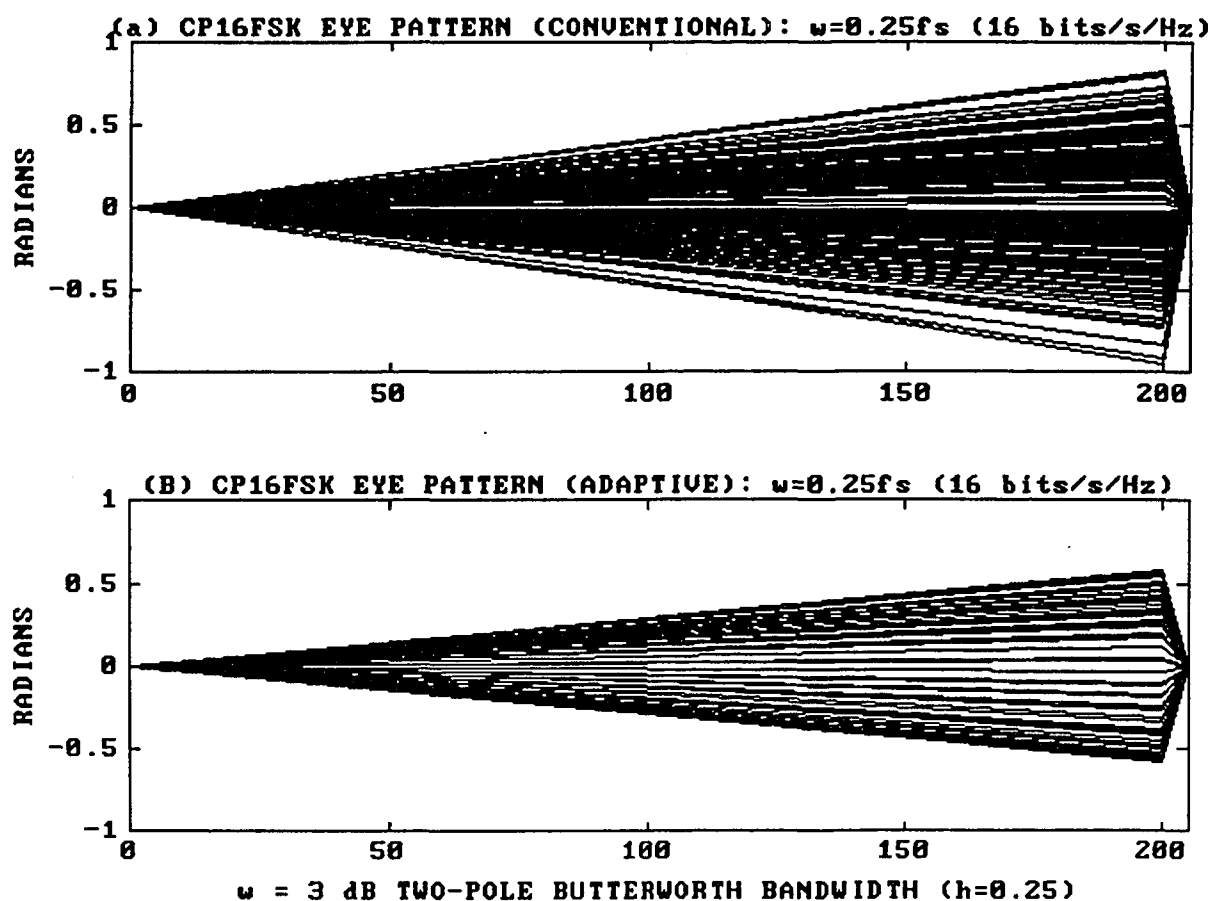


FIGURE 4.9.1: CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) WHEN 25 dB NOISE IS BEING ADDED USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$

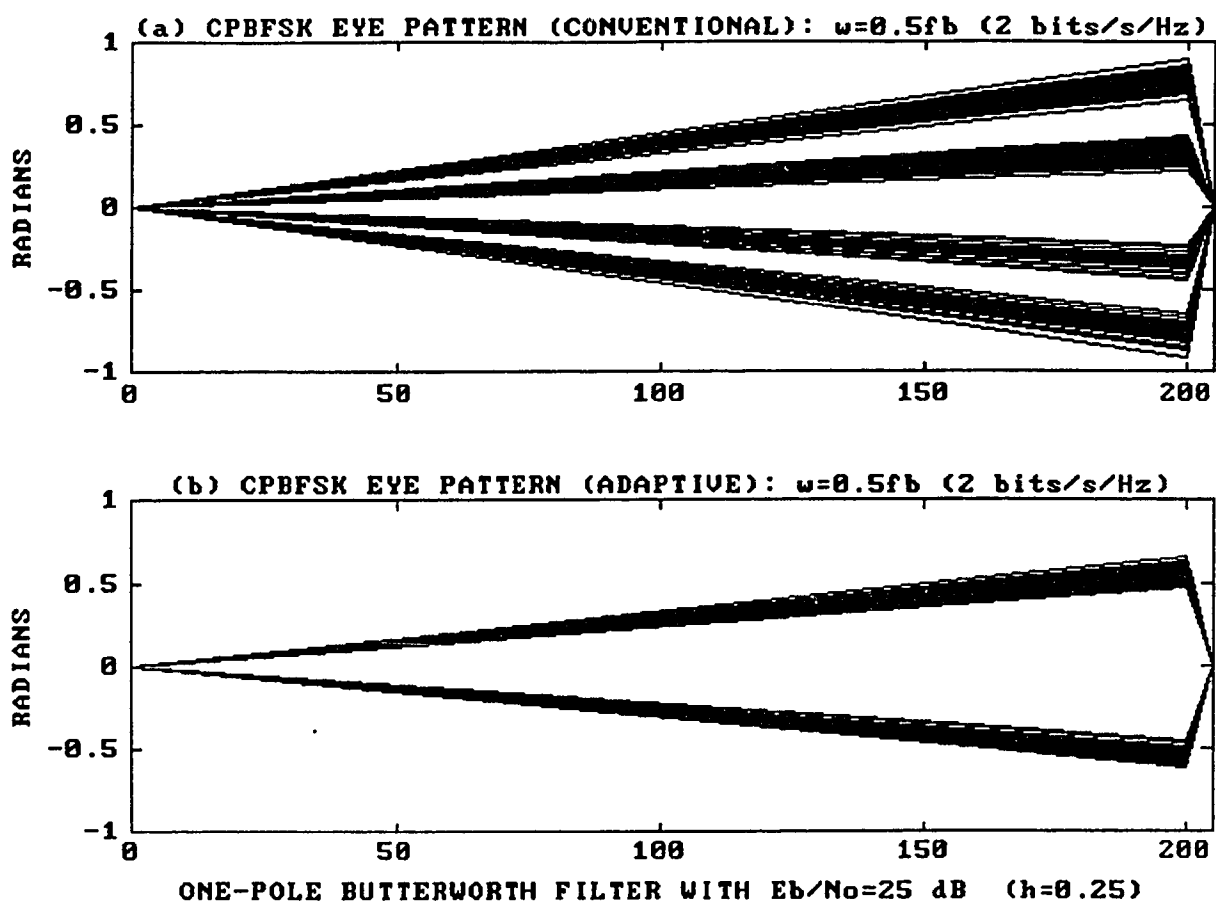


FIGURE 4.9.2: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) WHEN 20 dB NOISE IS BEING ADDED USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$**

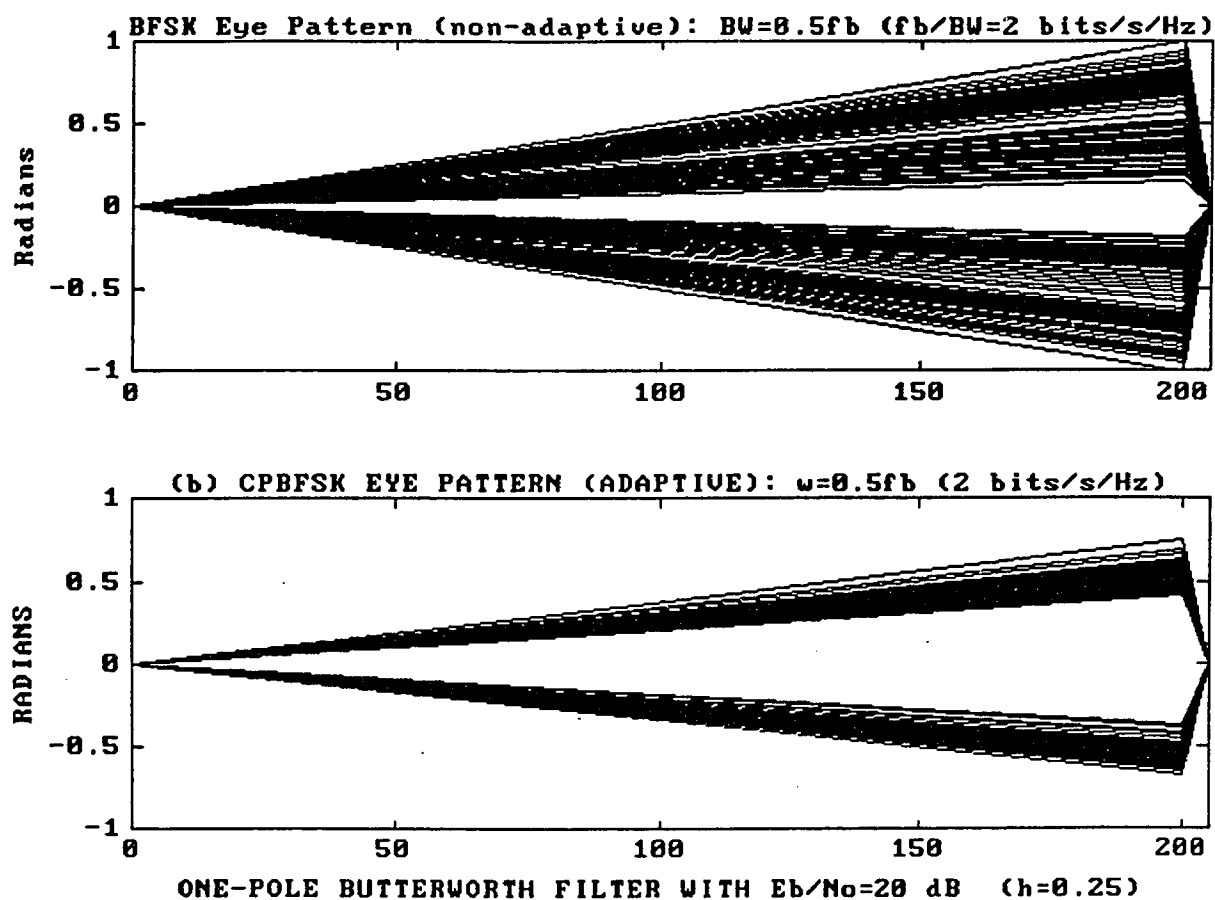


FIGURE 4.9.3: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) WHEN 15 dB NOISE IS BEING ADDED USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$**

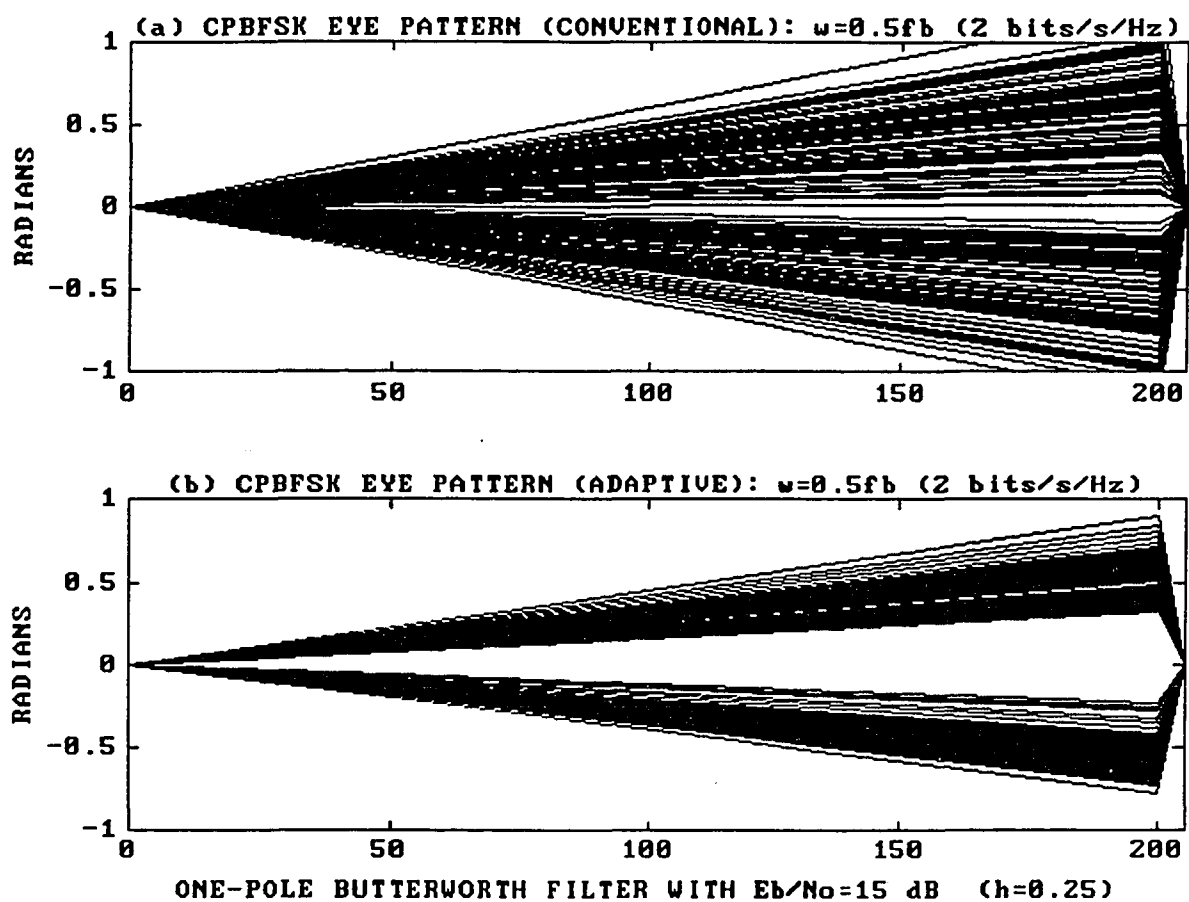
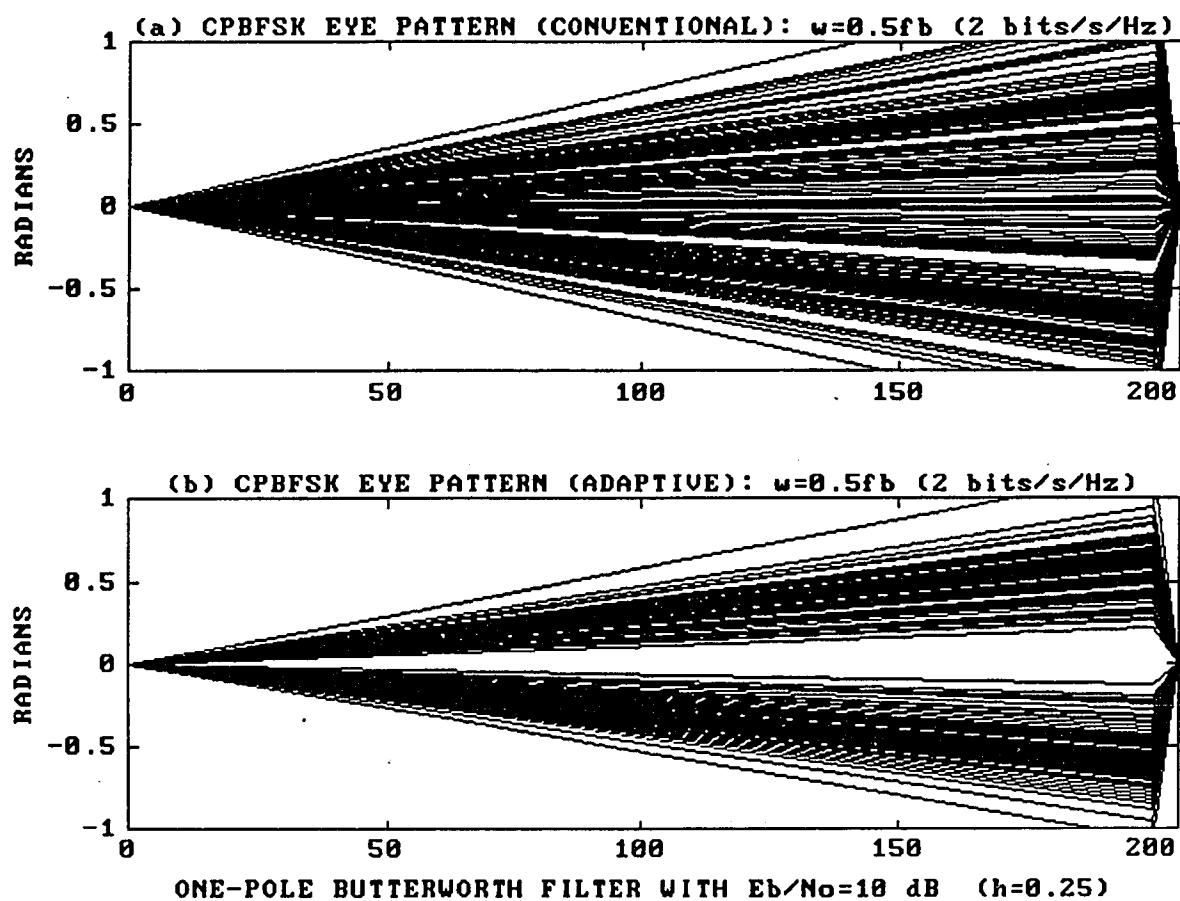


FIGURE 4.9.4: **CPBFSK EYE PATTERNS FOR THE CONVENTIONAL (a) AND ADAPTIVE THRESHOLD DECISION METHODS (b) WHEN 10 dB NOISE IS BEING ADDED USING A ONE-POLE BUTTERWORTH FILTER WITH $w = 0.5 f_b$**



the improvement is more than 10 dB compared to conventional method. Figure 4.10.2 reflects that there is a difference of about 1dB of required SNR when bandwidth of IF filter is reduced to reach 4 bits/s/Hz bandwidth efficiency. Figures 4.11-4.13 show the performance of the system when 4-ary, 8-ary and 16-ary continuous phase FSK signals has been utilized, respectively. BER performance of CPMFSK is depicted in Fig. 4.14 for the bandwidth efficiency of k bits/s/Hz when $k = 2, 4$ and 8 . These results demonstrate that the improved probability of error can be reached when adaptive threshold is applied.

Figures 4.15-4.18 present the comparison of BER performance for the conventional threshold (a) and adaptive threshold method (b) for binary, 4-ary, 8-ary and 16-ary signals when Two-pole Butterworth filter is employed for 2, 4 and 8 bits/s/Hz spectral efficiency. Figure 4.19 shows the system performance when CPMFSK signal with bandwidth efficiency of k bits/s/Hz is utilized where $k = 2, 4$ and 8 .

4.3 UNKNOWN BANDLIMITING FILTER

Although we have been concerned with the case in which the spectrum of the transmitted signal is restricted to a specific band, however, the band limiting filter is considered to be part of the receiver and the results are applicable to the case where part of the spectrum limitation occurs in a transmitting filter. In this section we have first assumed that only the bandwidth of the bandlimiting filter which is part of transmitter, is unknown by the nonlinear post-processor. We have also assumed

that the IF filter bandwidth is wide enough to have no effect on distorted signal phase. A computer program presented in appendix is written to track an lock on the correct bandwidth using adaptive steps. This program estimates the bandwidth by computing the eye opening and selecting the bandwidth which produces the largest eye. The eye diagrams found to be the same as the results presented in section 4.1 when above assumptions were made.

Next we assumed the type of the bandlimiting filter is unknown by the adaptive detector. In this example we assumed that the transmitting filter is a Two-pole Butterworth filter while the nonlinear post processor presumed the bandlimiting filter has only one pole. The BER results is presented in Fig. 4.20. Figure 4.20.1 shows the case when the postprocessor has no knowledge of the predemodulation filter (P21) versus the case when it knows the type of the IF filter (P2). To reach BER of 10^{-4} with 2 bits/s/Hz bandwidth efficiency, there is a difference of 0.4 dB in SNR when the adaptive post-processor has no knowledge of the bandlimiting filter type.

FIGURE 4.10.1: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

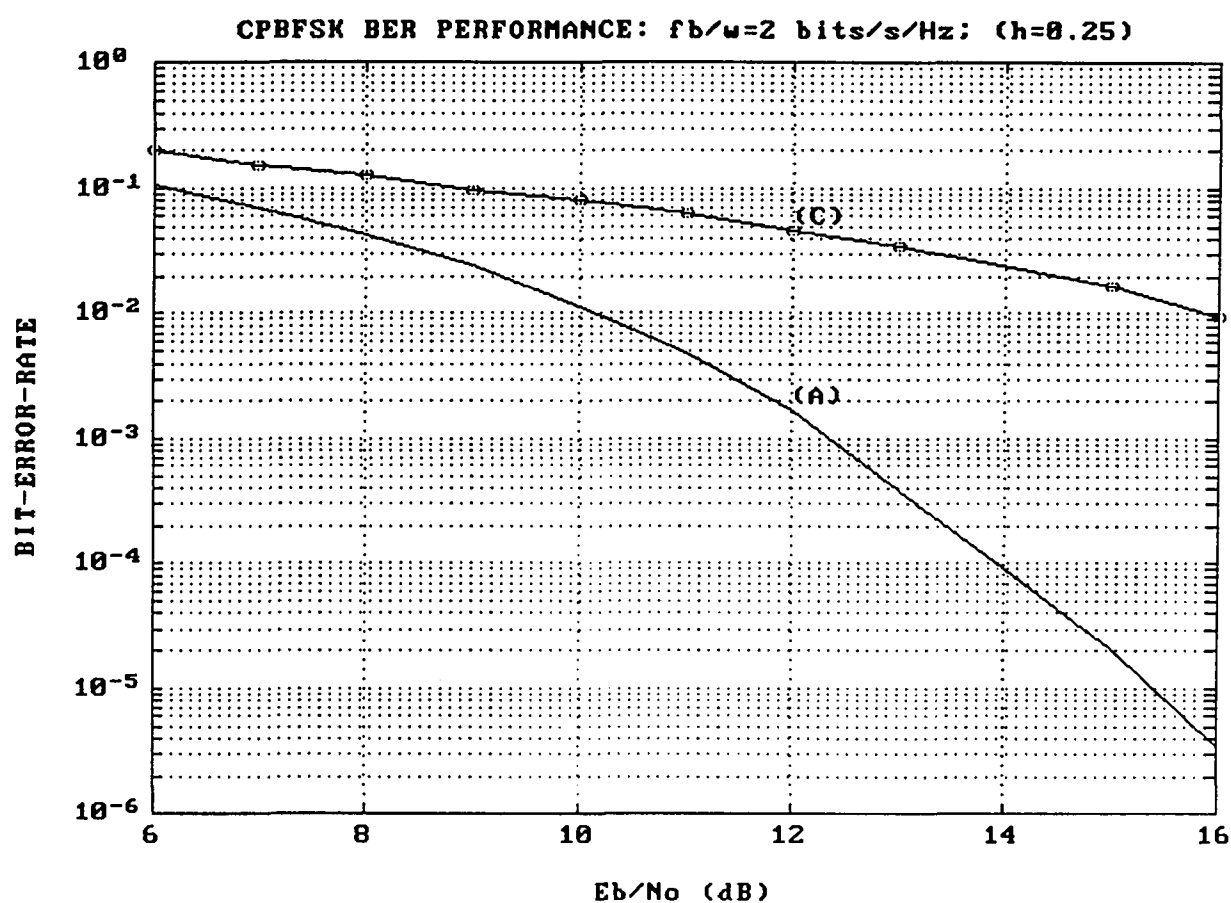


FIGURE 4.10.2: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

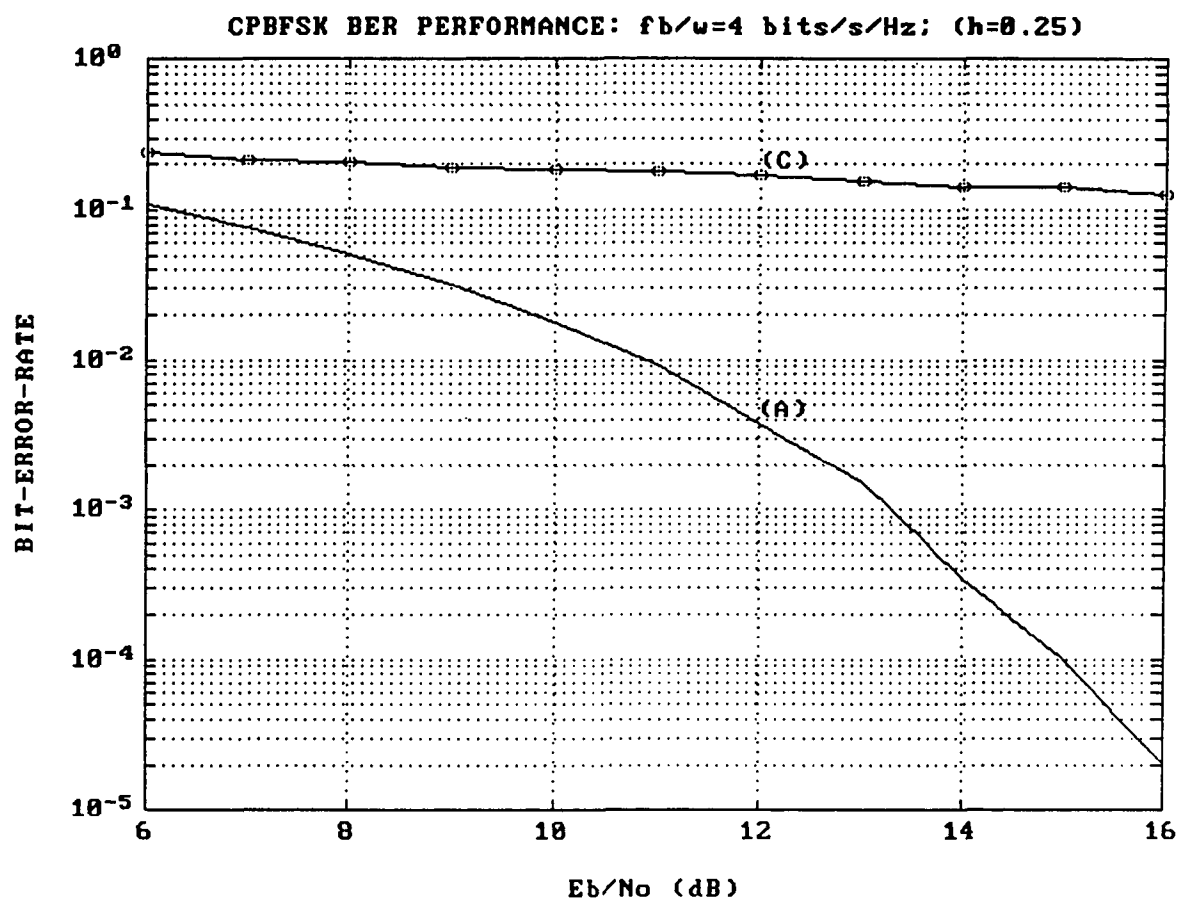


FIGURE 4.10.3: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

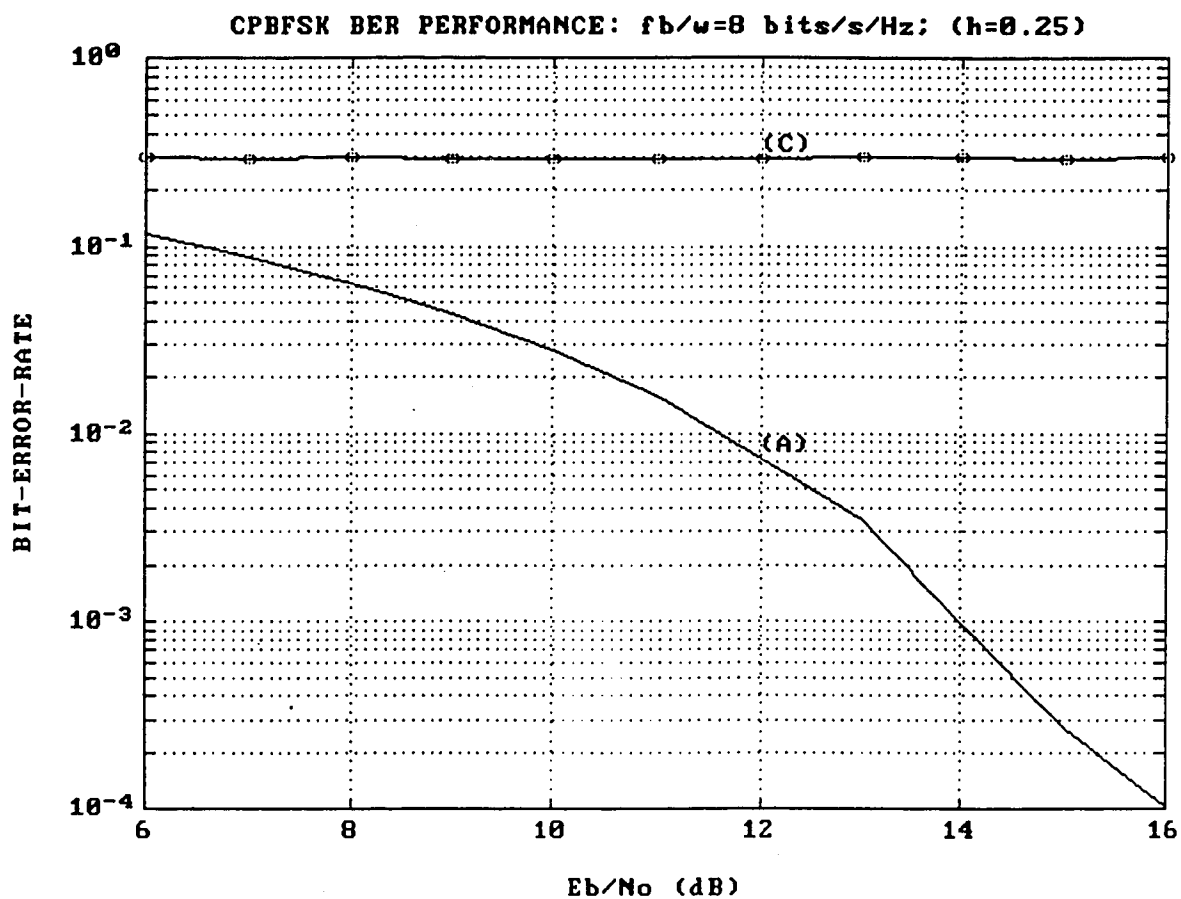


FIGURE 4.11.1: COMPARISON OF CP4FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

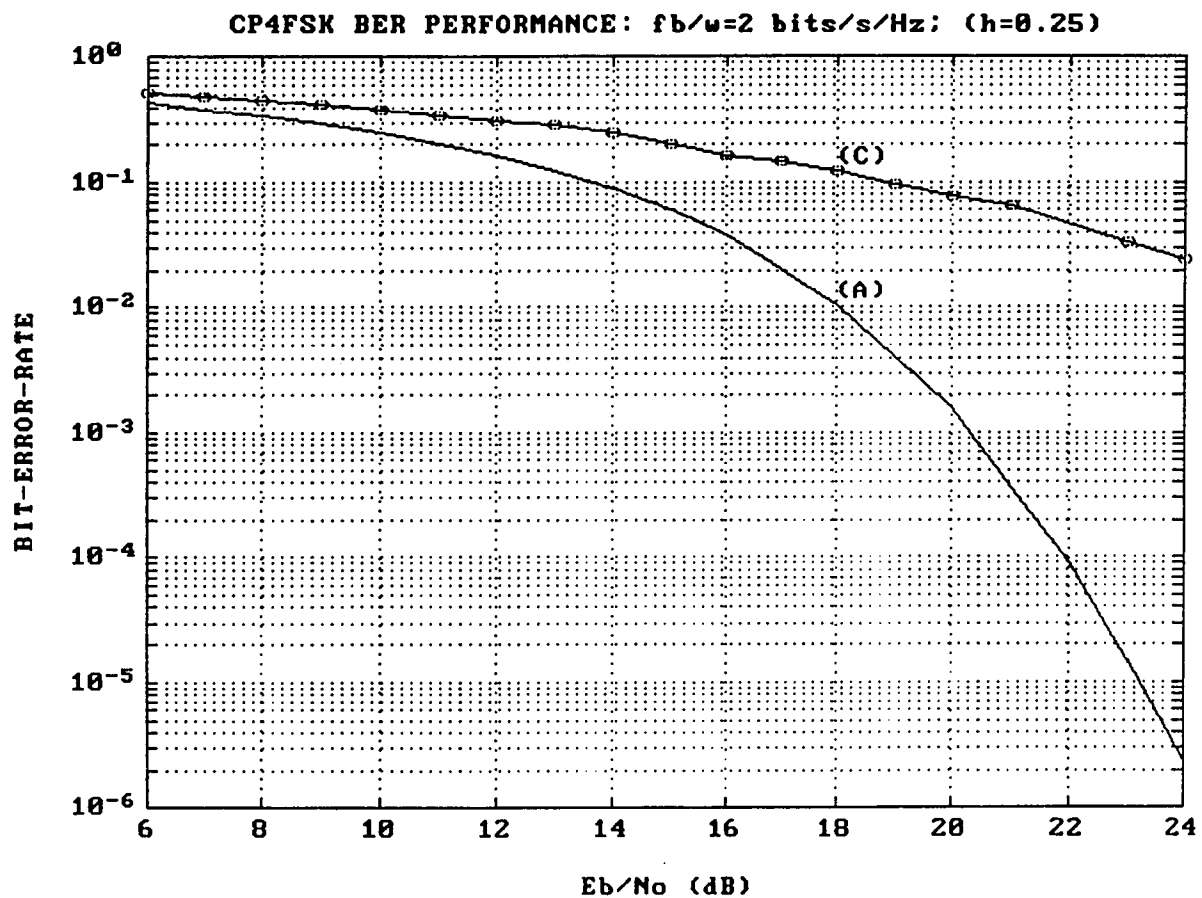


FIGURE 4.11.2: COMPARISON OF CP4FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

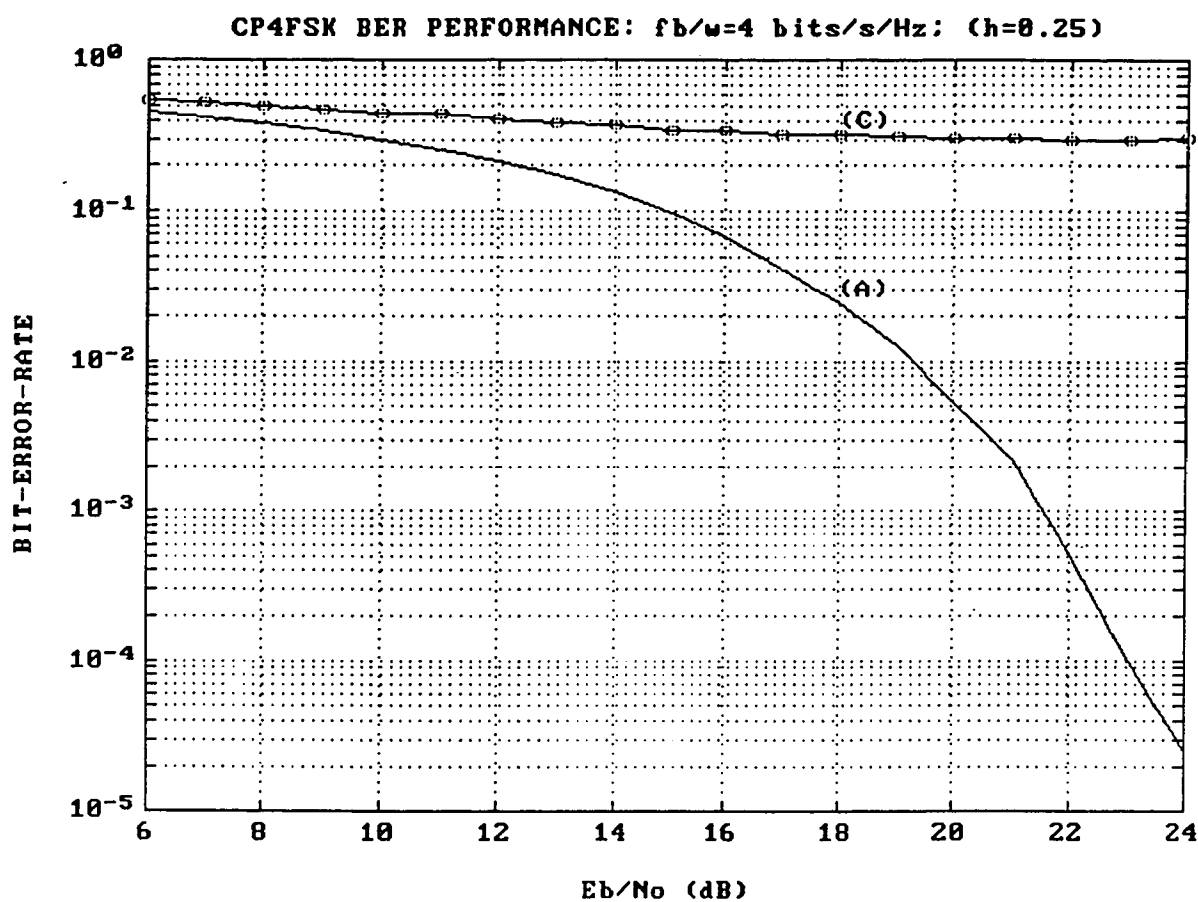


FIGURE 4.11.3: COMPARISON OF CP4FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/\omega = 8 \text{ bits/s/Hz}$

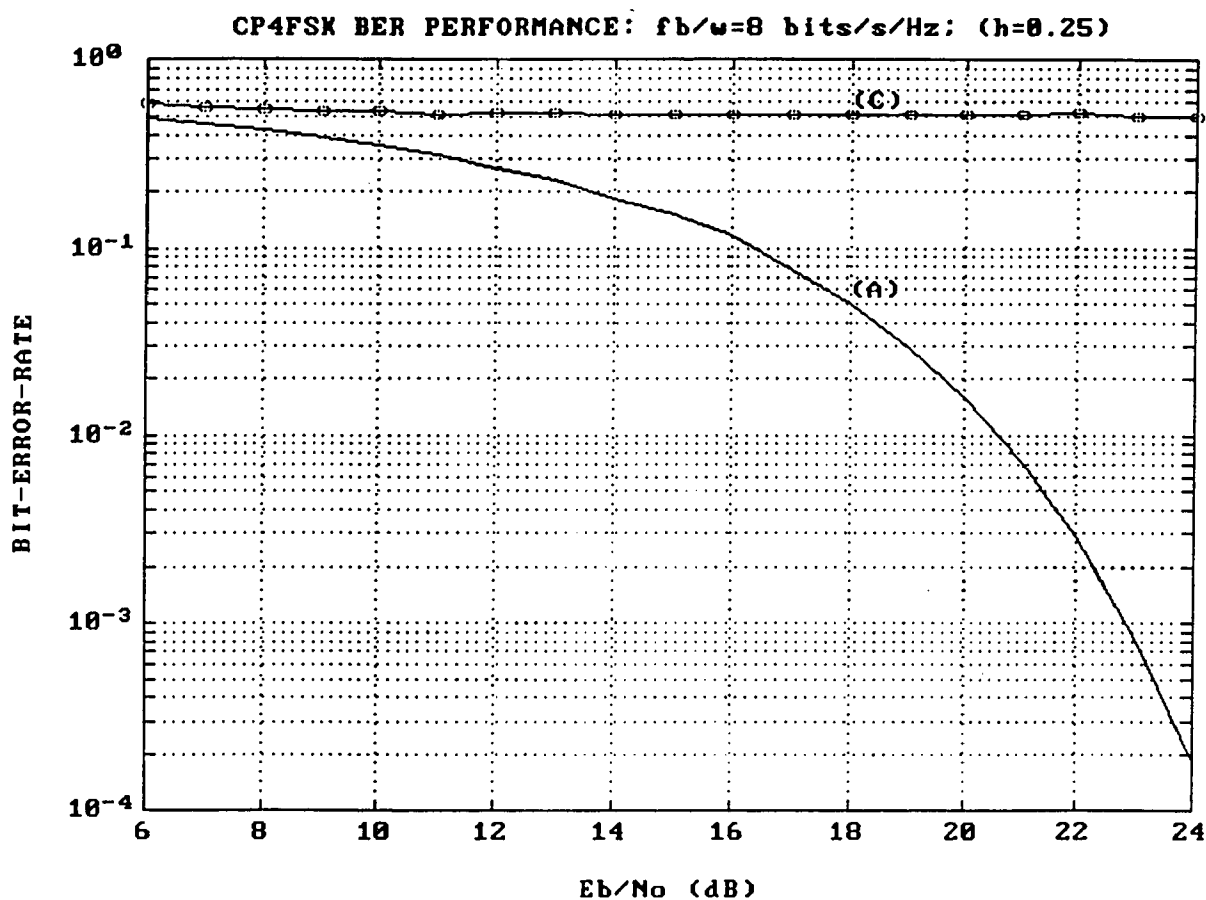


FIGURE 4.12.1: COMPARISON OF CP8FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

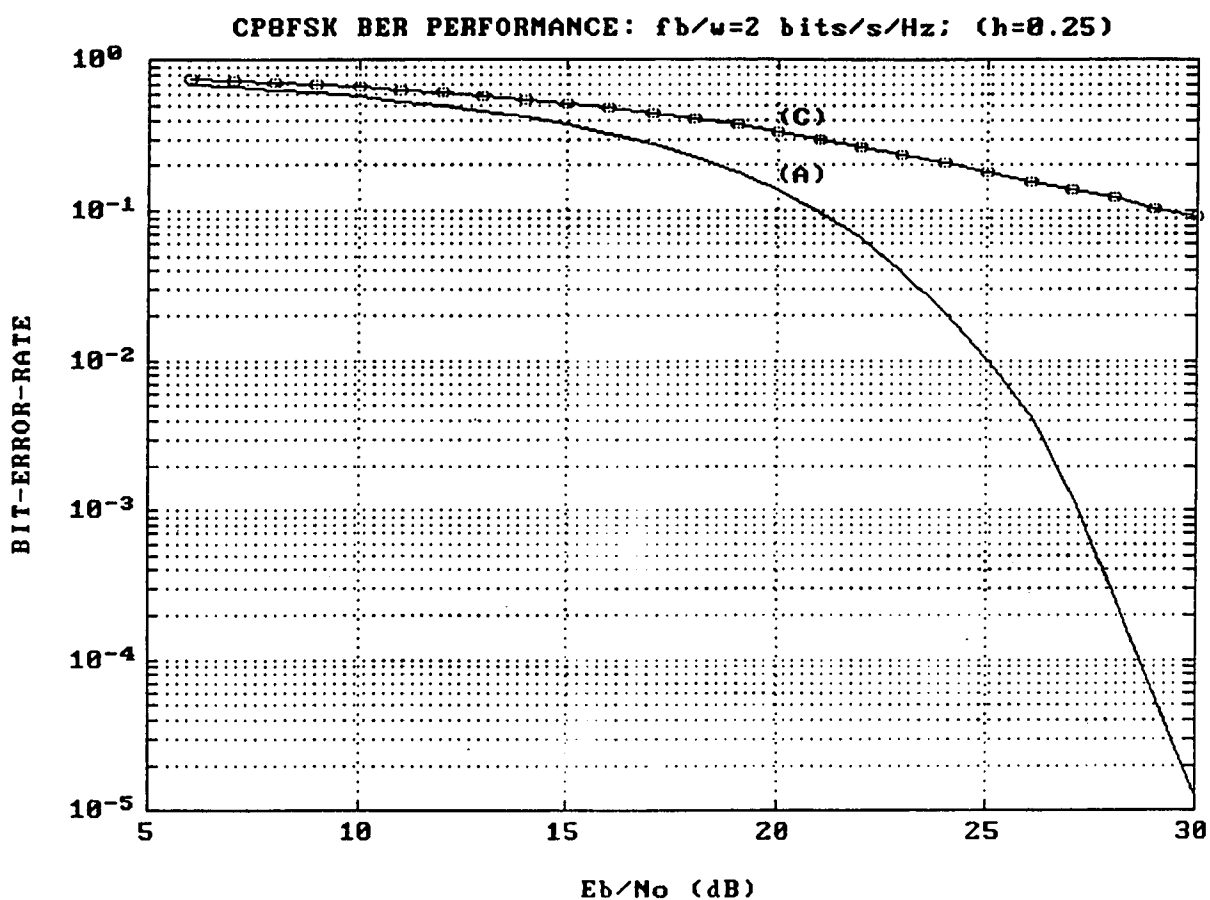


FIGURE 4.12.2: COMPARISON OF CP8FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

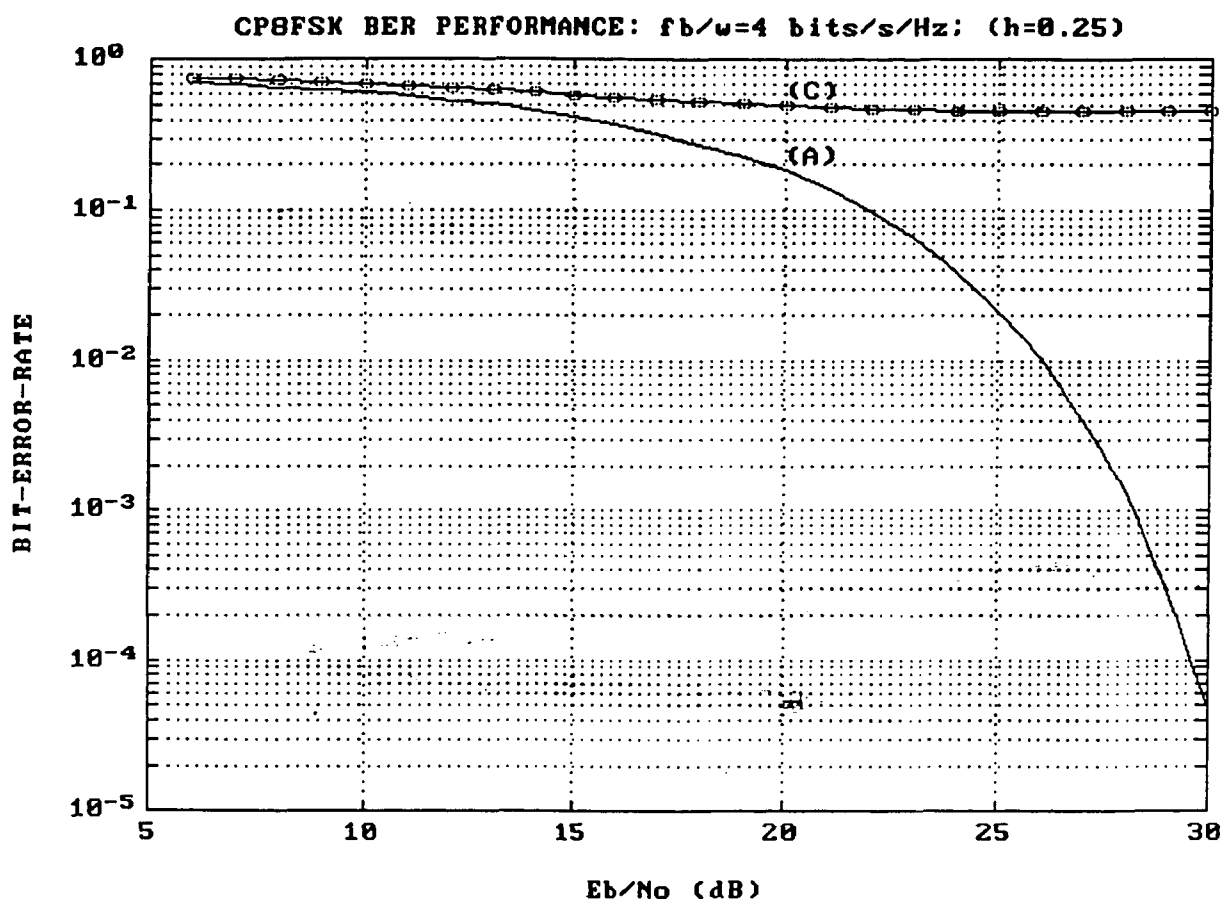


FIGURE 4.12.3: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

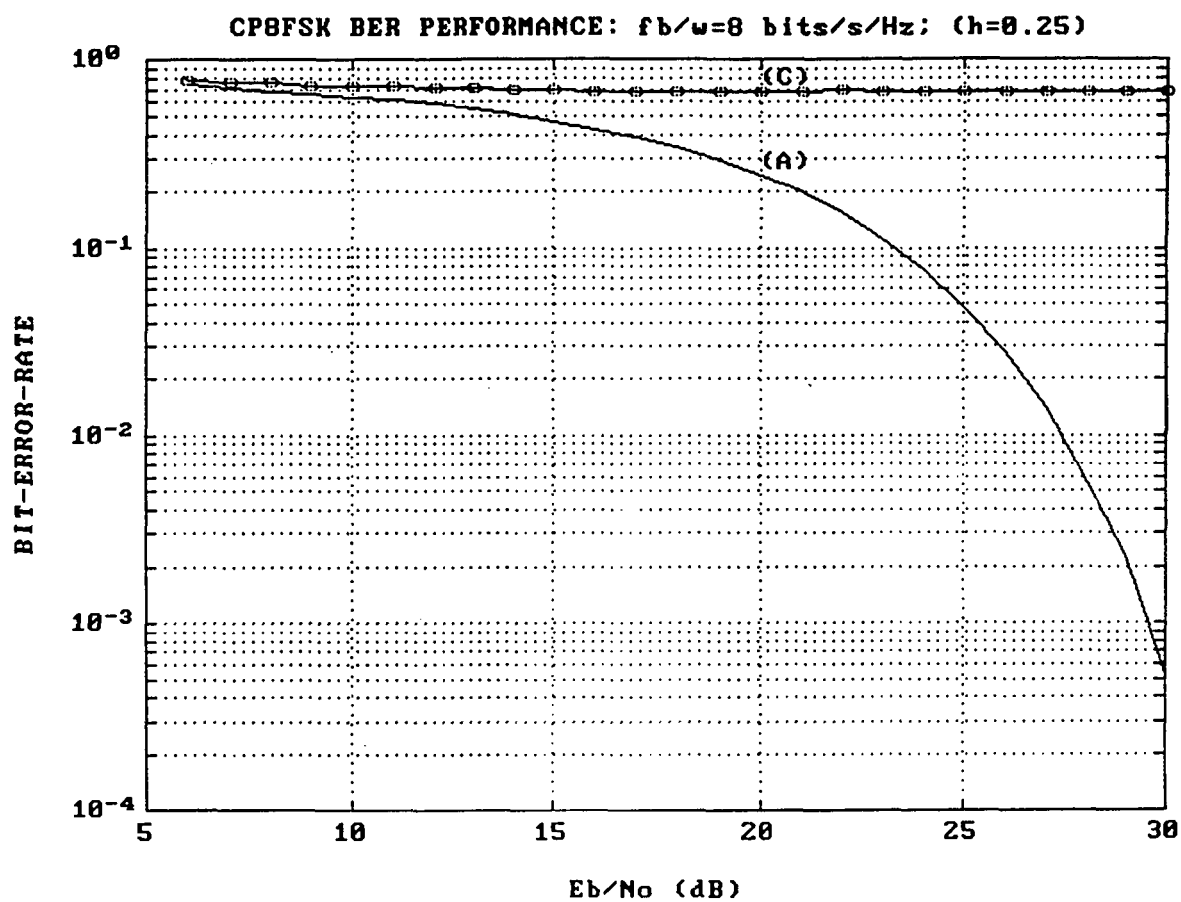


FIGURE 4.13.1: COMPARISON OF **CP16FSK** BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

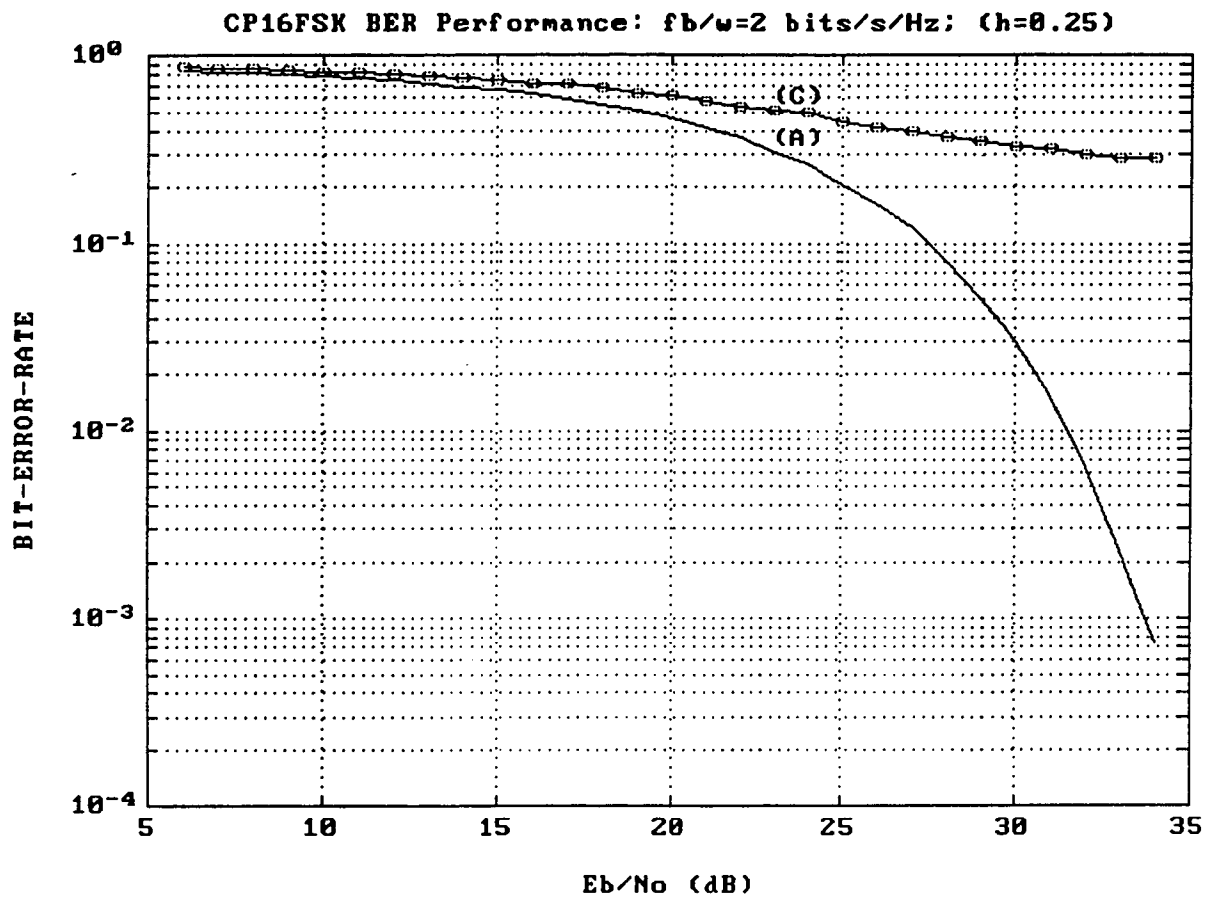


FIGURE 4.13.2: COMPARISON OF CP16FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

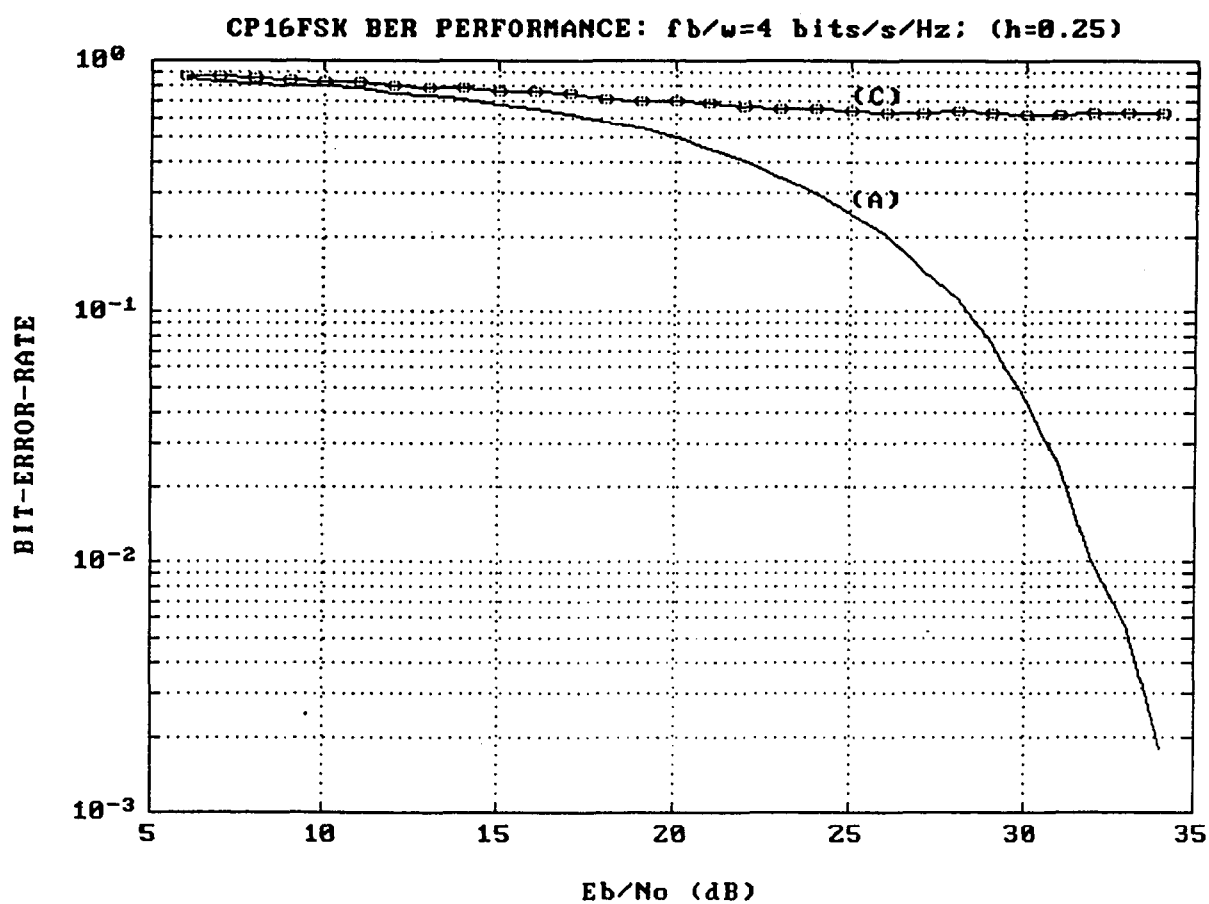


FIGURE 4.13.3: COMPARISON OF CP16FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

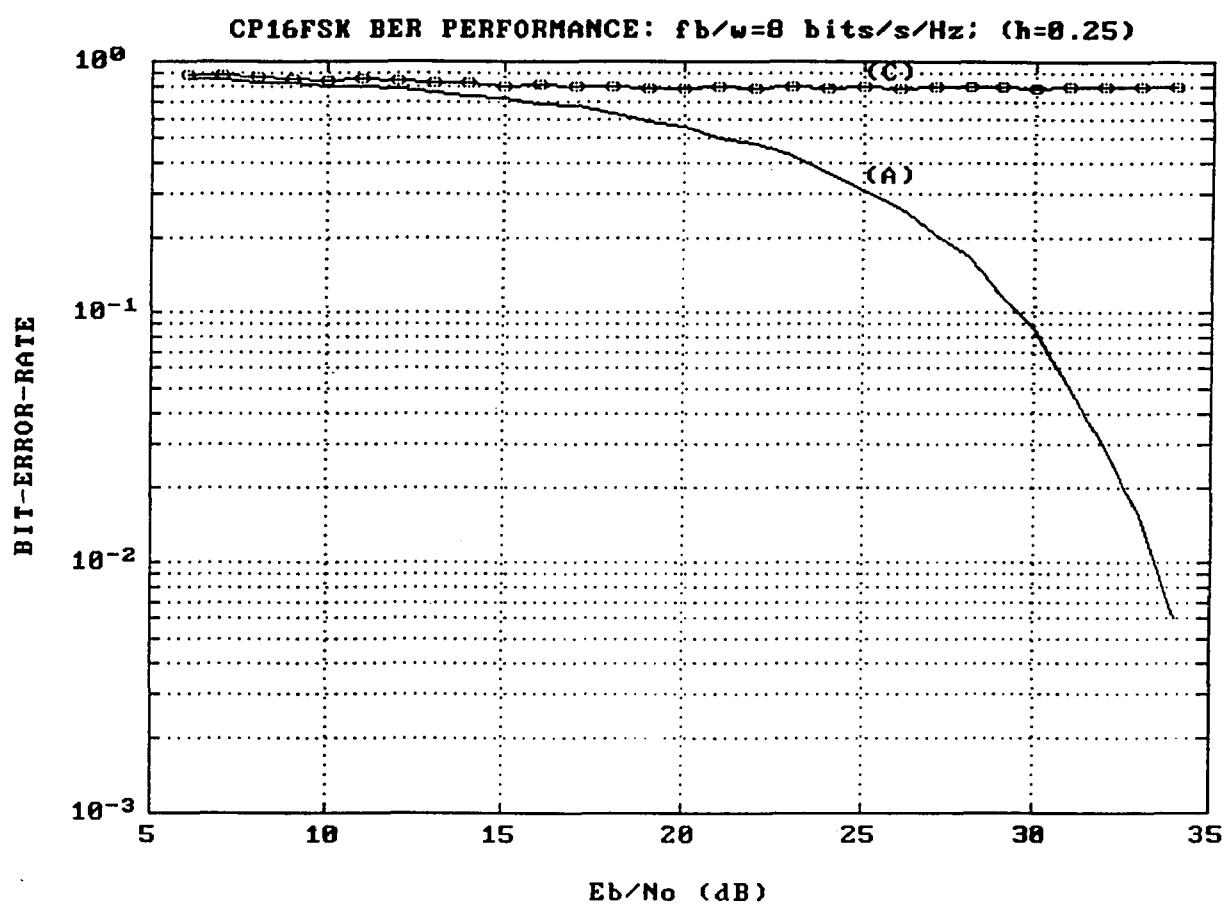


FIGURE 4.14.1: COMPARISON OF CPMFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

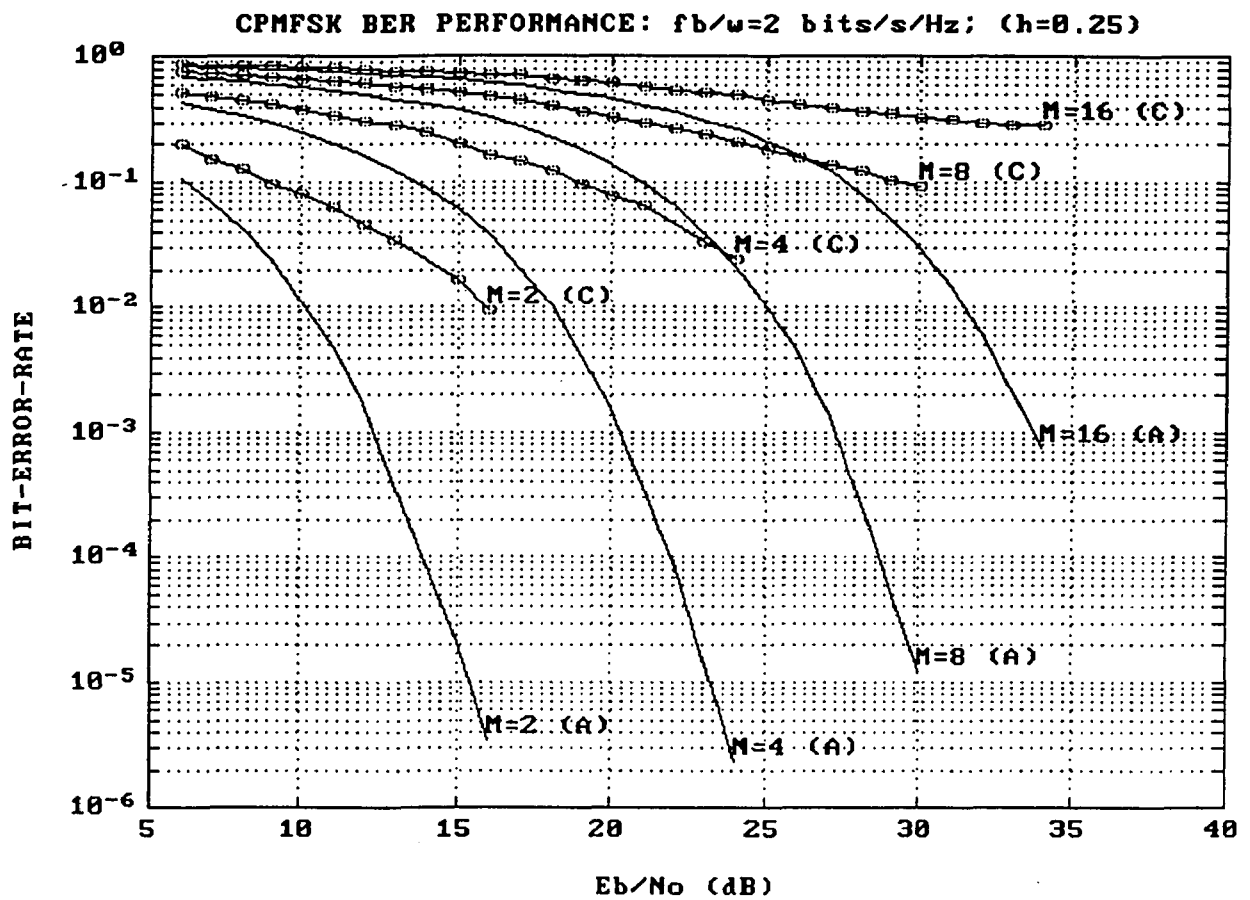


FIGURE 4.14.2: COMPARISON OF CPMFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

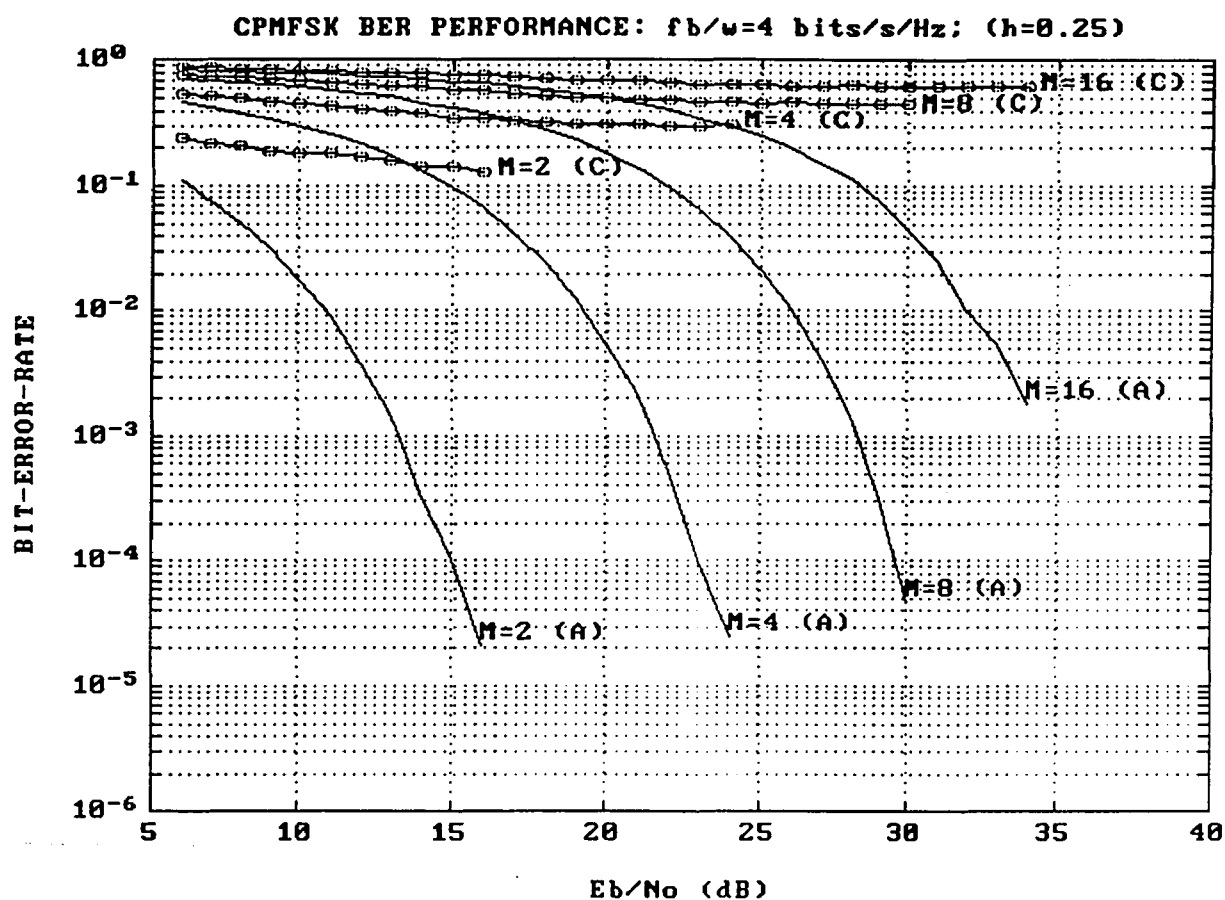


FIGURE 4.14.3: COMPARISON OF CPMFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A ONE-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

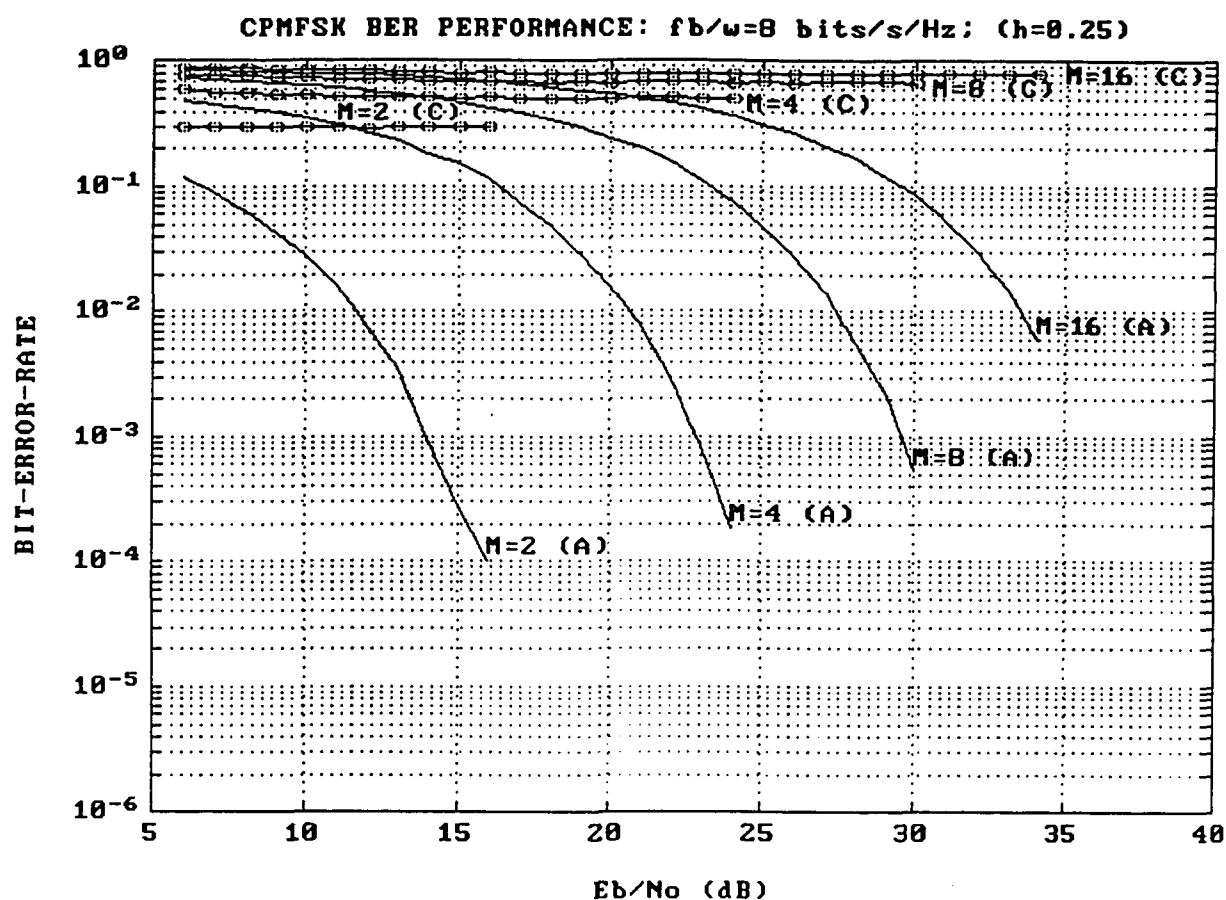


FIGURE 4.15.1: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

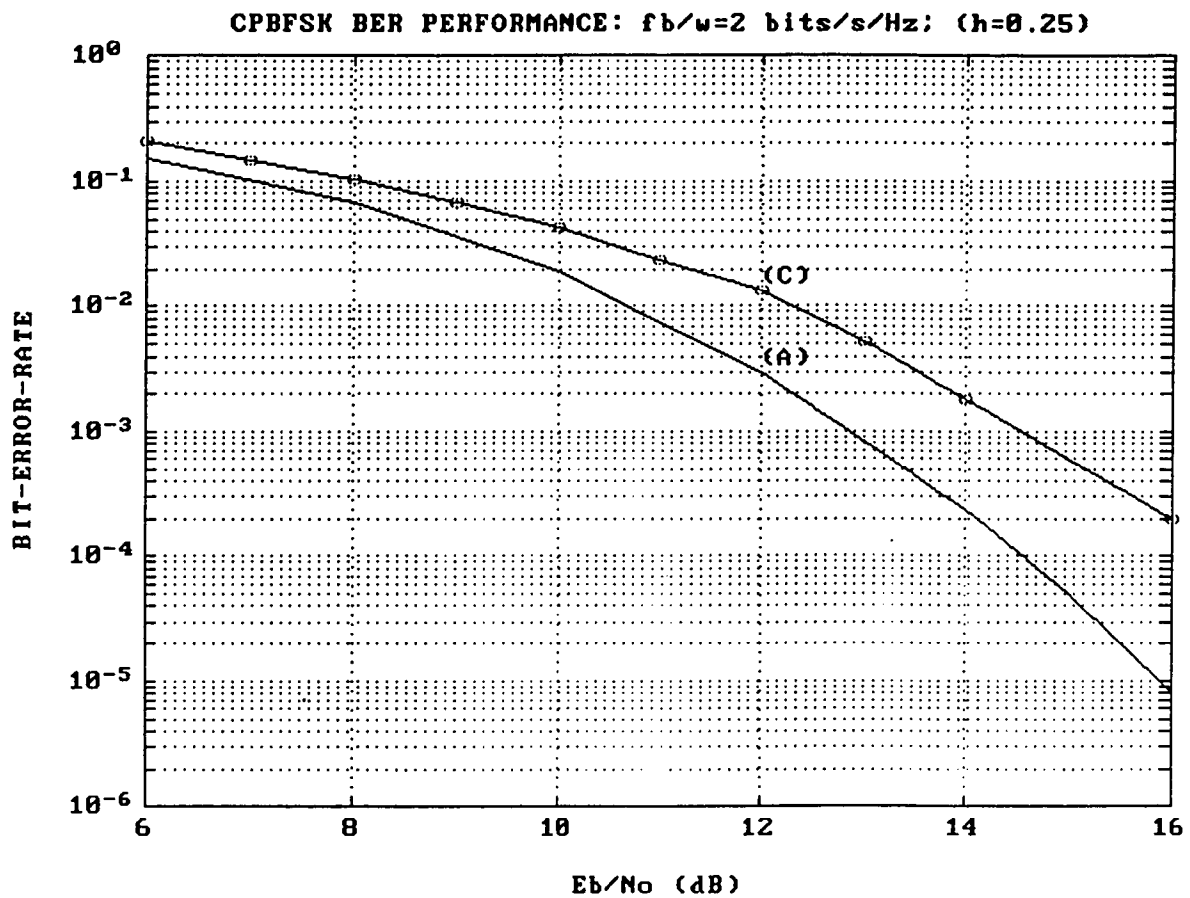


FIGURE 4.15.2: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

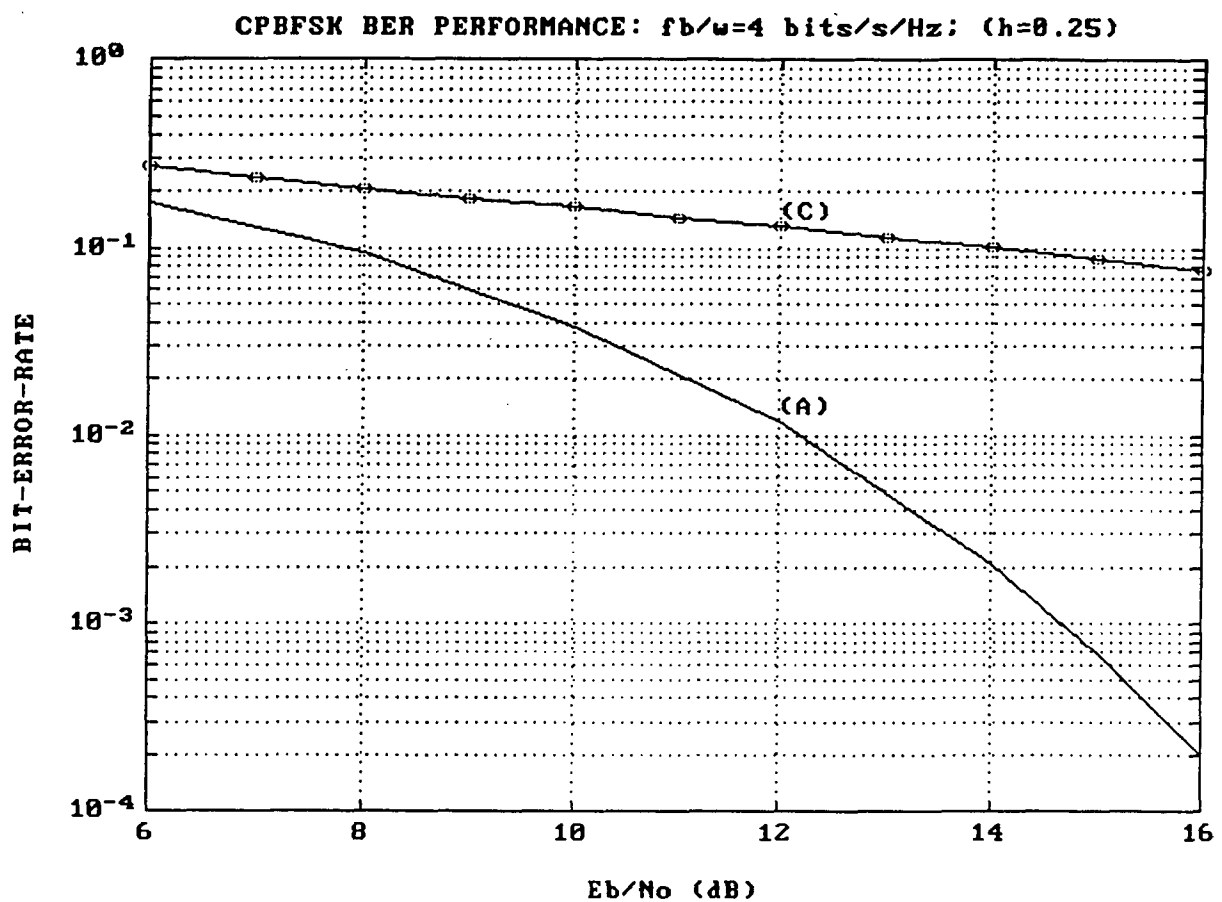


FIGURE 4.15.3: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

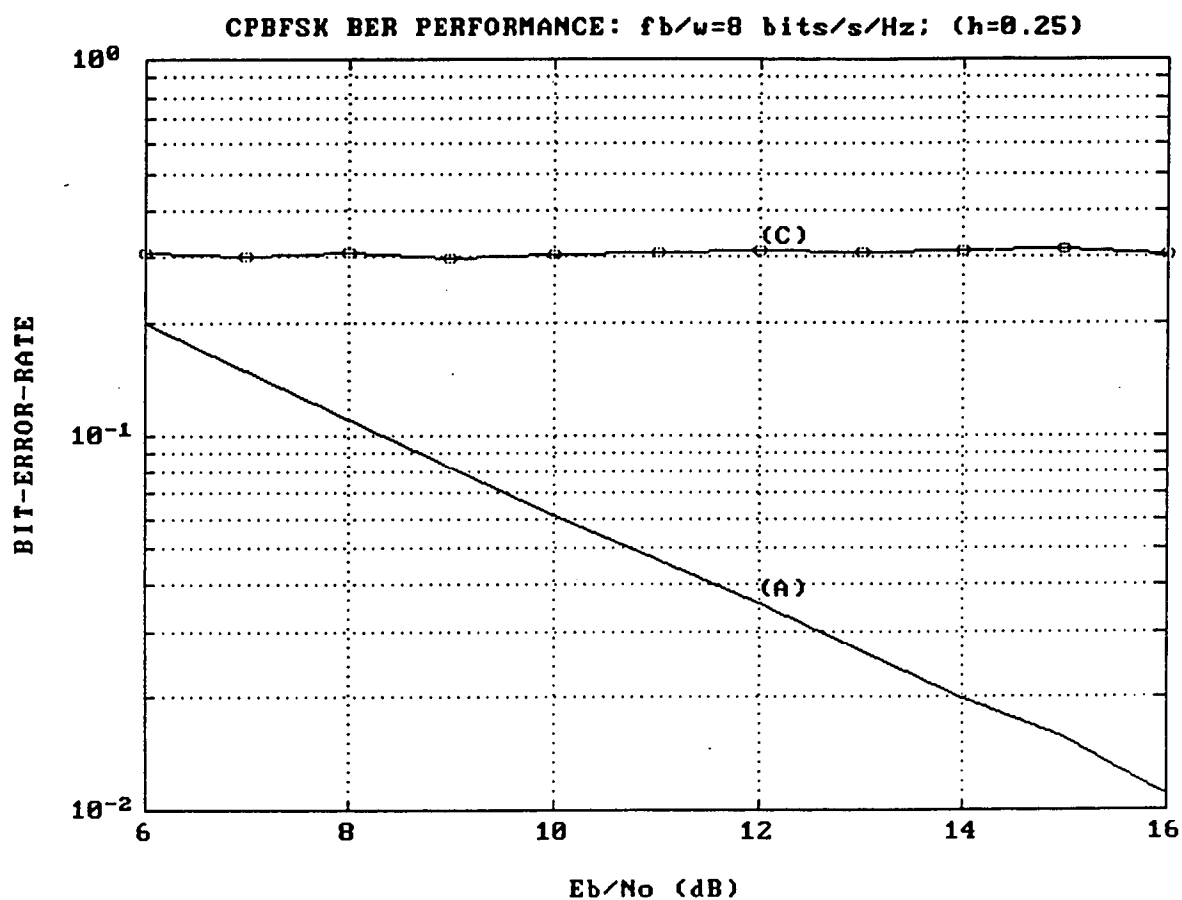


FIGURE 4.16.1: COMPARISON OF **CP4FSK** BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/\omega = 2 \text{ bits/s/Hz}$

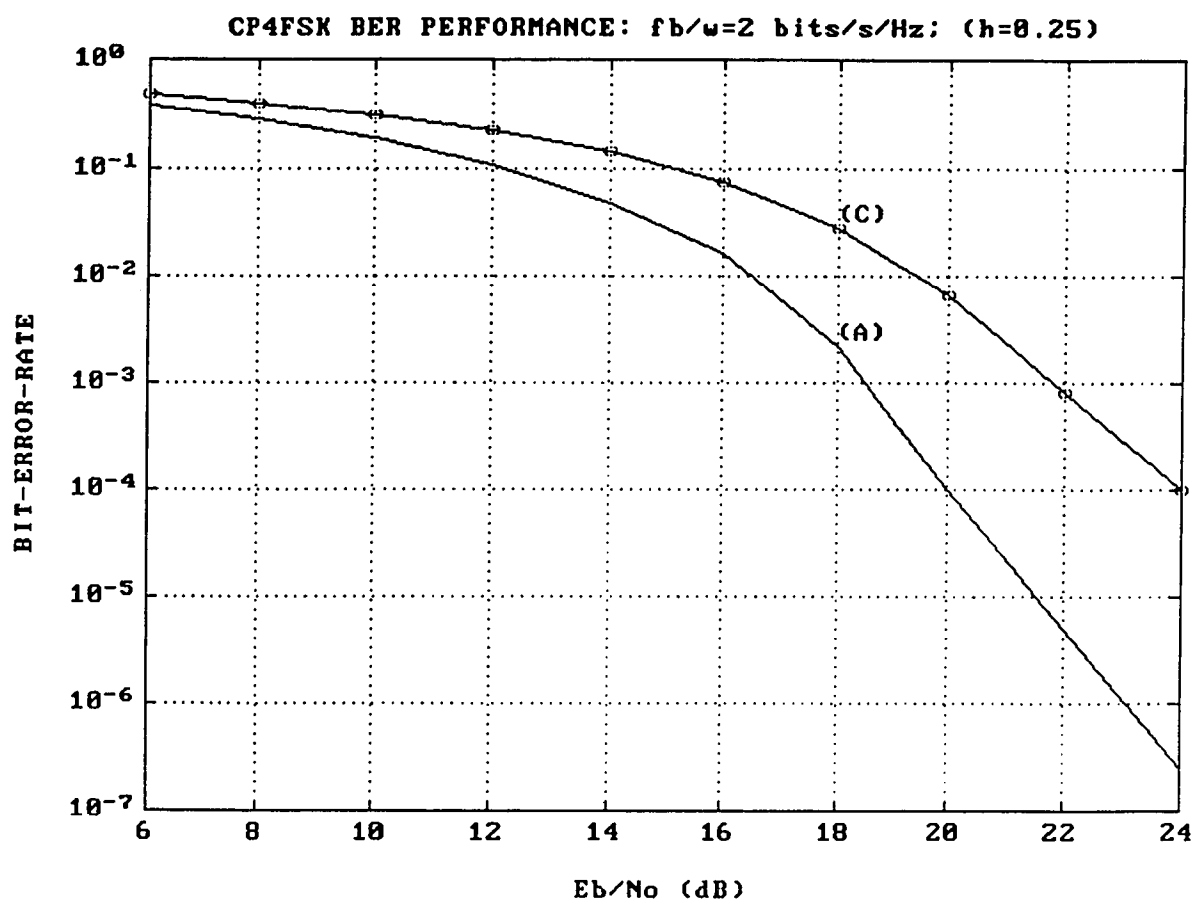


FIGURE 4.16.2: COMPARISON OF CP4FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

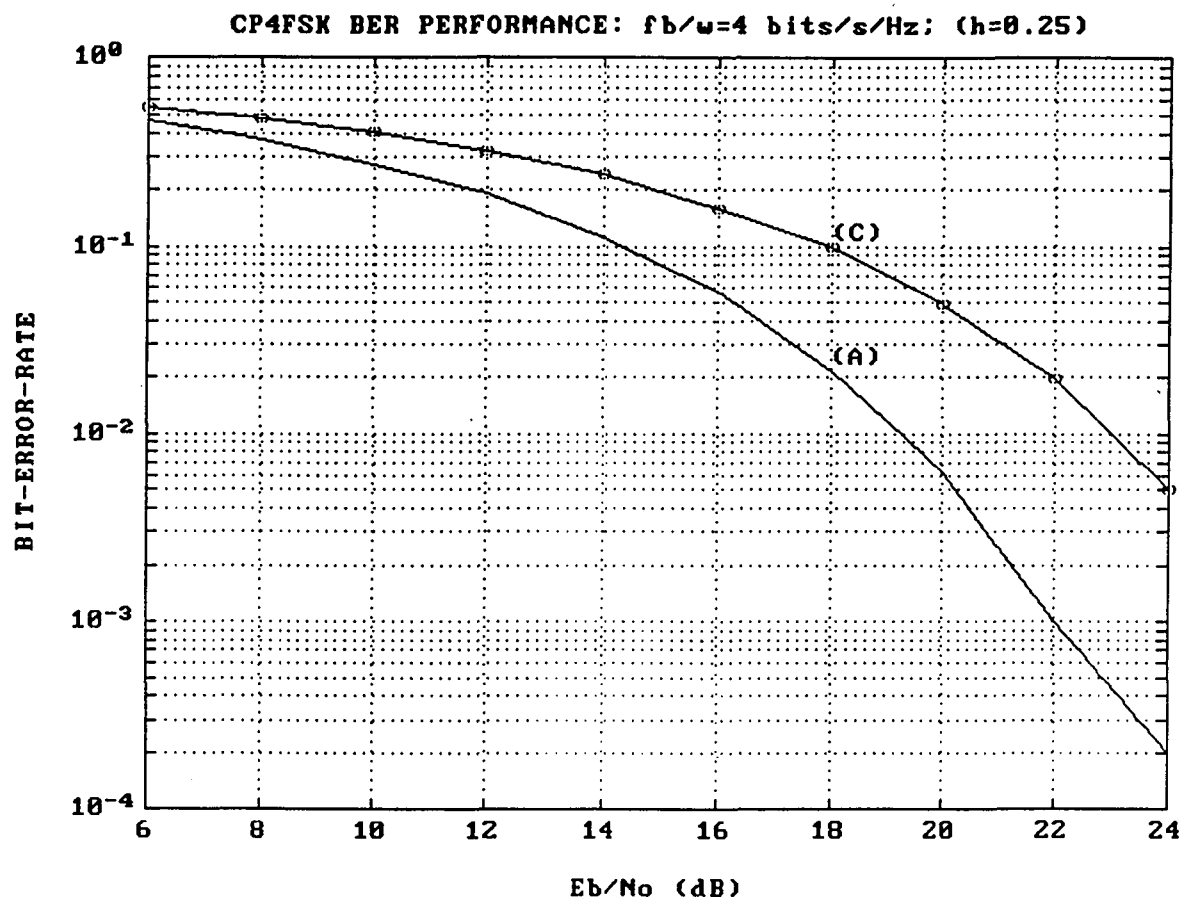


FIGURE 4.16.3: COMPARISON OF CP4FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

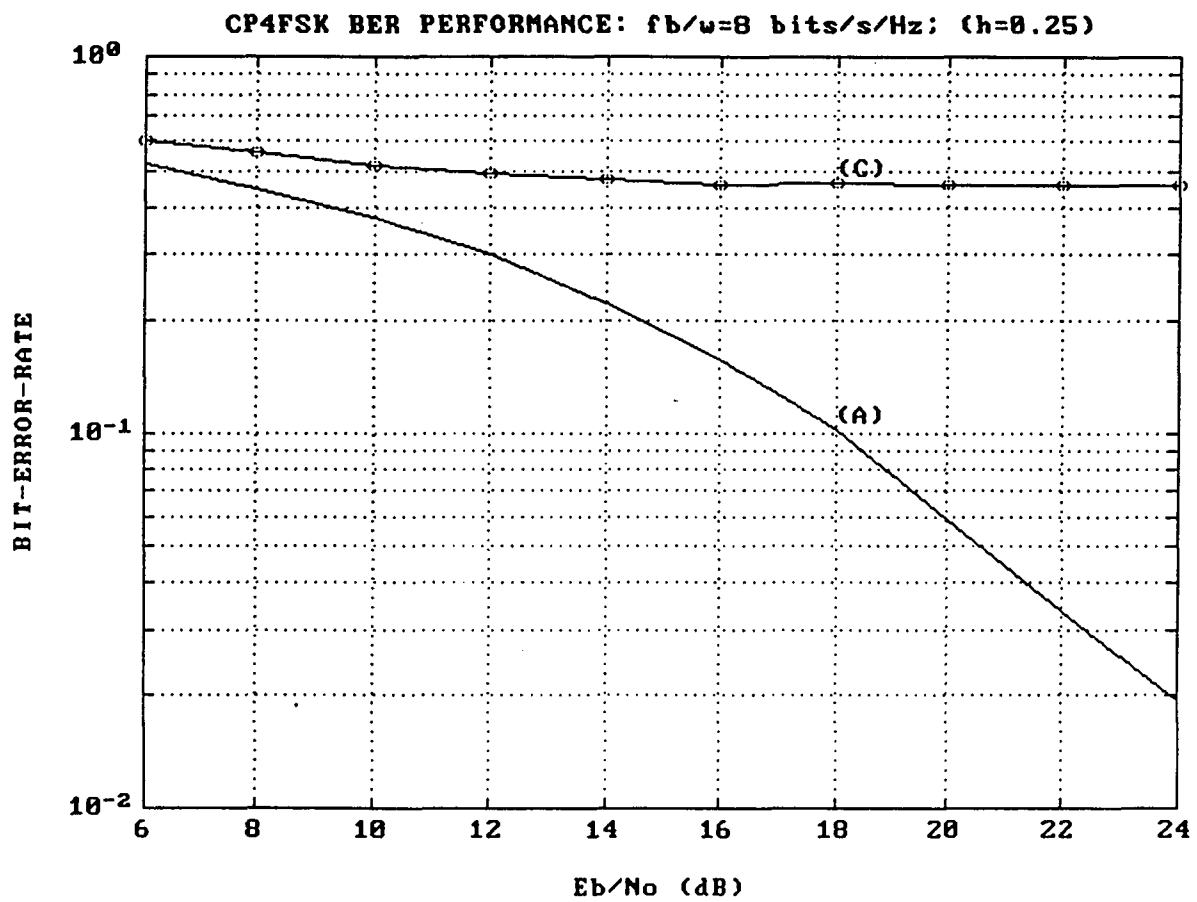


FIGURE 4.17.1: COMPARISON OF CP8FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

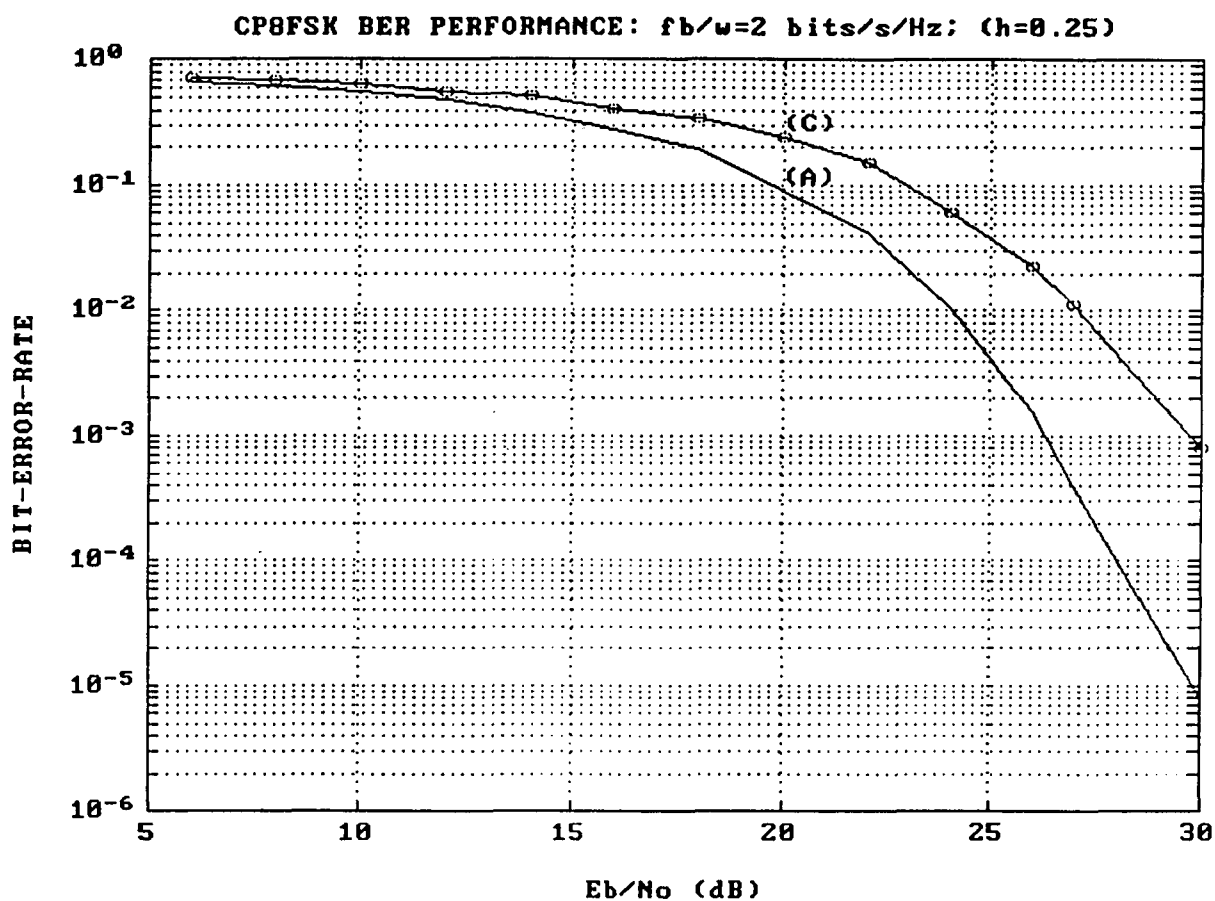


FIGURE 4.17.2: COMPARISON OF CP8FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

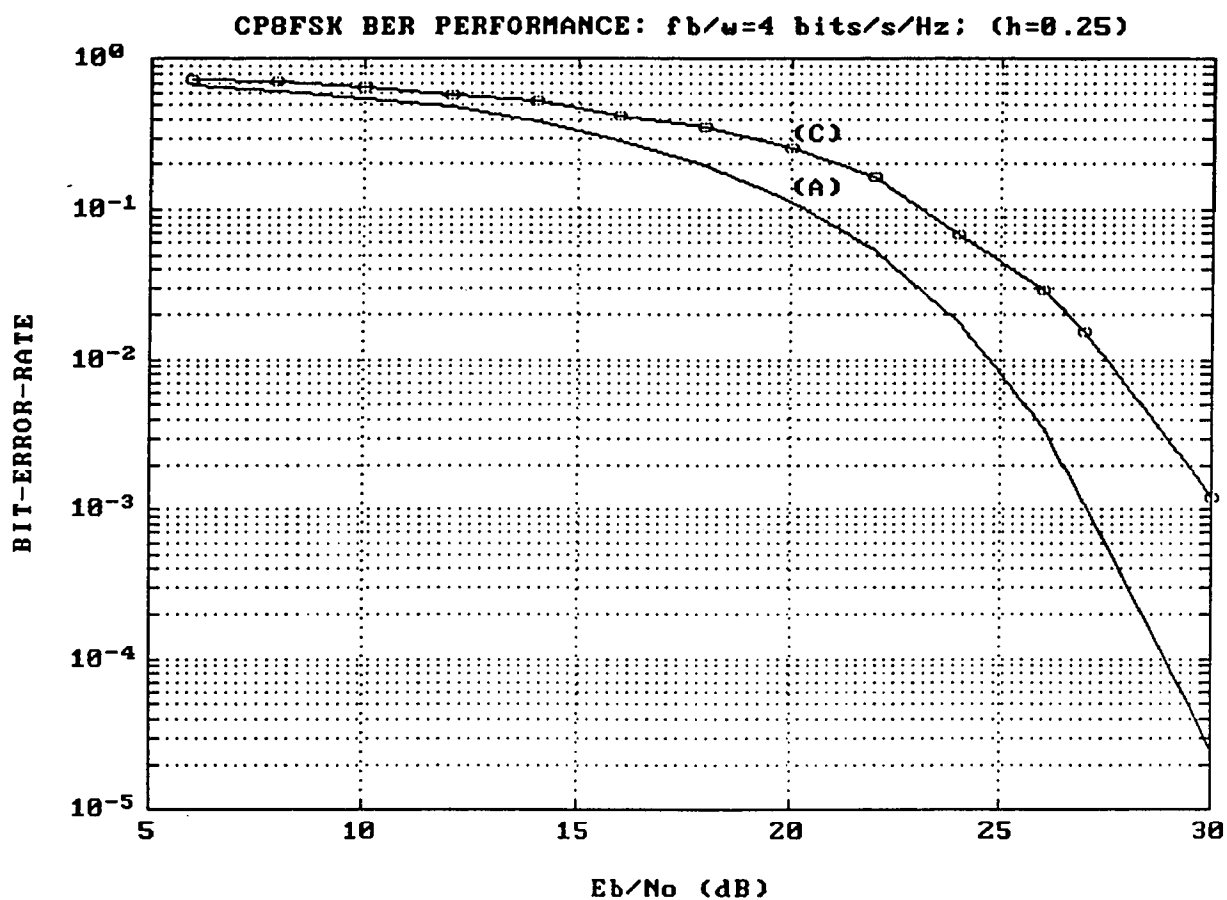


FIGURE 4.17.3: COMPARISON OF CP8FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

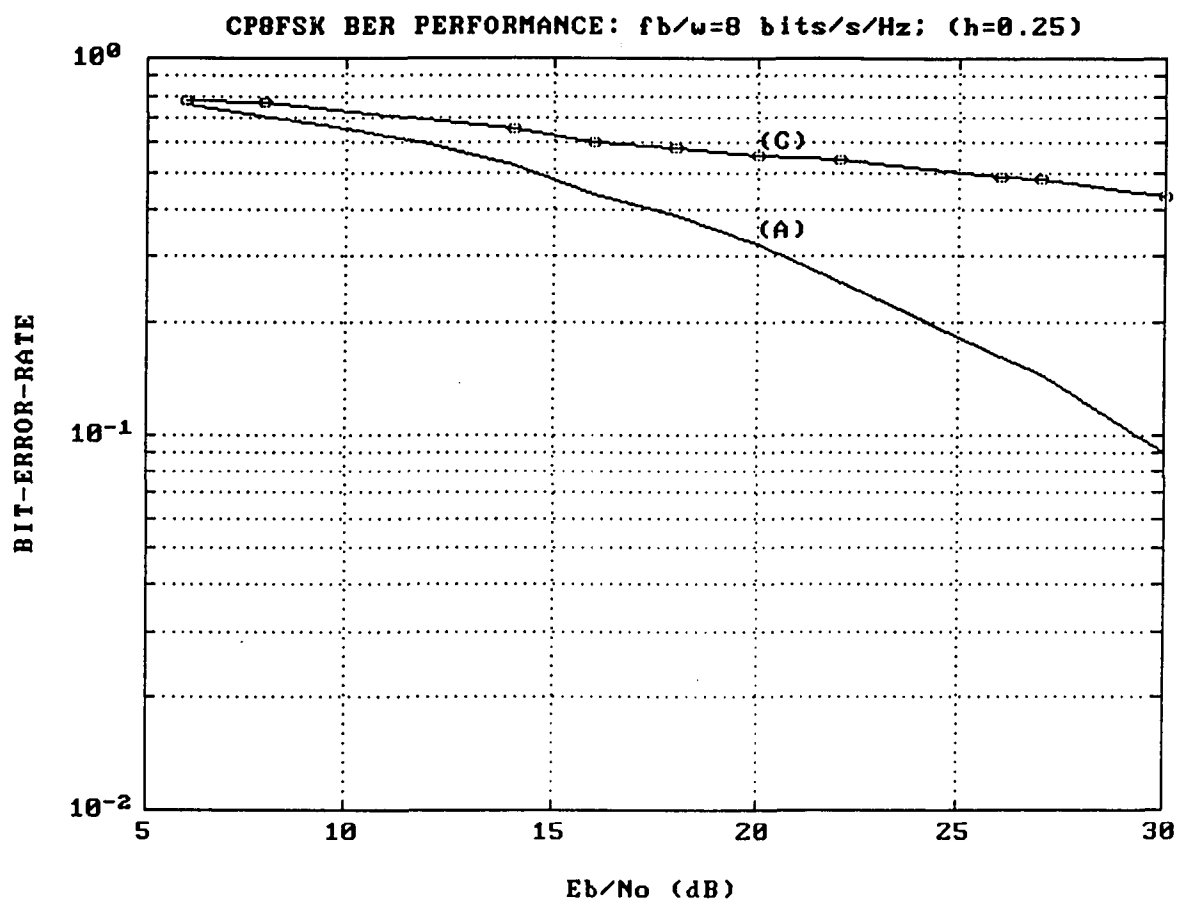


FIGURE 4.18.1: COMPARISON OF CP16FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

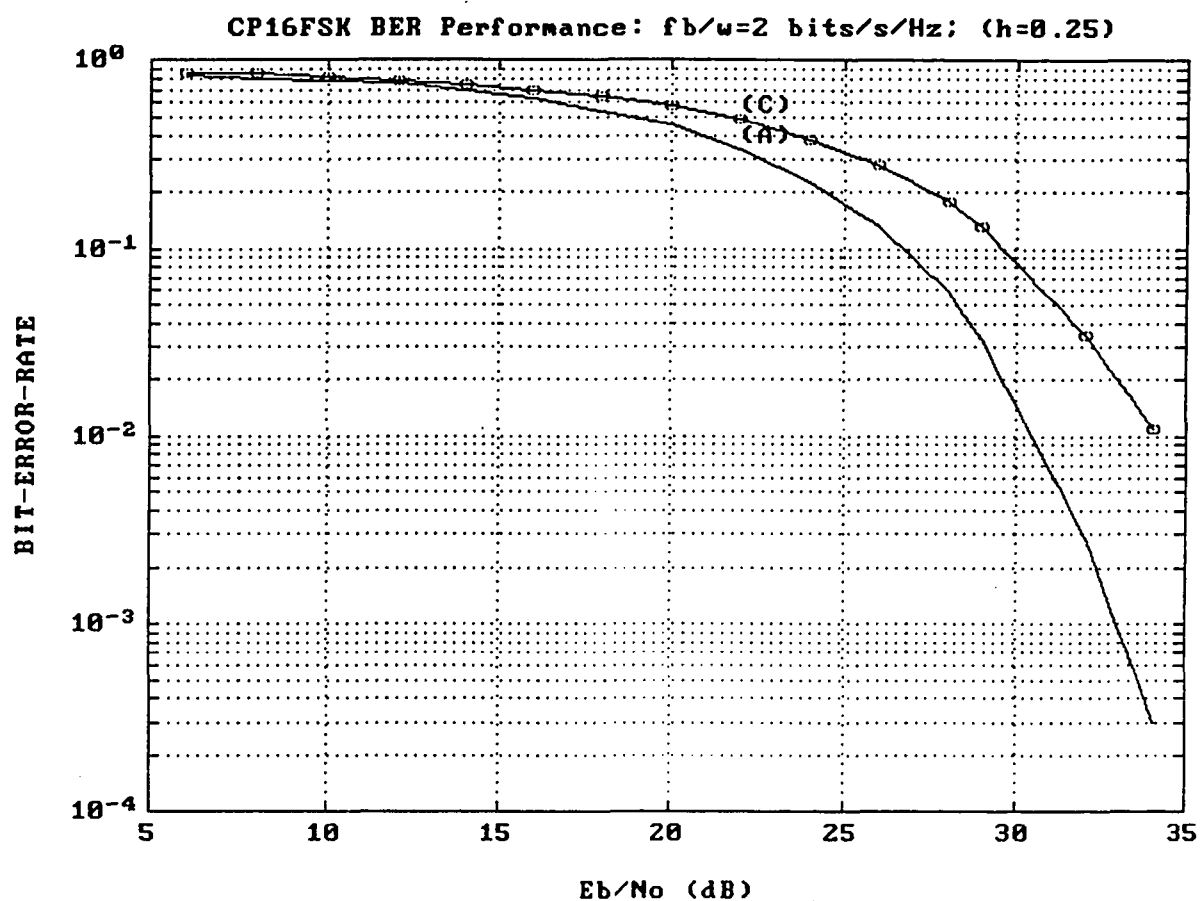


FIGURE 4.18.2: COMPARISON OF CP16FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 4 \text{ bits/s/Hz}$

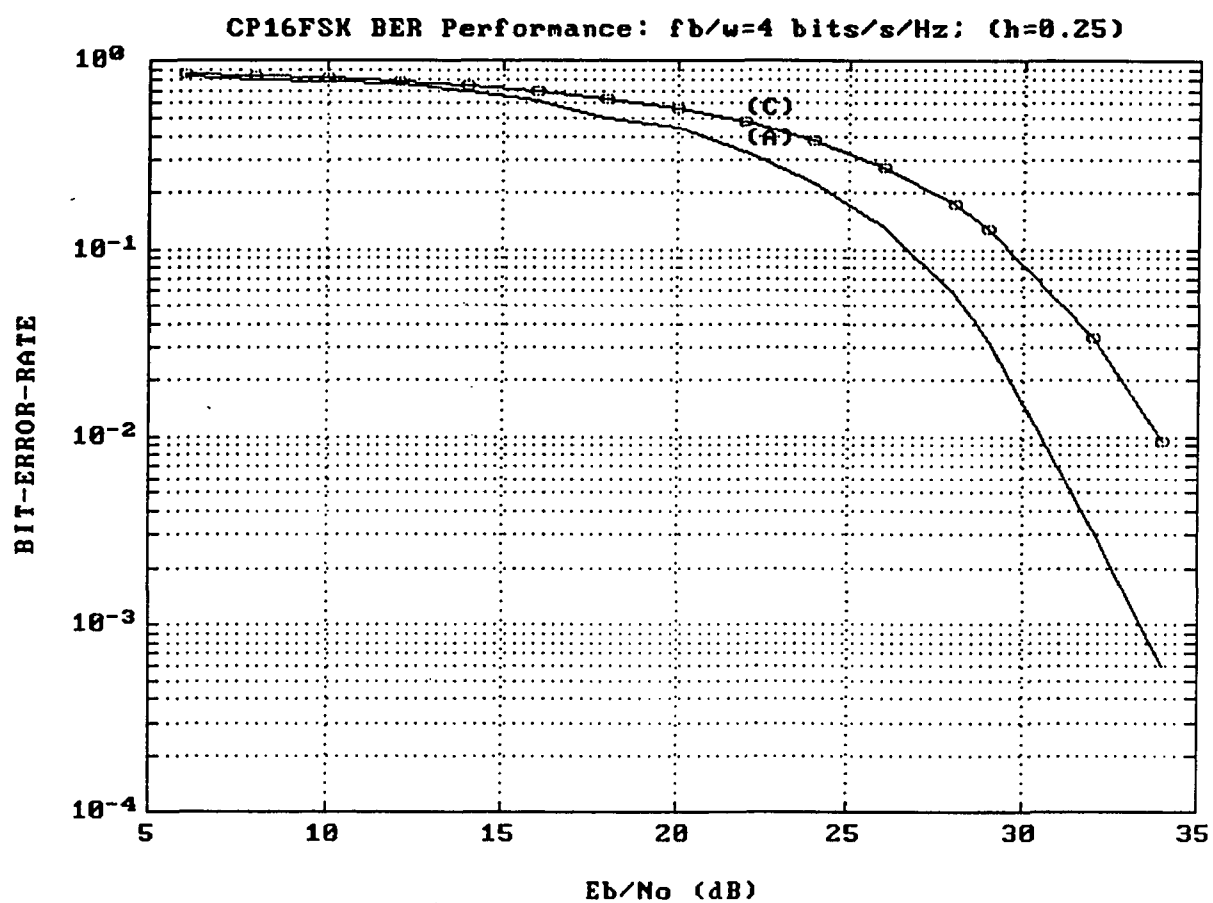


FIGURE 4.18.3: COMPARISON OF CP16FSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$

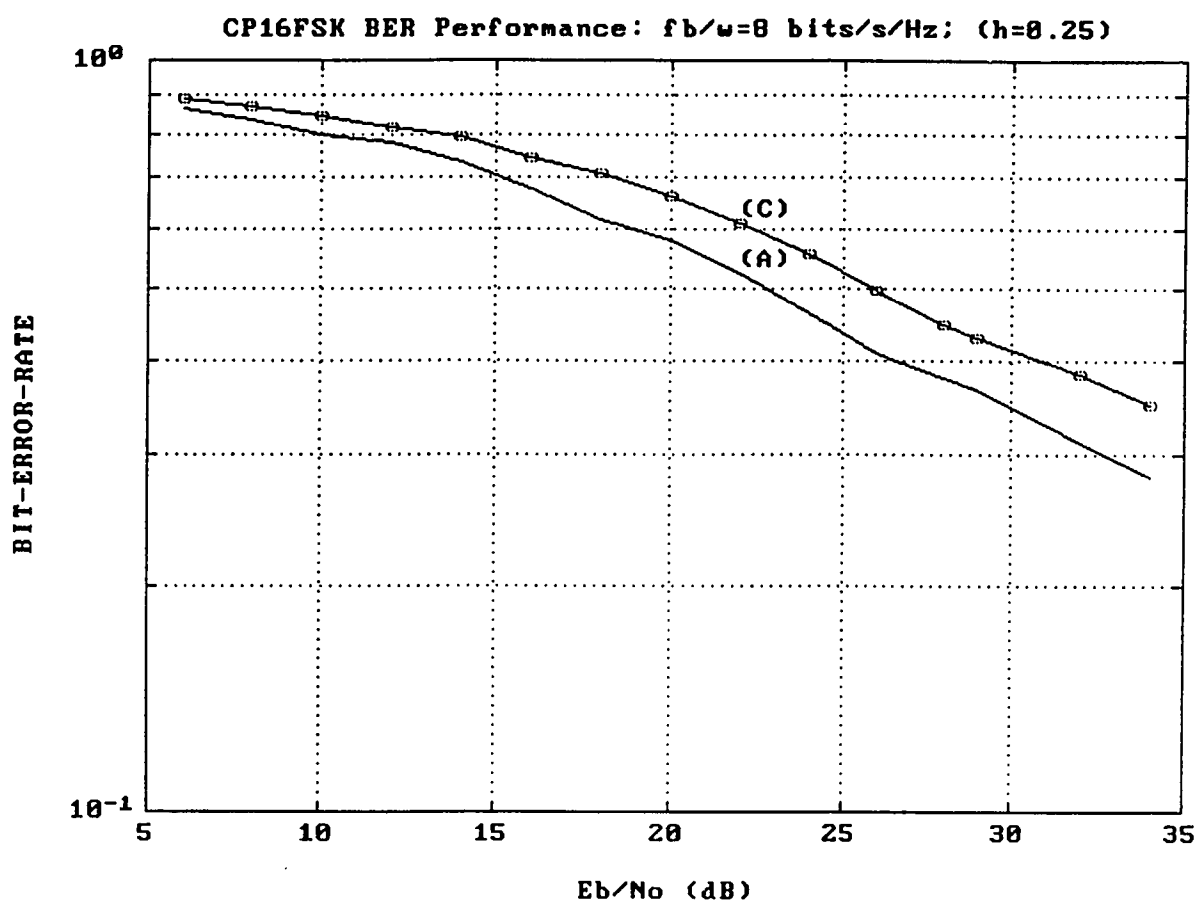


FIGURE 4.19.1: COMPARISON OF CPMFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 2 \text{ bits/s/Hz}$

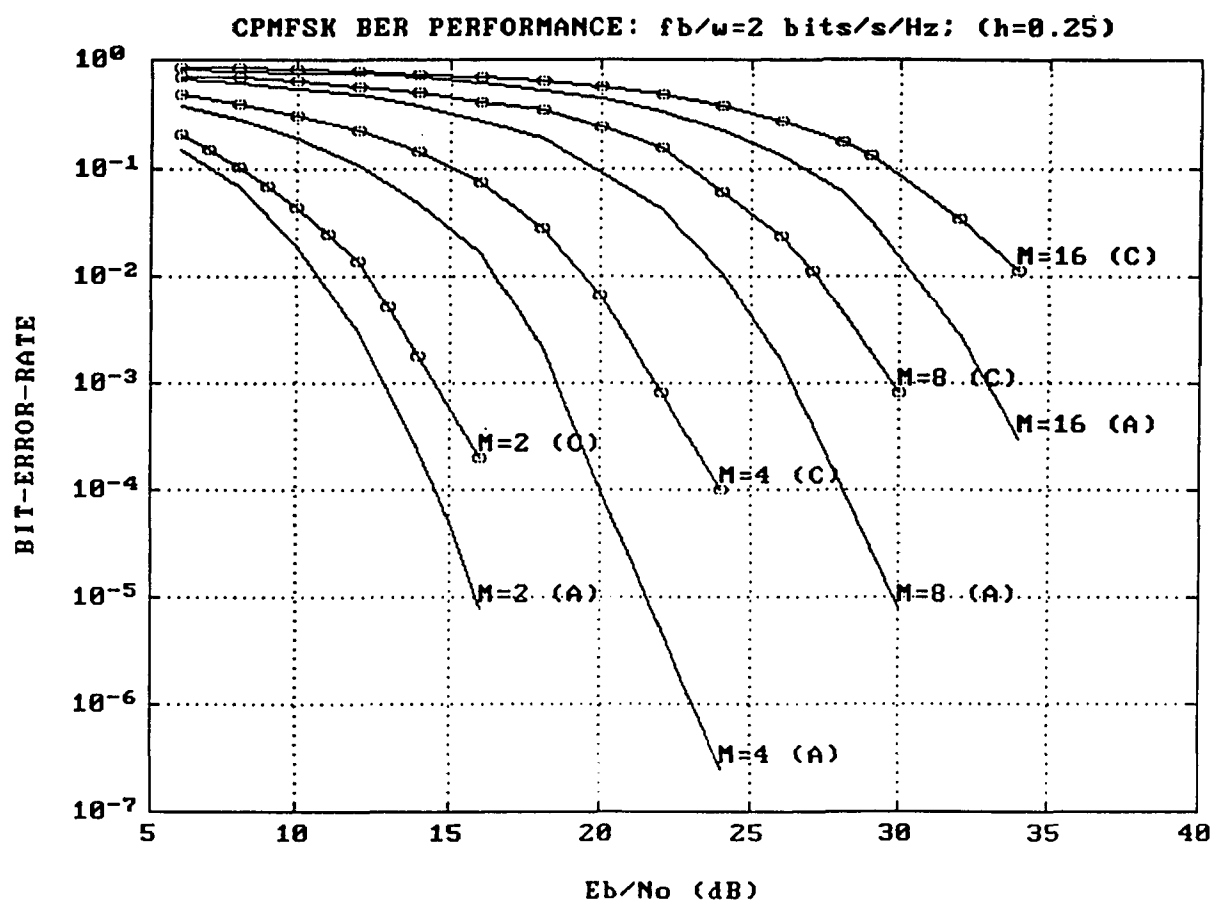


FIGURE 4.19.2: COMPARISON OF **CPMFSK** BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/\omega = 4 \text{ bits/s/Hz}$

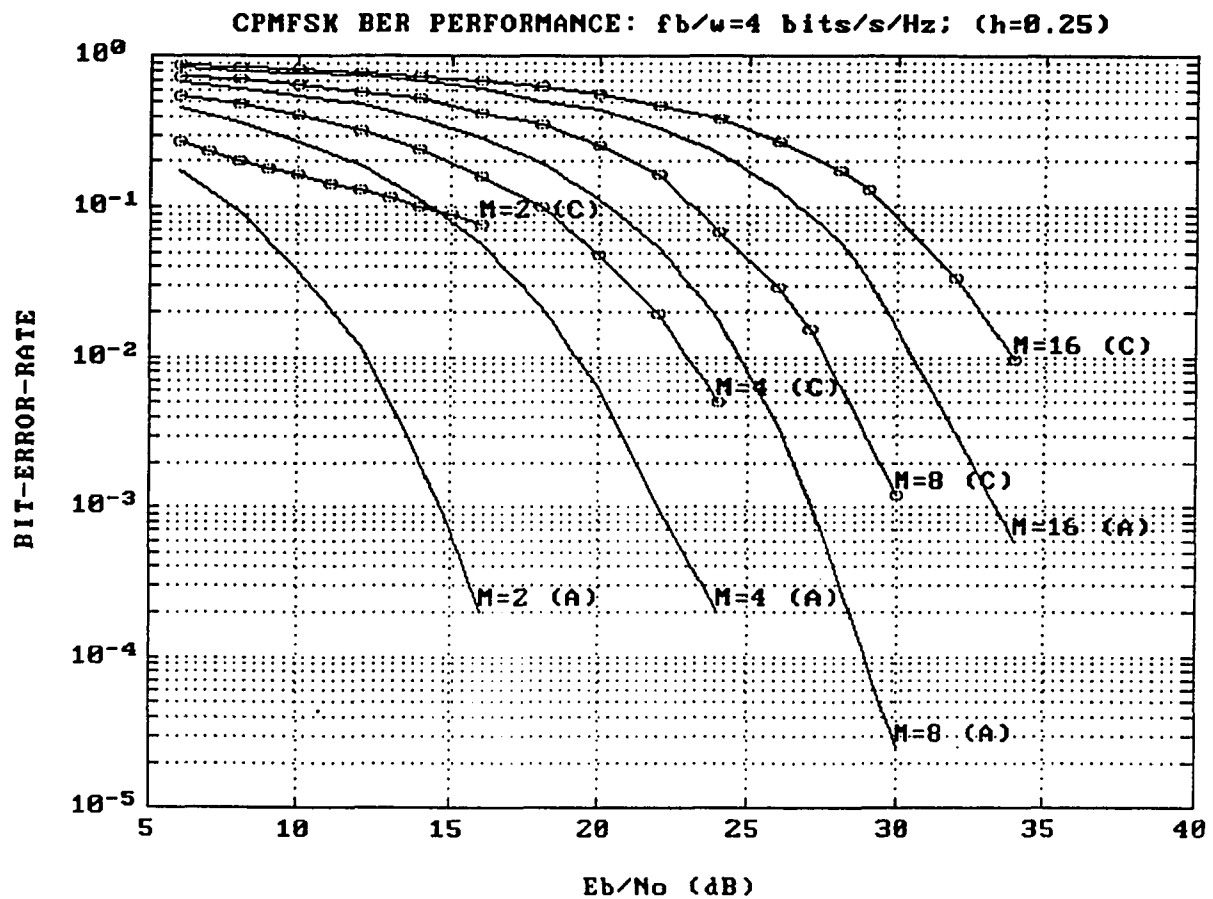
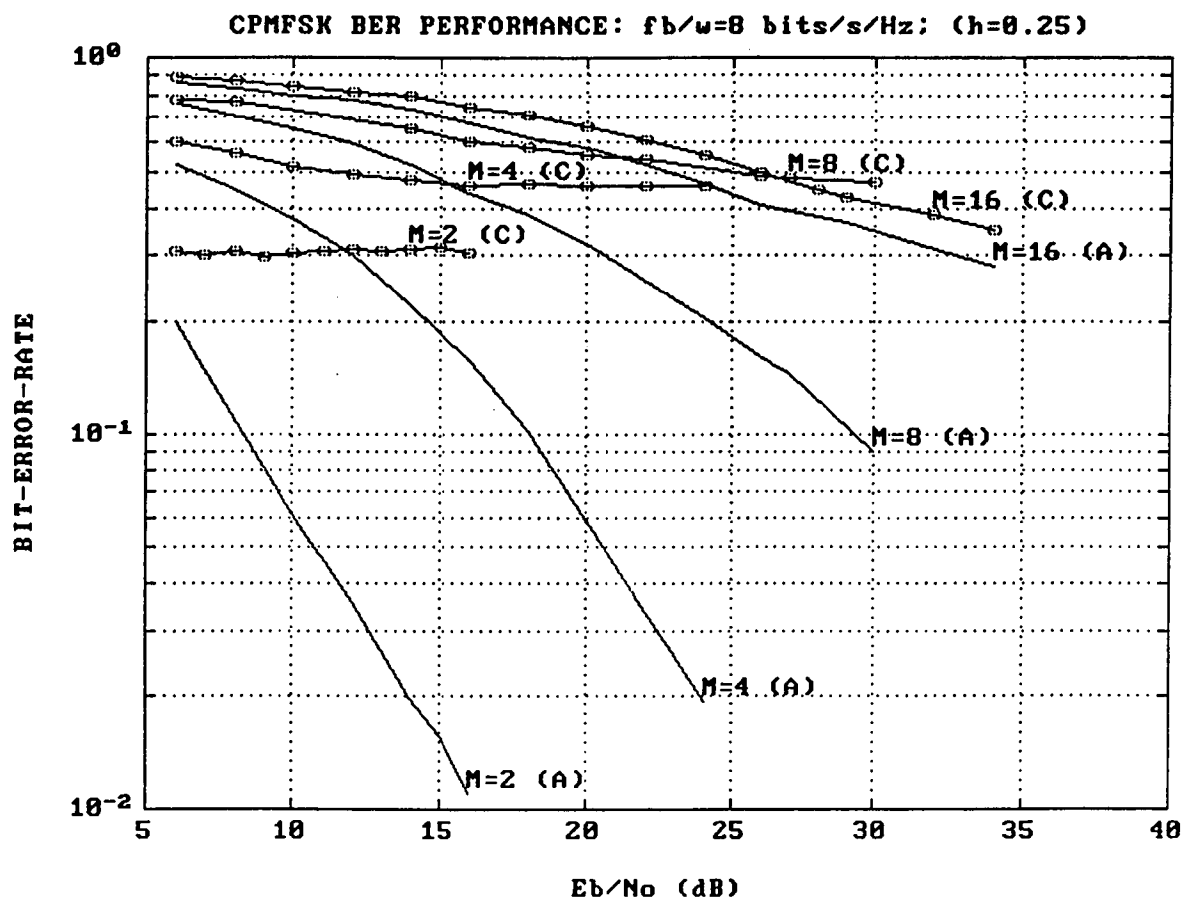


FIGURE 4.19.3: COMPARISON OF **CPMFSK** BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS USING A TWO-POLE BUTTERWORTH FILTER WITH $f_b/w = 8 \text{ bits/s/Hz}$



5 FUTURE WORK

The detection technique proposed in this research is for a digital signal corrupted by intersymbol interference (ISI) and additive white Gaussian noise (AWGN). This system is spectrally efficient unlike conventional narrowband FN system; Therefore, it can be used to build high data rate modems. Also, there is no need to lock into phase and frequency since the non-coherent detection is used. This property makes the described demodulation ideal for the fast fading environments such as mobile radio communication channels. This finding justifies and points toward addressing of the other main impairment to reliable communications, namely multipath fading which results in a very low received SNR when channel exhibit a deep fade.

FIGURE 4.20.1: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS WITH $f_b/w = 2 \text{ bits/s/Hz}$

(P2): FILTER TYPE IS KNOWN BY ADAPTIVE DETECTOR

(P21): FILTER TYPE IS UNKNOWN BY ADAPTIVE DETECTOR

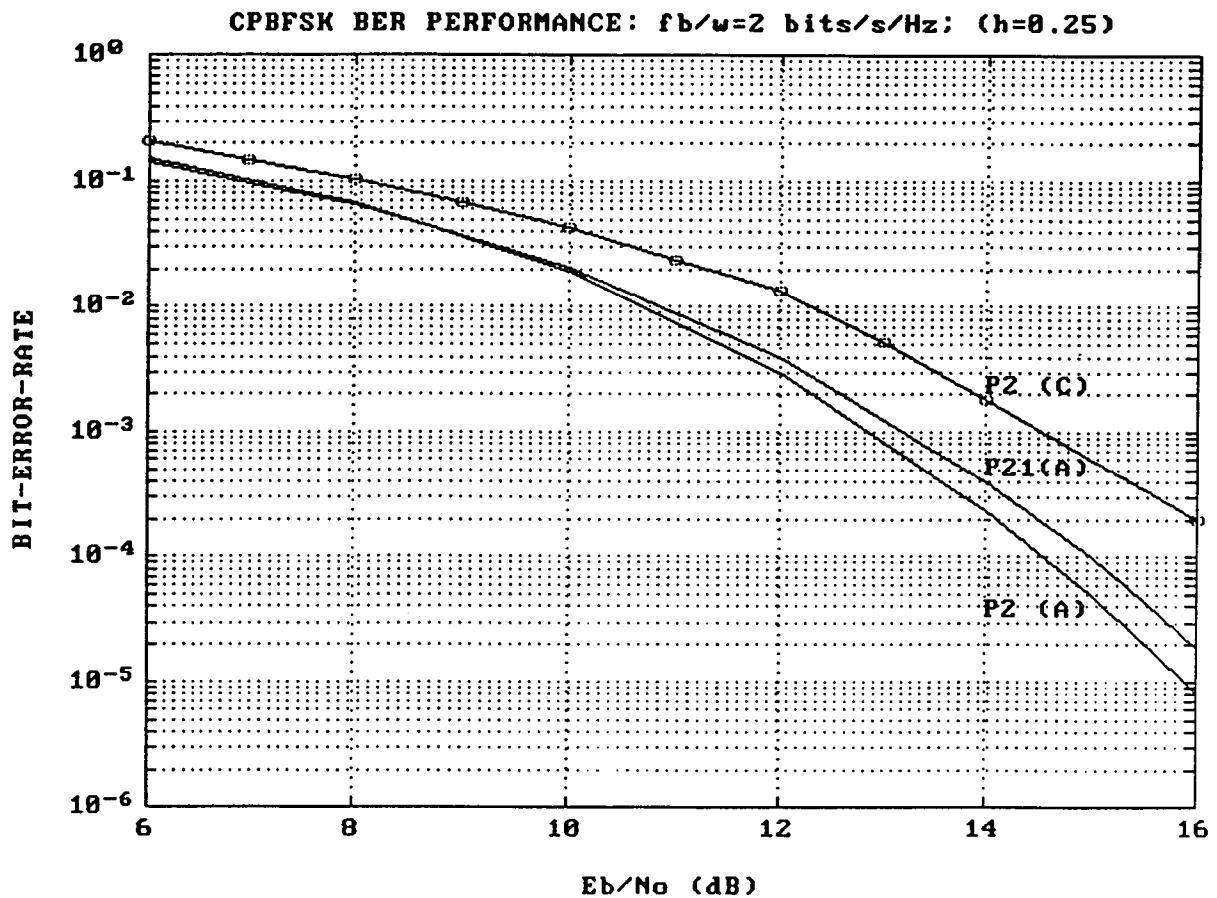


FIGURE 4.20.2: COMPARISON OF **CPBFSK** BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS WITH $f_b/w = 4 \text{ bits/s/Hz}$

(P2): FILTER TYPE IS KNOWN BY ADAPTIVE DETECTOR

(P21): FILTER TYPE IS UNKNOWN BY ADAPTIVE DETECTOR

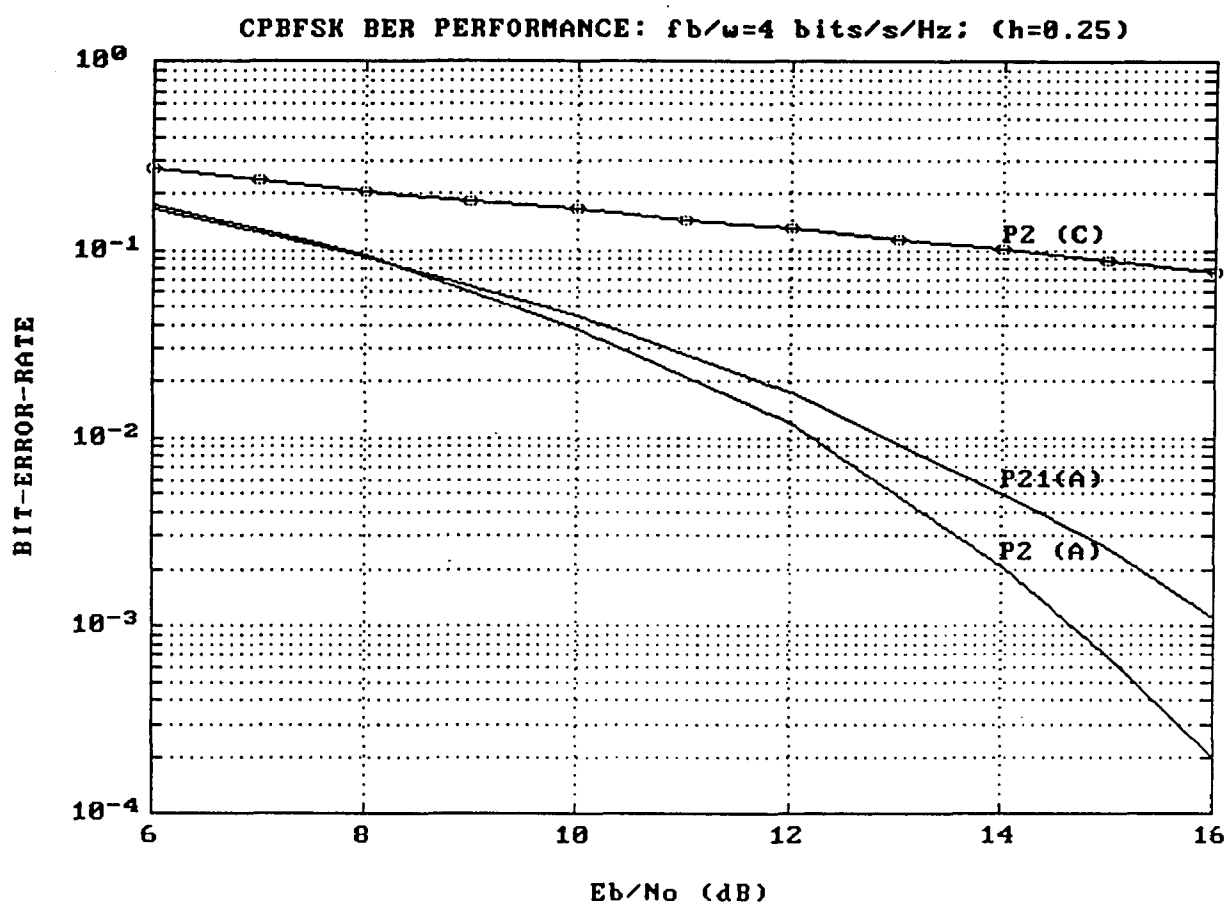
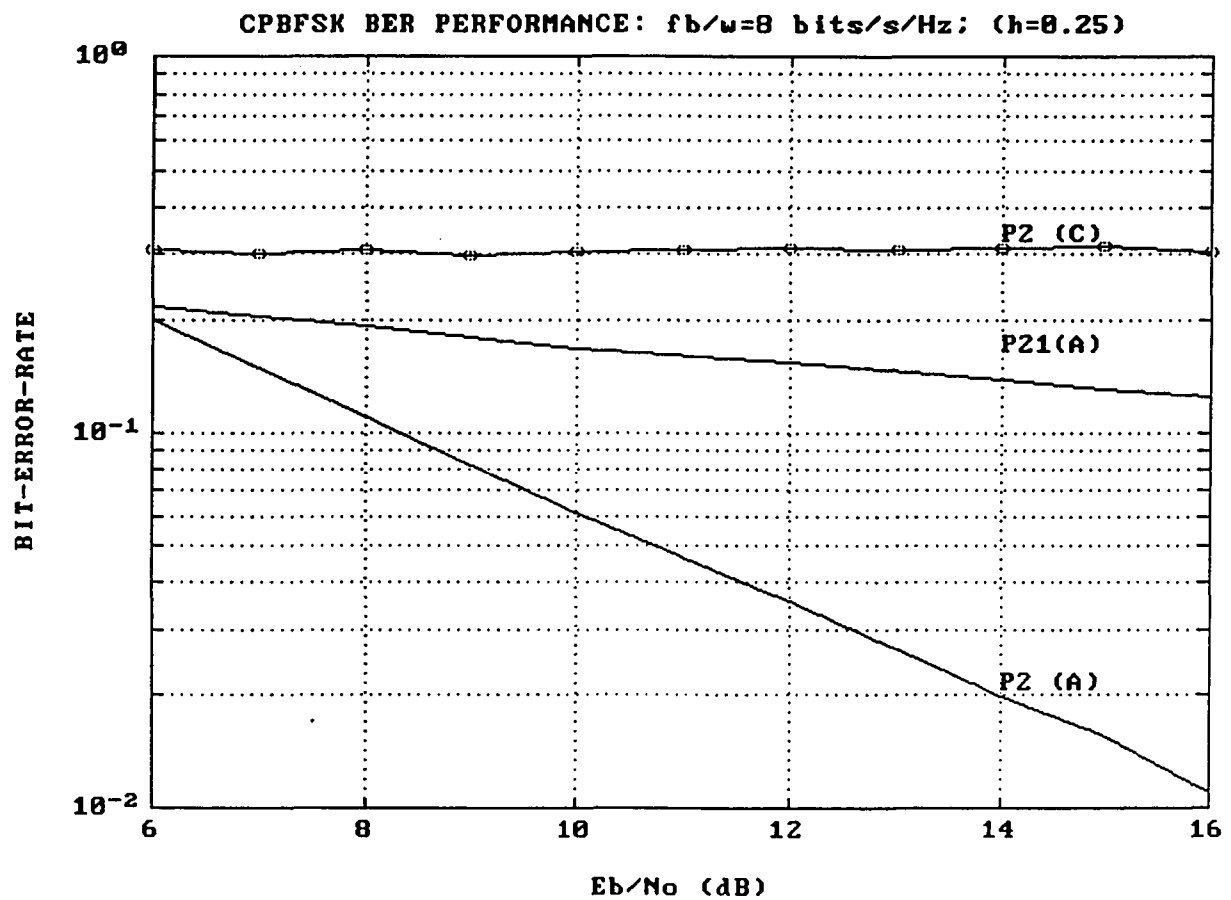


FIGURE 4.20.3: COMPARISON OF CPBFSK BER PERFORMANCES FOR THE CONVENTIONAL AND ADAPTIVE THRESHOLD DECISION METHODS WITH $f_b/w = 8 \text{ bits/s/Hz}$

(P2): FILTER TYPE IS KNOWN BY ADAPTIVE DETECTOR

(P21): FILTER TYPE IS UNKNOWN BY ADAPTIVE DETECTOR



6 APPENDIX

In this appendix a program is presented to track and lock on the correct bandwidth of IF filter when the value of the bandwidth is unknown by the nonlinear post-processor. This program uses adaptive step (ϵ) rather than constant steps.

We have a program called "EYE" which estimates the eye opening (eo) having the bandwidth of the IF filter when nonlinear post-processor is employed. We assume the value of the bandwidth is w . By computing eo and adjusting ϵ adaptively we find the maximum possible eye opening. The bandwidth which generates the largest eo is the desired bandwidth.

The following program is written in "MATLAB" which is a high-performance interactive software package entirely written in the C language, for scientific and engineering numeric computation [23].

% Nonlinear post-processor has no knowledge of IF filter bandwidth.

w = 1; % normalized bandwidth.

EYE; % estimates "eo".

ϵ = 1

w = 2; EYE; E1 = eo;

while (1 + ϵ) > 1

$\epsilon = \epsilon / 2$

while 1

E0 = E1; w = w + ϵ ; EYE; E1 = eo

if E1 < E0

break

end

end

$\epsilon = \epsilon / 2$

while 1

E0 = E1; w = w - ϵ ; EYE; E1 = eo

if E1 < E0

break

end

end

end

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