Performance of On-Off Modulated Lightwave Signals with Phase Noise

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Abstract

We analyze the performance of On-Off-Keying modulated signals in the presence of phase noise for different envelope detection structures. We obtain closed-form expressions for the error probability conditioned on a random envelope, and remove the conditioning via approximations. We also provide a tight lower bound in closed-form. We also compare the performance with that of Frequency Shift Keying.

1 Introduction

Coherent optical communication systems provide an efficient way of sharing the huge bandwidth of single mode optical fiber among different users of the future high-speed networks. Phase noise is a major problem in coherent technology, due to both spectral broadening which necessitates wider channel spacing in a network and the incomplete knowledge and the time-varying nature of the phase which makes the correct retrieval of the transmitted data bits more difficult for the receiver. On Off Keying (OOK) is an attractive modulation scheme for systems with phase noise because of the weak coupling between the noise in the phase and the information in the amplitude.

In this paper we analyze a set of Intermediate Frequency (IF) receiver structures and determine their bit error performances. These structures employ envelope detection which is insensitive to phase uncertainties. For the case where the phase uncertainty is constant over the bit period, the envelope detector is the optimal receiver [10].

Various performance analyses for OOK exist in the literature [2, 3, 4, 5, 6, 7]. A contribution of this work is to obtain closed-form expressions for the error probability conditioned on a normalized phase noisy envelope. This phase noisy envelope was first formulated by Foschini et.al. [3] in the context of the same receiver structures that we consider here. The exact error probabilities for these receivers can be obtained by taking the expectation of the conditional error probability. However a computationally attractive method to obtain the statistics of the phase noisy envelope is yet to be found, although significant progress has been achieved [7, 8]. Therefore, we use an approximation to the actual envelope for which the statistics can be easily obtained. This approximation is first introduced in [9]; it is identical to the desired envelope to the first order in phase noise strength and faithfully produces its first two moments. Having computed the density function of the approximate envelope in [9] we obtain the performance explicitly. We also provide a easily computed lower bound, and evaluate the accuracy of Gaussian approximation.

In the next section, we described the receiver structures that are considered. Then in Section 3 we present the analysis and discuss its results. We compare the performance of OOK with Frequency Shift Keying (FSK) in Section 4. Finally we present the conclusions from this work in Section 5.

2 Receiver Structures

The amplitude modulated optical field is assumed to reach the receiver in the transmitted form, i.e. we neglect the fiber dispersion. The received signal is processed by an optical heterodyne receiver. Heterodyning involves the addition of a local oscillator signal to the received signal, and the photodetection of the sum. The effect of this processing is the transformation of the signal from optical frequency to intermediate frequency. We assume that the local oscillator power is large enough to ensure that photodetection results in an addition of a white Gaussian noise process (shot noise). Thus the IF signal output $r(t)$ is given by

$$r(t) = d\,Acos\,(2\pi f_c t + \theta(t)) + n(t) \tag{1}$$

where $d = 0, 1$ is the current data bit, f_c is the IF carrier frequency, $\theta(t)$ is the combined phase noise process of the transmitter and local oscillator lasers, and $n(t)$ is the shot noise process with two-sided spectral density $N_0/2$. This signal will be the input to the IF receiver which is the main focus of this paper. We use the standard Brownian motion model for the phase noise process. According to this model, $\theta(t) = 2\pi \int_0^t \mu(\tau)\, d\tau$, where the frequency noise process $\mu(\tau)$ is white and Gaussian with spectral density $\beta/2\pi$, β is the sum of the laser linewidths.

The general structure of the IF receiver to be considered in this paper is shown in Figure 1. The received signal is first transformed to baseband via a standard quadrature demodulator. The integrators serve the purpose of limiting the additive noise power. The integration time is T' which is yet to be determined. The signal is effectively bandpass filtered around the center frequency by the use of demodulators and integrators; the filter bandwidth is $1/T'$. For a uniformly distributed phase uncertainty which is constant over the bit duration the optimum value of T' is the bit duration T [10]. In the presence of phase noise, however, the signal occupies a larger frequency band because of phase noise. This fact may necessitate a wider filter bandwidth, or equivalently smaller integration times. In order to simplify the analysis, we assume that the ratio T/T' is a positive integer M as in [3].

The outputs of the in-phase and quadrature branch integrators in Figure 1 are squared and then added to perform the envelope detection. The remainder of the signal processing depends on the value of M. For $M = 1$, the adder outputs are sampled at the end of the bit duration and the sampled value is compared to a threshold h to reach a decoding decision. This processing involves a single stage of filtering and corresponds to *conventional envelope detection* (CED). For $M \geq 2$, the bit duration consists of multiple integration windows, so a variety of signal processing options exists. A simple strategy is to sample only once per bit as before, this effectively discards all but the last one of M integration windows. We refer to this strategy as *modified envelope detection* (MED). Note that since M will be chosen so as to optimize the performance, and since for $M = 1$ MED reduces to CED, this modification is guaranteed to be at least as good. In Figure 1, MED is shown with M samples per bit duration, but the decision device discards all but the last of these samples. A better, but not optimal, processing is to average these M samples to get the input to the threshold device. This averaging can be performed by a lowpass by a lowpass filter that integrates the adder output over the bit duration. So we call this latter processing to be *double filter envelope detection* (DFED).

To summarize, we have outlined three reception strategies: conventional, modified and double filter envelope detection. The first two are single filter receivers with the difference being in their filter bandwidths. In the special case of $M = 1$ all three receivers are identical. These receiver structures were first suggested by Kazovsky et.al. in [11] in the context of multiport homodyne receivers. The remainder of this paper is devoted to the performance analysis of these receivers.

3 Performance Analysis and Results

Now we proceed to find the probability of error for the receiver structures outlined above. The method employed is similar to that of [9] in that we first condition the error probability on the phase noise process, then we remove this conditioning either by numerical integration or via analytical bounds.

We first consider the conventional envelope detection. The decision variable can be written as

$$Y = \left| \frac{A}{2} d \int_0^T e^{j\theta(t)}\, dt + n_c + jn_s \right|^2 \tag{2}$$

where n_c and n_s are independent identically distributed Gaussian random variables with zero mean and variance

$\sigma^2 = N_0 T/4$. The signal component of the decision variable is

$$X' = \left| \frac{A}{2} \int_0^T e^{j\theta(t)}\, dt \right|^2 .$$

When $d = 0$, the decision variable has an exponential distribution given by

$$p_Y(y \mid 0) = \frac{1}{2\sigma^2} e^{-y/2\sigma^2} \quad y \geq 0 .$$

For $d = 1$, both additive noise and phase noise are present. We first condition the density on the phase noisy signal component X' to obtain

$$p_Y(y|X') = \frac{1}{2\sigma^2} e^{-(y+X')/2\sigma^2} I_0 \left(\frac{\sqrt{X'y}}{\sigma^2} \right)$$

which is a well known noncentral Chi-square distribution with two degrees of freedom.

Let $P_e(0)$ be the probability of error when $d = 0$, and let $P_e(1|X')$ be the conditional probability of error given $d = 1$ and X'. Then we have

$$\begin{aligned} P_e(0) &= \Pr[Y > h | d = 0] \\ &= \int_h^\infty \frac{1}{2\sigma^2} e^{-y/2\sigma^2}\, dy = e^{-h/2\sigma^2} \end{aligned}$$

and

$$\begin{aligned} P_e(1|X') &= \Pr[Y \leq h | d = 1, X'] \\ &= \int_0^h \frac{1}{2\sigma^2} e^{-(y+X')/2\sigma^2} I_0 \left(\frac{\sqrt{X'y}}{\sigma^2} \right) dy \\ &= 1 - Q\left(\sqrt{\frac{X'}{\sigma^2}}, \sqrt{\frac{h}{\sigma^2}} \right) \end{aligned}$$

where $Q(\cdot,\cdot)$ is the Marcum's Q function defined as

$$Q(a,b) \triangleq \int_b^\infty e^{-(a^2+x^2)/2} I_0(ax) x\, dx .$$

It is convenient to define a normalize threshold $\bar{h}$ as

$$\bar{h} = \frac{h}{2\sigma^2}$$

and to rewrite X' as follows

$$X' = \left| \frac{A}{2} \int_0^T e^{j\theta(t)}\, dt \right|^2 = \frac{A^2 T^2}{4} \left| \int_0^1 e^{j\sqrt{\gamma}\psi(t)}\, dt \right|^2 .$$

where $\psi(t)$ is the standard Brownian motion ($E\,\psi^2(t) = t$) and γ is the phase noise strength defined as $\gamma \triangleq 2\pi\beta T$. Now with the definition of normalized phase noisy envelope $X(\gamma)$ as

$$X(\gamma) = \left| \int_0^1 e^{j\sqrt{\gamma}\psi(t)}\, dt \right|^2 \tag{3}$$

we have $\frac{X'}{2\sigma^2} = \xi X(\gamma)$ where $\xi = A^2 T/2N_0$ is the signal to noise ratio (SNR). Then the error probabilities become

$$\begin{aligned} P_e(0) &= e^{-\bar{h}} \\ P_e(1|X(\gamma)) &= 1 - Q\left(\sqrt{2\xi X(\gamma)}, \sqrt{2\bar{h}} \right) \end{aligned}$$

and the unconditional error probability is given by

$$P_e = \frac{1}{2}e^{-\bar{h}} + \frac{1}{2}\left[1 - E\left[Q\left(\sqrt{2\xi X(\gamma)}, \sqrt{2\bar{h}}\right)\right]\right] \tag{4}$$

which reduces to the well known OOK error probability for $\gamma = 0$.

If the probability density function of $X(\gamma)$ is explicitly known, then the expectation in (4) can be evaluated to obtain the error probability. It will seen that only this density is required for the other receiver structures as well. The phase noisy envelope plays the same critical role for envelope detection of phase noisy FSK [3, 9]. Because of this considerable effort has gone into characterizing its statistics. While there is progress in this direction [7, 8], the techniques known so far are too complicated. Therefore, we pursue a different approach which attempts to approximate the random variable $X(\gamma)$ with a random variable whose statistics can be easily obtained. This approach of random variable fitting was first used in [3] where a linear approximation was used with $X_L(\gamma) = 1 - \gamma X'(0)$. ($X'(0)$ is the derivative of X at $\gamma = 0$, it is given in terms of the normalized Brownian motion as $\int_0^1 \left(\psi(t) - \int_0^1 \psi(\tau)\, d\tau\right)^2 dt$.) This approximation retains the properties of the original random variable for small γ. However, since it can take on negative values it yields poor results when used in conjunction with the conditional error probabilities. In order to remedy this situation, another approximation was developed in [9] which uses the random variable $X_E(\gamma) = \exp(-\gamma X'(0))$. This variable is still a faithful replica of the original random variable to the first order in γ, it also has the same range $(0,1)$. For a detailed discussion of the properties of this exponential approximation the reader is referred to [1, 9].

The approximate probability density function, $p_\gamma(x)$, of $X(\gamma)$ was obtained in [9]. Therefore the error probability in (4) can be evaluated numerically. The results are shown in Figure 2 for different γ values. The threshold $\bar{h}$ has been optimized at each point to minimize the error probability. It is seen that the performance of the conventional envelope detector deteriorates rapidly with the introduction of phase noise.

Next we consider the modified envelope detector (MED) . The analysis for this case is very similar to that of the previous case, since the decision variable consists of a single sample as well. The additive noise variances increase by a factor of M, while the effective phase noise strength decreases by the same factor. This is due to the reduction of the integration period by M. As a result, the error probability of MED is given by

$$P_e = \frac{1}{2}e^{-\bar{h}} + \frac{1}{2}\left[1 - E\left[Q\left(\sqrt{2\xi X(\gamma/M)/M}, \sqrt{2\bar{h}}\right)\right]\right] \tag{5}$$

The value of M that minimizes P_e must be found as well as the normalized threshold $\bar{h}$. The former optimization poses a computational problem since the evaluation of the density function $p(x)$ for many values of γ/M is likely to be difficult. However due to the exponential approximation for $X(\gamma)$ introduced in [9], this problem can be avoided as follows. We have $p_{\gamma/M}(x) = p_\gamma(x^M)Mx^{M-1}$ which results in

$$E\left[f\left(X(\gamma/M)\right)\right] = \int_0^1 f(x^{1/M})p_\gamma(x)\, dx$$

for any function $f(\cdot)$ defined on $(0,1)$. Thus

$$P_e = \frac{1}{2}e^{-\bar{h}} + \frac{1}{2}\left[1 - \int_0^1 Q\left(\sqrt{2\xi x^{1/M}/M}, \sqrt{2\bar{h}}\right) p_\gamma(x)\, dx\right] \tag{6}$$

which requires only one density function, instead of possibly many. The results are shown in Figure 3. Again we see that the performance degradation due to phase noise is severe. The modified receiver introduces a small improvement over the conventional receiver at no additional hardware complexity. All that is needed is an estimate of the signal to noise ratio and the phase noise strength, and an a priori computation of the optimum bandwidth expansion factor M.

Finally we consider the performance of the double filter envelope detection. In this case, the decision variable is a sum of M statistically independent random variables, $\{Y_k\}$, under both hypotheses. The additive noise components are Gaussian with variance $\alpha^2 = N_0T/4M$. It is convenient to normalize both the decision variable and the threshold as $V = Y/2\alpha^2$, $\bar{h} = h/2\alpha^2$. Then for $d = 0$, V is the sum of squares of $2M$ Gaussian random variables, thus it has a Gamma density given as

$$p_V(v) = \frac{v^{M-1}}{(M-1)!}e^{-v} .$$

For $d = 1$, we first condition on the phase noise process, and obtain the Chi-square distribution with $2M$ degrees of freedom

$$p_V\left(v \mid \{\theta(t)\}\right) = \left(\frac{v}{r}\right)^{(M-1)/2} e^{-(v+r)} I_{M-1}(\sqrt{4rv})$$

where the dependence on the phase noise process is exhibited in the parameter r. This parameter is defined as

$$r = \frac{1}{2\alpha^2}\sum_{k=1}^{M}|x(k)|^2$$

where

$$x(k) = \frac{A}{2}\int_{(k-1)T/M}^{kT/M} e^{j\theta(t)}\,dt \quad k = 1,\ldots,M.$$

It can be easily seen that r can be expressed as

$$r = \frac{\xi}{M}\sum_{k=1}^{M} X_k(\gamma/M) \tag{7}$$

where $X_k(\cdot)$ are independent observations of the random variable $X(\cdot)$. The parameter r may be regarded as a phase noisy signal to noise ratio. It is always less than the actual SNR ξ, and approaches ξ with probability 1 as $M \to \infty$ due to both the central limit theorem and the fact that the means of X_k approach 1 as the phase noise strength vanishes. Interestingly, the dependence of error probability on phase noise is through a single random variable r. The error probability for $d = 0$ is easily seen to be

$$P_e(0) = \sum_{k=0}^{M-1}\frac{\bar{h}^k}{k!}e^{-\bar{h}}$$

while the error probability for $d = 1$ conditioned on r is obtained as

$$\begin{aligned} P_e(1|r) &= 1 - \int_{\bar{h}}^{\infty}\left(\frac{v}{r}\right)^{(M-1)/2} e^{-(v+r)} I_{M-1}(\sqrt{4rv})\,dr \\ &= 1 - Q_M(\sqrt{2r}, \sqrt{2\bar{h}}) \end{aligned}$$

where $Q_M(\cdot,\cdot)$ is the generalized Marcum's Q function [12] of order M defined as

$$Q_M(a,b) \triangleq \int_b^{\infty}\frac{x^M}{a^{M-1}}e^{-(x^2+a^2)/2} I_{M-1}(ax)\,dx .$$

Note that there is no symmetry between $P_e(0)$ and $P_e(1)$ in OOK, unlike FSK [3, 9]. For a given r, $P_e(1|r)$ increases with M, as well as $P_e(0)$. On the other hand, for a given M, $P_e(1|r)$ decreases with r. However, r is a function of M as well, and it increases with probability 1 with M. Therefore there is a tradeoff in choosing M, between the additive and phase noise processes. The optimal values of M will be lower than those of FSK, since $P_e(0)$ is uniformly increasing in M.

The unconditional probability of error is given as

$$P_e = \frac{1}{2}\sum_{k=0}^{M-1}\frac{\bar{h}^k}{k!}e^{-\bar{h}} + \frac{1}{2}\left[1 - E\left[Q_M\left(\sqrt{2r}, \sqrt{2\bar{h}}\right)\right]\right] \tag{8}$$

which is to be optimized over M and $\bar{h}$. Since r is a sum of M independent identically distributed random variables, the expectation involves M-fold self-convolution of the density $q_{\gamma/M}(x)$. Thus we have, once again, the problem of many density computations as well as many convolutions. For MED, this problem could be solved by a change of variable; for double filter FSK, the form of the conditional error probability allowed us eliminate the problem as well [9]. However, this does not seem possible in this case, due to the fact that generalized Marcum's Q function does not have a compact explicit representation (for example, a finite series). Therefore, we will use three different techniques to obtain a reliable estimate of the error probability. First, we use Jensen's inequality to obtain a lower bound, which provides a rather tight estimate of the error probability for FSK [9]. Here we observe that the quantity $P_e(1|r)$ is a convex $\cup$ function of r for a given threshold $\bar{h}$, so that an interchange of the expectation operator with the function Q_M will result in a lower bound to the error probability. The lower bound corresponds to the error probability in the optimistic scenario where the random variable r does not deviate from its mean. The resulting bound is

$$P_e \geq \frac{1}{2}\sum_{k=0}^{M-1}\frac{\bar{h}^k}{k!}e^{-\bar{h}} + \frac{1}{2}\left[1 - Q_M\left(\sqrt{2\xi\bar{X}(\gamma/M)}, \sqrt{2\bar{h}}\right)\right] \tag{9}$$

where $\bar{X}(\cdot)$ denotes the mean of $X(\cdot)$. This mean can be easily computed as [9]

$$\bar{X}(\gamma) = \frac{4}{\gamma}\left[1 - \frac{2}{\gamma}(1 - e^{-\gamma/2})\right] . \tag{10}$$

Equation (9) in conjunction with (10) describe a lower bound that we will use in the following discussion.

The second technique to be employed in estimating (8) is the Chernoff bound which yields

$$\Pr(V \leq \bar{h}) \leq e^{s\bar{h}}E(e^{-sV})$$

for all $s \geq 0$. This provides an upper bound to $P_e(1)$ in terms of the moment generating function of the decision variable under $d = 1$. We note that the moment generating function of V is given by [10]

$$E\left[e^{-sV} \mid r\right] = \frac{1}{(1+s)^M}\exp\left(-\frac{sr}{1+s}\right) .$$

Since r is a sum of M independent, identically distributed random variables, the conditioning on r is removed to yield

$$E\left[e^{-sV}\right] = \frac{1}{(1+s)^M}\left[E\left[\exp\left(-\frac{s\xi}{M(1+s)}X(\gamma/M)\right)\right]\right]^M$$

and finally the Chernoff bound is given by

$$P_e \leq \frac{1}{2}\sum_{k=0}^{M-1}\frac{\bar{h}^k}{k!}e^{-\bar{h}} + \frac{1}{2}\frac{1}{(1+s)^M}\left[\Phi_{\gamma/M}\left(\frac{s\xi}{M(1+s)}\right)\right]^M \tag{11}$$

where we denote the moment generating function of $X(\gamma)$ by $\Phi_\gamma(s) = E(e^{-sX(\gamma)})$. Lastly we relate $\Phi_{\gamma/M}(\cdot)$ to $q_\gamma(\cdot)$ as

$$\Phi_{\gamma/M}(s) = \int_0^1 \exp(-sx^{1/M})q_\gamma(x)\,dx \tag{12}$$

which eliminates the need for computation of many density functions. An additional optimization needs to be performed over nonnegative s to obtain the tightest upper bound.

The Jensen bound and the Chernoff bound are shown in Figure 4 for various γ values. Note that the bounds are very close for all values of γ and ξ. Therefore, the computationally simpler lower bound may be used reliably.

A third approach in estimating the performance of the double filter receiver is Gaussian approximation. Since the normalized decision variable V is the sum of M independent identically distributed random variables under both hypotheses, it is tempting to use a Gaussian approximation for V. For $d=0$, both the mean and the variance of V are easily obtained to be M. For $d=1$, the conditional mean of V is $M+r$, while its conditional variance is $M+2r$. Therefore we have with the Gaussian approximation

$$\begin{aligned} P_e(0) &\simeq Q\left(\frac{\bar{h}-M}{\sqrt{M}}\right) \\ P_e(1|r) &\simeq Q\left(\frac{M+r-\bar{h}}{\sqrt{M+2r}}\right) . \end{aligned}$$

Due to the difficulty of removing the conditioning exactly, we will further approximate $P_e(1)$ via a Jensen-type interchange by

$$P_e(1) \simeq Q\left(\frac{M+E(r)-\bar{h}}{\sqrt{M+2E(r)}}\right) .$$

Under these approximations the error probability becomes

$$P_e \simeq \frac{1}{2}Q\left(\frac{\bar{h}-M}{\sqrt{M}}\right) + \frac{1}{2}Q\left(\frac{M+\xi\bar{X}(\gamma/M)-\bar{h}}{\sqrt{M+2\xi\bar{X}(\gamma/M)}}\right) \tag{13}$$

which is to be optimized over M and $\bar{h}$. This is the error probability of a system in which the decision variables are Gaussian with nonidentical variances. The optimal setting of the threshold is complicated. A conventional threshold setting is one that equalizes the two error probabilities. This results in an error probability of

$$P_e \simeq Q\left(\frac{\xi\bar{X}(\gamma/M)}{\sqrt{M}+\sqrt{M+2\xi\bar{X}(\gamma/M)}}\right) \tag{14}$$

where the underlying threshold setting is

$$\bar{h}_G = M + \frac{\sqrt{M}\xi\bar{X}(\gamma/M)}{\sqrt{M}+\sqrt{M+2\xi\bar{X}(\gamma/M)}} . \tag{15}$$

It is known that this nonoptimal threshold setting gives results that are very close to the optimal threshold setting for the Gaussian approximation [13]. The error probability predicted by the Gaussian approximation is compared with the previously obtained lower bound in Figure 5. For each γ value the Gaussian curve is the upper curve. It is seen that the two results are in good agreement uniformly over the set of γ and ξ values. Therefore we conclude that the lower bound given in Equation (9) satisfactorily predicts the performance of the double filter receiver.

The closeness of the error probability prediction of the Gaussian approximation to the exact performance was shown in [13] in the absence of phase noise. It was also noticed that the Gaussian threshold estimate is much lower than the actual optimal threshold. If the threshold is set according to the Gaussian approximation, the resulting performance would be far worse than what is predicted. This rather surprising result is true even when the value of M is large [13]. The presence of phase noise does not help correct this discrepancy. In Figure 6 we show the threshold predictions of the Gaussian approximation and the Jensen bound. It is seen that threshold predictions remain very different.

The performance predicted by this work is considerably better than that predicted in [3] for conventional and modified envelope detectors when γ is large, while the results agree closely for the double filter performance. For the latter receiver the optimal value of M is large enough to ensure small effective phase noise strength γ/M; this ensures that the exponential and linear approximations are close. On the other hand, the value of M is much smaller for the single filter receivers, this causes the linear approximation to overestimate the error probability.

4 Comparison of OOK and FSK with Envelope Detection

A set of error probability curves of double filter envelope detection of OOK for a larger collection of γ values is given in Figure 7 for improved visual clarity. It is seen from the figure that at a bit error rate of 10^{-9}, the SNR penalty due to phase noise is 0.25 dB for $\gamma = 1$, 0.38 dB for $\gamma = 2$, 0.62 dB for $\gamma = 4$, 0.75 dB for $\gamma = 6$, and 1.13 dB for $\gamma = 16$. The robustness of OOK to phase noise surpasses that of FSK in terms of required excess signal energy to maintain a certain error probability level. This is seen clearly in Figure 8 which shows the penalty in SNR as a function of phase noise strength for both OOK and FSK at an error probability of 10^{-9}. The improved robustness does not mean improved error performance, however. Since semiconductor lasers are limited in the peak power they can generate, as opposed to being average power limited, envelope detection of OOK will always have an inferior error probability with respect to envelope detection of orthogonal signals. This is demonstrated in Figure 9 for $\gamma = 0$ and $\gamma = 1$. The reason behind this conclusion can be seen as follows. In orthogonal signaling, envelope detection is performed for two frequencies, resulting in two decision variables. If one of the decision variables is neglected, and the other is compared to a fixed threshold for the decision, then the resulting performance will be identical to that of OOK. However, this clearly is a suboptimal processing of the available decision variables; symmetry dictates a comparison of the two decision variables for deciding on the most likely hypothesis. Nonetheless, OOK is still a modulation format of interest; the choice between OOK and FSK must be based on factors such as the amount of desired simplicity for the transmitters and the receivers, the achievability of orthogonality (i.e. wide frequency separation) in FSK, the severity of frequency dispersion during propagation along the fiber, etc.

5 Conclusions

We have provided a performance analysis for different envelope detection structures when the signal is OOK modulated. This analysis is different from previously reported ones in that it provides exact closed-form conditional error probabilities. The only obstacle to the final exact result remains the statistics of a normalized random envelope which is currently receiving considerable attention. Here we have approximated this envelope by another random variable to obtain the error performance. This enables us to compare the three envelope detection strategies and to compare OOK with FSK with similar receivers. Double filter envelope detection provides robustness against phase noise. While OOK is more robust in terms of phase noise induced degradation, FSK has a better performance for peak power limited semiconductor lasers.

6 Acknowledgements

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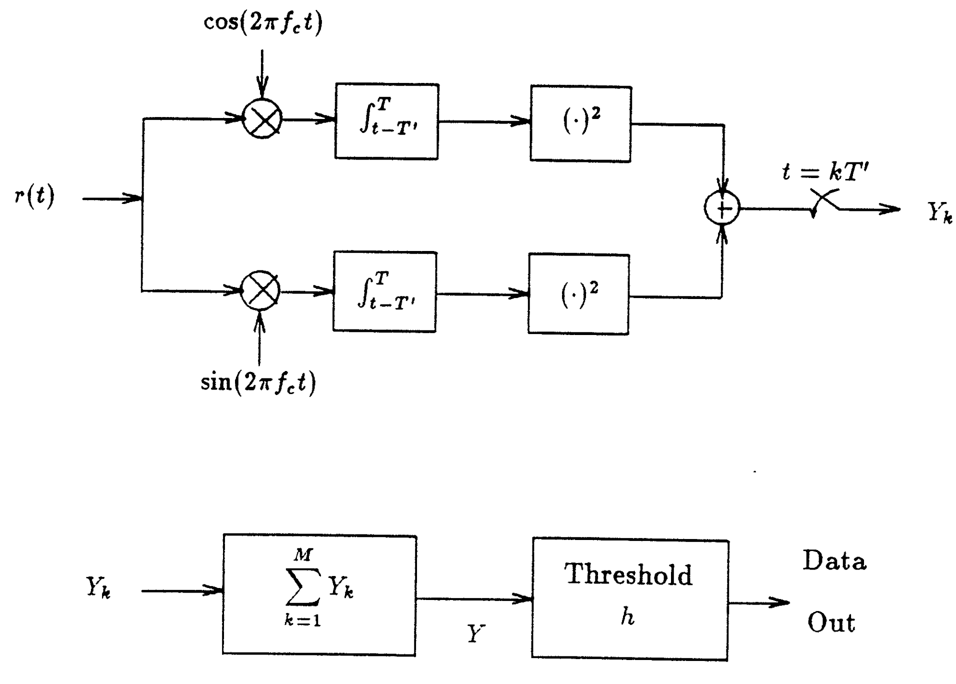

Figure 1: IF receiver for envelope detection of OOK signals. For single filter receivers the adder is absent, and $Y = Y_M$.

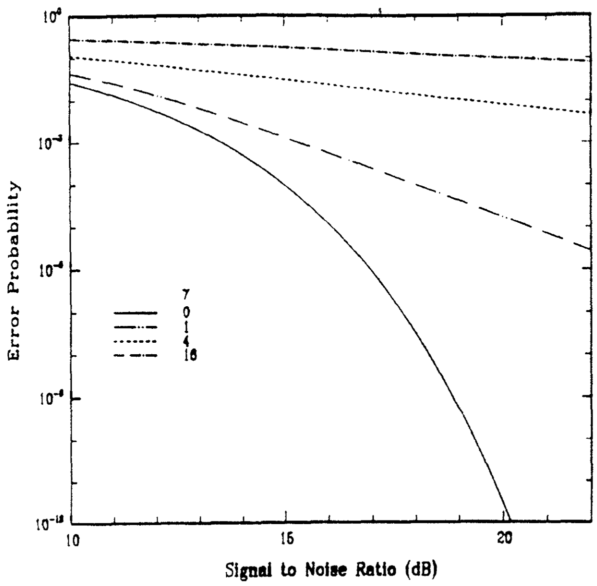

Figure 2: Error probability for the conventional envelope detector.

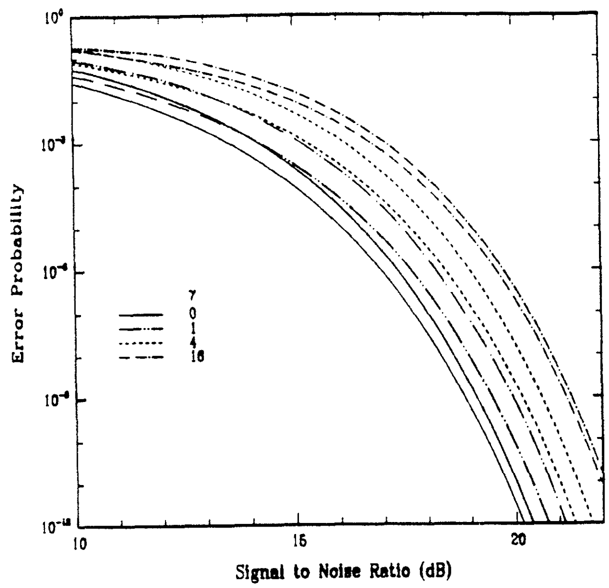

Figure 4: Comparison of Jensen and Chernoff bounds for the double filter envelope detector.

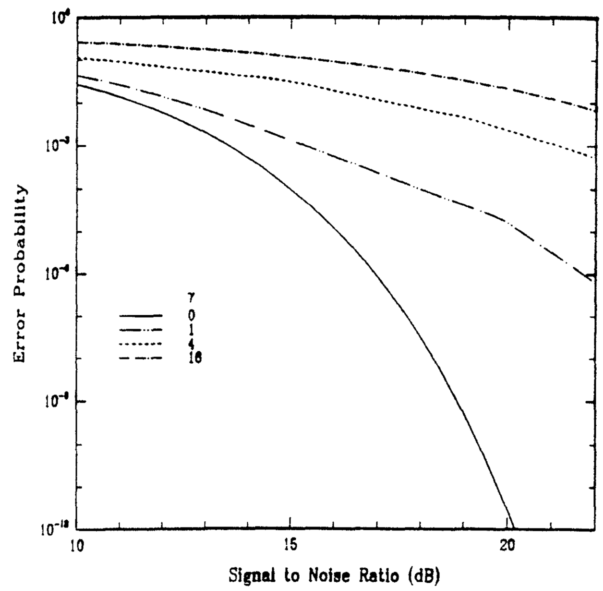

Figure 3: Error probability for the modified envelope detector.

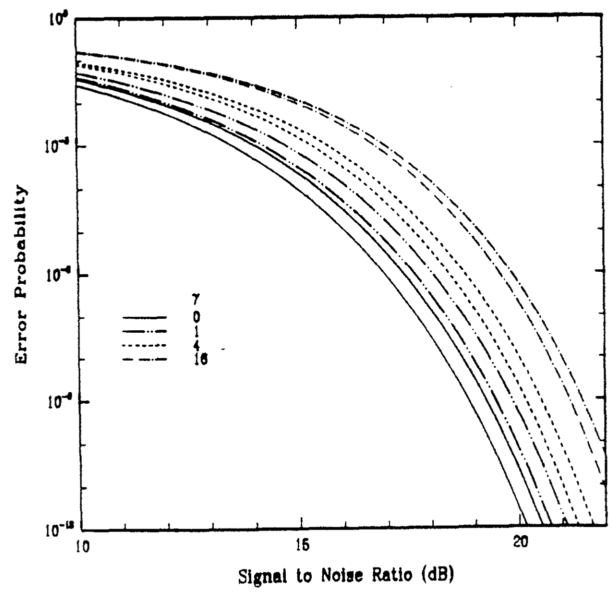

Figure 5: Comparison of Gaussian approximation and the lower bound for the double filter receiver.

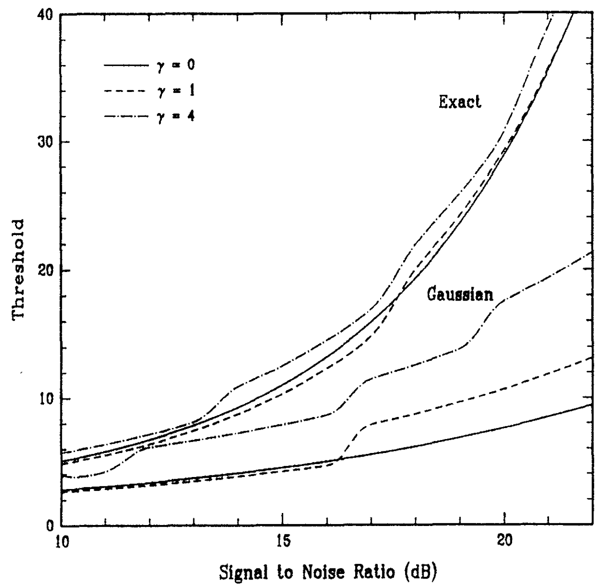

Figure 6: Optimal thresholds as functions of signal-to-noise ratio for various γ values.

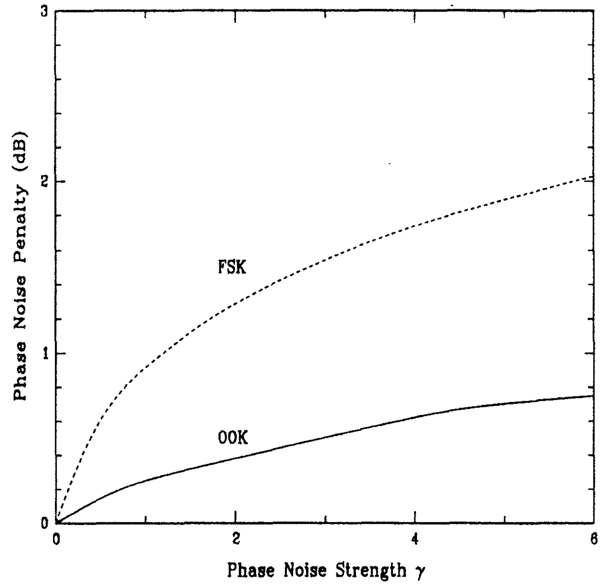

Figure 8: Comparison of SNR penalties for double-filter OOK and FSK due to phase noise.

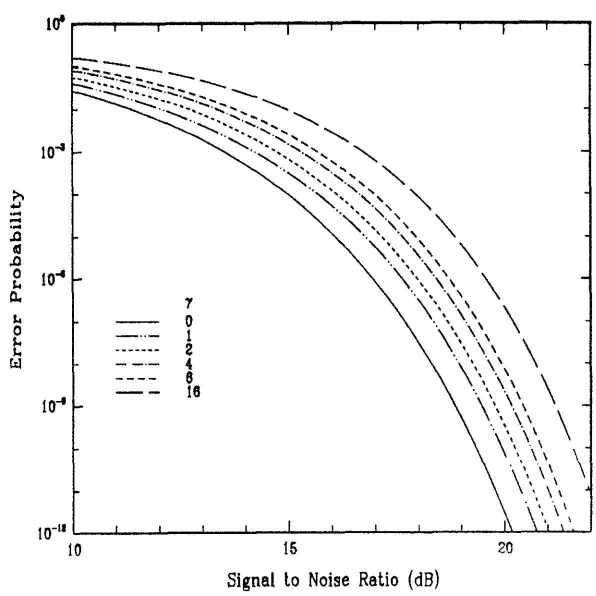

Figure 7: Estimated performance of the double filter receiver.

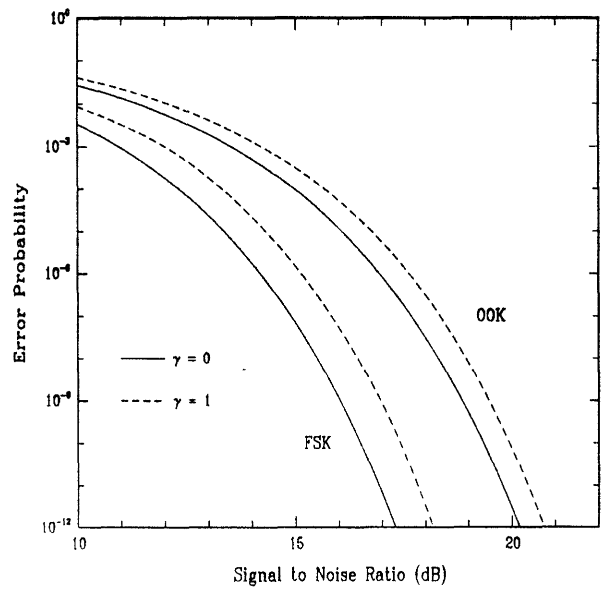

Figure 9: Comparison of error probabilities of double-filter OOK and FSK for $\gamma = 0$ and $\gamma = 1$.