A COMPARISON OF COHERENT AND DIFFERENTIALLY COHERENT DETECTION SCHEMES FOR FADING CHANNELS

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ABSTRACT

Regarding carrier recovery as the estimation of the fading distortion we reveal the common basis of coherent and differentially coherent detection. In the differentially coherent receiver a very simple estimate of the fading distortion is used whereas the coherent receiver uses the optimal estimate [1]. Further, the bit error rates for M-ary PSK and DPSK transmission are calculated using a single method of calculation for both detection schemes. The calculation takes into account non-perfect carrier recovery, cochannel interference, and diversity. The results allow a direct comparison of the two schemes and show that coherent detection is preferable in many realistic fading environments.

I. INTRODUCTION

It is well known that coherent detection schemes are superior to differentially coherent or noncoherent schemes in terms of power efficiency, if a stable carrier reference can be established. However, carrier recovery is the key problem on a fading channel with its amplitude fluctuations and random frequency modulation. Weber studied the performance of a PLL on a fading channel quantitatively [2]. It is common opinion that if the fading is rapid enough this precludes any phase-locked type of carrier recovery [3]. A popular alternative is to use differential detection since then the problem of acquiring a phase-tracking loop in this highly degraded environment is avoided and since it is assumed that the detection loss due to phase jitter on the carrier reference in the coherent receiver exceeds the signal-to-noise ratio penalty associated with differential detection [4]. However, differential detection suffers in the same way from channel disturbances as the coherent scheme does. E.g., a detailed analysis of differential detection of binary PSK is found in [5]. However, a direct comparison of the two detection schemes that takes into account the performance degradation of both the coherent receiver (due to a noisy phase reference) and the differentially coherent receiver (due to channel phase changes over two consecutive signalling intervals) has not yet been given.

The aim of this paper is to present such a comparison for the frequency-nonselective Rayleigh fading channel. The common origin of the two detection schemes is revealed by viewing carrier recovery as estimation of the multiplicative fading distortion. In a previous paper we found the optimal carrier recovery structure for the channel model under consideration [1]. Since this structure is linear, phenomena like hang-up and threshold effect do not occur. Therefore the error variance of the carrier recovery is a suitable measure to assess its performance. Taking into account the nonperfect carrier reference we calculate the error rate for coherent and differentially coherent detection of M-ary PSK by applying Stein's method [6]. In addition to thermal noise we consider cochannel interference as a further source of degradation. Besides the results for the coherent receiver the exact formulas for the error rate of differential detection for M = 4,8 seem to have not yet been published.

This paper is organized as follows. After the description of the channel model and the coherent receiver in section II we show the common basis of coherent and differentially coherent detection (section III). Section IV contains the error rate calculation and in section V the performance of the two detection schemes is compared.

II. CHANNEL MODEL AND RECEIVER STRUCTURE

We consider linearly phase modulated signals transmitted over a frequency-nonselective Rayleigh fading channel in the presence of both thermal noise and multiple cochannel interference. The received signal, in complex baseband notation, is

$$r(t) = h(t)\sqrt{E_s/T}\sum_{i=0}^{\infty} a_i g(t-iT) + \tilde{n}(t) + \sum_{n=1}^{N} h_n(t-\tau_n)\sqrt{E_{sn}/T}\sum_{i=0}^{\infty} a_i^n g(t-iT-\tau_n) \quad (1)$$

where a_i is the i-th transmitted symbol and $g(t)$ is the signal pulse. With the normalization $\int g(t)g^*(t)dt = T$ and $|a_i| = 1$, E_s denotes the energy per symbol. T is the symbol period, and $\tilde{n}(t)$ is complex additive white gaussian noise with

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doublesided spectral density N_0. The fading distortion h(t) is a zero mean complex gaussian noise process with unity power. The last term in eq. (1) is caused by cochannel interference. E_{sn}, $h_n(t)$, a_i^n, τ_n represent the energy, the fading distortion, the i-th transmitted symbol and the time shift of the n-th interferer relative to the time scale of the useful signal, respectively. We assume that the fading distortions $h_n(t)$, n = 1,..,N, are independent of each other and of h(t) and have the same statistical properties as h(t).

In the receiver the incoming signal is filtered by a matched filter with impulse response $g^*(-t)$. Sampling the matched filter output at t = kT and using a normalization to simplify later calculations gives[1]

$$z(k) := \frac{1}{\sqrt{E_s/T}} \frac{1}{T} \int r(t) g^*(t-kT) dt \tag{2}$$

$$= h(k)a_k + n(k) + \sum_{n=1}^{N} \sqrt{E_{sn}/E_s}\, \frac{1}{T} \int h_n(t-\tau_n) \sum_{i=0}^{\infty} a_i^n g(t-iT-\tau_n) g^*(t-kT) dt \tag{3}$$

with

$$h(k) = \frac{1}{T} \int h(t) |g(t-kT)|^2 dt \tag{4}$$

If h(t) is constant during the integration interval then the sampled matched filter output contains all the information of the continuous-time signal that is relevant for detection. n(k) is complex white gaussian noise with variance $R = N_0/E_s$. Here z(k) is denoted as received signal sample. From eq. (3) it follows that the degradation caused by the cochannel interferers depends on their time shifts τ_n, n = 1,..,N. In the case of not too severely bandlimited pulses maximum interference occurs for $\tau_n = 0$, n = 1, ..N. Using this in (3) one obtains

$$z(k) = h(k)a_k + n(k) + \sum_{n=1}^{N} \sqrt{E_{sn}/E_s}\, a_k^n h_n(k) \tag{5}$$

Due to the randomness of the interfering symbols the cochannel interference induced term in (5) is white gaussian noise with variance Σ_{hh}/CIR where

$$\Sigma_{hh} = E[|h(k)|^2]$$
$$= \frac{1}{T^2} \iint |g(t)|^2 |g(\tau)|^2 \rho_h(t-\tau) dt d\tau \tag{6}$$

and CIR is the carrier-to-interference ratio

$$CIR = E_s / \left(\sum_{n=1}^{N} E_{sn} \right) \tag{7}$$

$\rho_h(\tau)$ is the (normalized) autocorrelation function of the fading distortion

$$\rho_h(\tau) = E[h(t+\tau) h^*(t)] \tag{8}$$

In [1] we derived the maximum a posteriori detector of the transmitted symbol sequence in the case of absence of cochannel interference. However since cochannel interference causes an additional white noise term which can be combined with n(k) to form an overall white noise term of variance $(R + \Sigma_{hh}/CIR)$, the results of [1] are still valid. The derivation showed that carrier recovery equals the estimation of the fading distortion h(k), a result that was perviously found by Kam and Teh [7]. If a linear state space model of the fading distortion exists then this signal model determines the Kalman filter as the optimal state estimator [8] and thus as the optimal carrier recovery unit. Fig. 1 shows a receiver structure that employs this carrier recovery and uses a decision-directed elimination of the symbol phase for [illegible] rier synchronization. To avoid phase amb[illegible] the information is associated with [illegible] transitions, i.e. a_k is the output of [illegible]tial encoder which maps the unco[illegible]sequence $\{\alpha\}$ onto $\{a\}$ according to

$$a_k = a_{k-1} \alpha_k \tag{9}$$

In the receiver the reverse operation must be performed (see fig. 1). For M-ary phase shift keying a_k (and α_k) is element of the set $\{\exp(j0), \exp(j2\pi/M), \ldots, \exp(j2\pi(M-1)/M)\}$. We assume that each symbol is equally likely and that the symbol sequence is white.

Note that the receiver presented is easily implemented digitally since a fixed oscillator is used in the receiver and all synchronizer and detector operations are performed on the sampled matched filter output (no VCO is needed).

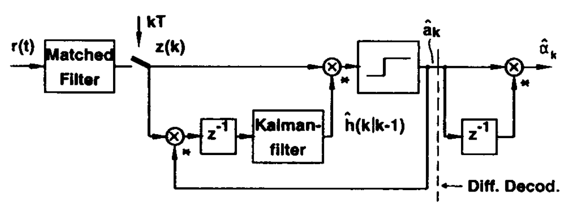

Fig.1: Blockdiagram of coherent receiver

III. DIFFERENTIALLY COHERENT DETECTION

Let us assume absence of modulation for the moment (i.e. $a_k = a_k^1 = .. = a_k^N = 1$ for all k). The Kalman filter delivers the optimal estimate $\hat{h}(k|k-1)$ for

[1] We assume perfect bit timing and absence of intersymbol interference.

$h(k)$ given all received signal samples up to time step $(k-1)$: $Z(k-1) = \{z(0), \ldots, z(k-1)\}$:

$$\hat{h}(k|k-1) = E[h(k)|Z(k-1)] \tag{10}$$

However other, suboptimal estimators are possible. Let us assume that only the last received signal sample, $z(k-1)$, is available to obtain an estimate for $h(k)$. Since $h(k)$ and $z(k-1)$ are jointly gaussian the optimal estimate in the sense of minimal mean square error

$$\hat{h}'(k) = E[h(k)|z(k-1)] \tag{11}$$

is linear [8]:

$$\hat{h}'(k) = \Sigma_{hz}\Sigma_{zz}^{-1}\, z(k-1) \tag{12}$$

Using (3) and (4) one finds

$$\Sigma_{hz} = E[h(k)z^*(k-1)] = E[h(k)h^*(k-1)]$$
$$= \frac{1}{T^2}\iint |g(t)|^2 |g(\tau)|^2 \rho_h(T+t-\tau)\,dt\,d\tau \tag{13}$$

$$\Sigma_{zz} = E[|z(k-1)|^2] = \Sigma_{hh} + R + \Sigma_{hh}/CIR \tag{14}$$

The error variance of the estimate (11) is

$$\sigma^2 = E[|h(k) - \hat{h}'(k)|^2]$$
$$= \Sigma_{hh} - \Sigma_{hz}\Sigma_{zz}^{-1}\Sigma_{zh} \approx 1 - \frac{\rho_h^2(T)}{1+R+1/CIR} \tag{15}$$

where we used the approximations

$$\Sigma_{hh} \approx 1 \tag{16a}$$
$$\Sigma_{hz} \approx \rho_h(T) \tag{16b}$$

which are valid if $h(t)$ is approximately constant for the length of the signal pulse. Of course the estimation error variance (15) can never be smaller than that of the Kalman filter, however the actual difference depends on the signal model.

Now let us use the estimator (11) as the carrier recovery unit. For phase modulation schemes the error rate of the receiver only depends on the phase of the recovered carrier. Since the real constant $\Sigma_{hz}\Sigma_{zz}^{-1}$ does not affect this phase, the error rate of a receiver that uses $\hat{h}''(k) = z(k-1)$ instead of $\hat{h}'(k)$ will be the same. The receiver that uses this simple carrier recovery is equal to the receiver of fig. 1 except that the Kalman filter is omitted.

The receiver structure can be further simplified by shifting the multiplication with $\hat{a}_k^*$ behind the decision stage. It can be shown that this does not affect the overall symbol decision $\hat{\alpha}_k$. Fig. 2 shows the resulting receiver structure. The operations performed after the decision device are just differential encoding and differential decoding which cancel out and thus can be omitted.

What remains is nothing else but a baseband realization of the optimal differentially coherent detector. The matched filter is the optimal receiving filter (see remarks after eq. (4)). Note that successive noise samples $n(k)$ are independent. In other publications often a somewhat vague description of the receiving filter is used ("filter bandwidth such that additive noise is suppressed but signal waveform remains undistorted"). Here we have found a common basis for the description of coherent and differentially coherent detection. The differentially coherent detector is just a special case of the coherent detector where a very simple carrier recovery is used. Having found this common basis we are now able to compare the two schemes with each other in terms of bit error rate using the same method of calculation for both receivers.

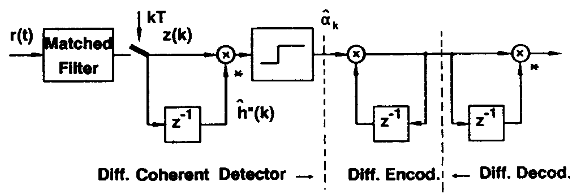

Fig.2: Baseband realization of differentially coherent receiver

IV. BIT ERROR RATE OF MPSK TRANSMISSION

A lower bound of the error probability of the coherent receiver is obtained when it is assumed that the symbol phase is known for synchronization and thus can be perfectly removed in front of the Kalman filter. In this case no differential encoding is needed since no phase ambiguity can occur. We have calculated this lower bound in [1] using the approximations (16) and assuming absence of cochannel interference. Here we will give up these restrictions and extend the calculation to the differentially coherent case.

Because of the rotational symmetry of the detection problem we can assume without loss of generality that $a_k = 1$ ($a_k a_{k-1}^* = 1$ for diff. coherent detection) has been transmitted. Applying Stein's result for Rayleigh fading channels Maciejko [9] showed that if the error probability can be expressed as

$$P_e = \text{Prob}\{|Y_1|^2 - |Y_2|^2 < 0\} \tag{17}$$

with

Y_1, Y_2: complex gaussian random variables then the error probability is

$$P_e = \frac{1}{2}(1 - \mu) \tag{18}$$

with

$$\mu = \frac{R_{11} - R_{22}}{\sqrt{(R_{11}+R_{22})^2 - 4R_{12}R_{21}}} \tag{19}$$

$$R_{ij} = E[Y_i Y_j^*] \tag{20}$$

E.g., the error rate of binary PSK is (see fig. 1)

$$P_{b,2PSK} = \text{Prob}(\text{Re}\{z(k)\hat{h}^*(k)\} < 0) \tag{21}$$

$$= \text{Prob}\left\{\left|\frac{z(k)+\hat{h}(k)}{2}\right|^2 - \left|\frac{z(k)-\hat{h}(k)}{2}\right|^2 < 0\right\} \tag{22}$$

where $\hat{h}(k)$ equals $\hat{h}(k|k-1)$ in the coherent and $\hat{h}''(k) = z(k-1)$ in the diff. coherent case. In [1] we showed that also the bit error rate of 4PSK and 8PSK can be cast in the desired form (17). For the computation of the error rate we now only need to calculate the correlation terms R_{ij} for the two detection schemes and insert the result into (18). The R_{ij}'s can again be easily calculated from the variances and covariances of z(k) and $\hat{h}(k)$. Note that the variance of z(k) has already been given in (14). Here we consider only Rayleigh fading. However Stein's method can also be applied to Rice fading. In that case the calculations are a bit more tedious (Marcum's Q-function is involved).

A) Coherent detection

Using the orthogonality theorem of estimation theory [8] the variance of the estimate $\hat{h}(k|k-1)$ results in

$$\Sigma_{\hat{h}\hat{h}} = E[|\hat{h}(k|k-1)|^2] = \Sigma_{hh} - \sigma^2 \tag{23}$$

and, similarly

$$\Sigma_{\hat{h}z} = E[\hat{h}(k|k-1)z^*(k)] = \Sigma_{hh} - \sigma^2 \tag{24}$$

where $\sigma^2 = E[|h(k) - \hat{h}(k|k-1)|^2]$ is the stationary solution of the Ricatti equation [8].

Now the correlation terms are calculated using (14), (23), and (24). Inserting them in (19) gives the error rate of 2PSK via (18)

$$P_{b,2PSK} = \frac{1}{2}\left[1 - \frac{1}{\sqrt{1 + \dfrac{\sigma^2 + \dfrac{1}{E_s/N_0} + \Sigma_{hh}/CIR}{\Sigma_{hh} - \sigma^2}}}\right] \tag{25}$$

For $\sigma^2 = 0$ and $\Sigma_{hh} = 1$ we obtain the well-known result for coherent detection when a perfect carrier reference is achievable [11]. Using the results of [1] the bit error rate of 4PSK and 8PSK can be found to be

$$P_{b,4PSK} = \frac{1}{2}\left[1 - \frac{1}{\sqrt{1 + \dfrac{2\sigma^2 + \dfrac{1}{E_b/N_0} + 2\Sigma_{hh}/CIR}{\Sigma_{hh} - \sigma^2}}}\right] \tag{26}$$

with $E_b = E_s/\log_2 M$

and

$$P_{b,8PSK} = \frac{1}{3}\left[1-\mu_1 + \frac{1}{2}(1+\mu_1)(1-\mu_2)\right] \tag{27}$$

with

$$\mu_i = \frac{\cos\alpha_i}{\sqrt{1 + \dfrac{\sigma^2(1+\sin^2\alpha_i) + \dfrac{1}{3E_b/N_0} + \Sigma_{hh}/CIR - \Sigma_{hh}\sin^2\alpha_i}{\Sigma_{hh} - \sigma^2}}}$$

$$i = 1,2$$

where

$$\alpha_1 = 3\pi/8;\ \alpha_2 = \pi/8$$

The error rates (25-27) are still a function of the performance σ^2 of the carrier recovery. However σ^2 can be calculated from the channel parameters by solving the Ricatti equation. If this dependency is inserted the error rate can be directly expressed as a function of the signal-to-noise ratio, the carrier-to-interference ratio, and the spectral characteristics of the fading.

B) Differentially coherent detection

Since $\hat{h}''(k) = z(k-1)$ the variance of the estimate $\hat{h}''$ is equal to (14). Inserting (5) and using (13) one finds for the crosscorrelation

$$\Sigma_{\hat{h}z} = E[z(k-1)z^*(k)] = \Sigma_{hz}\left[1 + \sum_{n=1}^{N} \frac{E_{sn}}{E_s} a_k^n (a_{k-1}^n)^*\right] \tag{28}$$

which depends on the interfering symbols. Therefore the error rate of the differentially coherent receiver first has to be calculated as a function of the interfering symbols. Afterwards the mean error rate is obtained by averaging over these symbols. Note that in the coherent case the Kalman filter performs this averaging operation, since for the carrier recovery the cochannel interference term is simply an additional white gaussian noise term.

Using the same method of calculation as for the coherent case one obtains

$$P_{b,2DPSK}(a) = \frac{1}{2}\left[1 - \frac{\Sigma_{hz}\left(1 + \sum_{n=1}^{N} \frac{E_{sn}}{E_s} a_k^n (a_{k-1}^n)^*\right)}{\Sigma_{hh}\left(1 + \frac{1}{CIR}\right) + \frac{1}{E_s/N_0}}\right] \tag{29}$$

In the following we assume $E_{s1} = E_{s2} = \ldots = E_{sn}$ for convenience. For a white symbol sequence with equally probable symbols the probability for each possible value of $a_k^n(a_{k-1}^n)^*$ is 1/M. In (29) we need the sum of the interferer phases. The probability that the phase exp(j0) occurs K_0-times, exp(j2π/M) occurs K_1-times etc. with the restriction that

$$\sum_{i=0}^{M-1} K_i = N \tag{30}$$

is ("generalized Bernoulli trial" [10])

$$P(K_0, K_1, \ldots K_{M-1}) = \frac{N!}{K_0! K_1! \ldots K_{M-1}!} \frac{1}{M^N} \tag{31}$$

Hence the average error probability is

$$P_{b,DPSK} = \sum P_{b,DPSK}(a) P(K_0, K_1, \ldots, K_{M-1}) \tag{32}$$

where the summation is over all combinations of K_i's that obey (30). In the case of M = 2 averaging is very simple since $P_{b,2DPSK}(a)$ depends linearly on the interferer symbols, eq. (29). Thus the averaging operation can be directly applied to the interferer symbols. Since $E[a_k^n(a_{k-1}^n)^*] = 0$ we obtain

$$P_{b,2DPSK} = E[P_{b,2DPSK}(a)] = \frac{1}{2}\left[1 - \frac{\Sigma_{hz}}{\Sigma_{hh}\left(1 + \frac{1}{CIR}\right) + \frac{1}{E_s/N_0}}\right] \tag{33}$$

For M = 4,8 the error rate (as a function of the interferer symbols) is calculated in the same way as for the coherent case. Since its dependence on the interfering symbols is no longer linear the averaging (32) does not lead to such compact results as in (33). For M = 4 and absence of cochannel interference one obtains again a quite simple result

$$P_{b,4DPSK}\Big|_{CIR \to \infty} = \frac{1}{2}\left[1 - \frac{\Sigma_{hz}}{\sqrt{2(\Sigma_{hh} + \frac{2}{E_b/N_0})^2 - \Sigma_{hz}^2}}\right] \tag{34}$$

For both detection schemes the above calculations can be extended to diversity reception. The combiner of the L independent diversity branches received that achieves best performance is the "maximal ratio combiner" [11] where the decision variable is the sum of the L matched filter output signals scaled by the optimal estimates of the complex channel gain $\hat{h}_\ell^*$, $\ell=1,..L$. This combiner can easily be implemented in our coherent receiver since a Kalman filter for each diversity branch delivers the optimal estimates $\hat{h}_\ell(k|k-1)$. In the differentially coherent case the estimates $\hat{h}_\ell''(k)$ are used instead. Using this combiner, eq. (18) must be replaced by [11]

$$P_e = \frac{1}{2}\left[1 - \mu \sum_{\ell=0}^{L-1} \binom{2\ell}{\ell} \left(\frac{1-\mu^2}{4}\right)^\ell\right] \tag{35}$$

Hence inserting μ for the detection and modulation schemes under consideration gives the performance of the coherent and differentially coherent receiver when using diversity.

V. PERFORMANCE COMPARISON

To exemplify the results, fig. 3 shows the bit error rate of unfiltered MPSK for the land mobile radio channel with its typical doppler spectrum [12] versus E_b/N_0. Here we assumed a normalized (on bit rate $1/T_b = \log_2 M/T$) cutoff frequency of the doppler spectrum of $\omega_D T_b$ = .09 (corresponds to vehicle speed v = 180 km/h, carrier freq. f_c = 900 MHz, data rate 10 ksymbols/s) and absence of cochannel interference. In the coherent receiver we

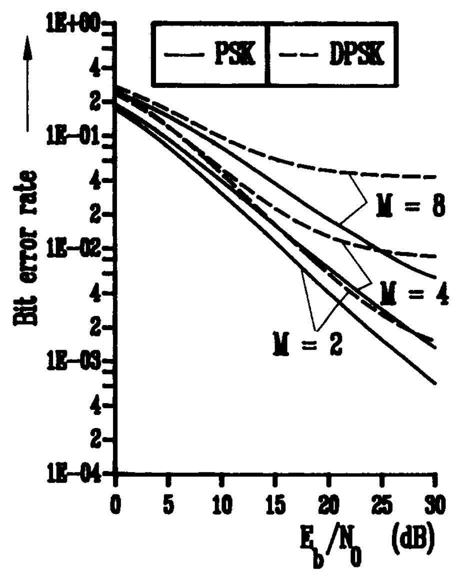

Fig.3: Bit error rate of MPSK transmission on land mobile radio fading channel ($\omega_D T_b$ = 0.09, CIR → ∞)

used a second order Kalman filter for carrier recovery. Details of the filter design and the channel simulator are described in [13]. For ideal carrier reference [11] (for both coherent and differentially coherent detection) the error rate is inversely proportional to E_b/N_0 and no irreducible error rate exists. However here all curves exhibit an irreducible error rate. The higher the alphabet size of the modulation scheme is, the better the performance of the coherent receiver is in comparison to differentially coherent detection. This is due to the fact that for high level modulation schemes the error rate is mainly determined by synchronization inaccuracies. Since the carrier recovery of the coherent receiver is superior, the coherent receiver clearly outperforms differential detection for large M.

In fig. 4 the influence of cochannel interference is discussed for the same channel conditions as in fig. 3. It shows the carrier-to-interference ratio required to achieve a bit error rate of 10^{-2} versus E_b/N_0. We assumed a frequency reuse cluster of 7 cells which corresponds to 6 cochannel interferers with equal interferer energy. Since cochannel interference appears to the Kalman filter to be white noise it is suppressed by the filter. For differential detection a similar effect is not present. Therefore coherent detection is more resistant to cochannel interference. E.g., to achieve $P_b = 10^{-2}$ for $E_b/N_0 \rightarrow \infty$ and for 4PSK, a CIR = 20 dB is needed in the coherent case and CIR = 27 dB for the differentially coherent receiver.

Clearly, the error rate of the coherent receiver calculated in section IV is a lower bound, since perfect elimination of the symbol phase for synchronization has been assumed whereas the results for differential detection are exact. If nonperfect elimination of the symbol phase and the influence of differential encoding of the information are taken into account the error rate is by about a factor of two larger than the lower bound. From fig. 3 it is clear that then the coherent receiver is no longer superior for a wide range of E_b/N_0. However the irreducible error rate is still smaller in the coherent case.

Things change when diversity techniques are used. The error rate then decreases approximately with the L-th power of of the signal-to-noise ratio, see fig. 5. The SNR loss of coherently detected MPSK due to the increase in the error rate by a factor of two because of differential encoding becomes increasingly smaller with increasing diversity degree L. Therefore coherent detection is again superior to differentially coherent detection. The above conclusions are also valid if channel coding together with interleaving is used instead of diversity. Diversity can be interpreted as just a special kind of coding.

VI. CONCLUSIONS

In this paper we compared coherent with differentially coherent detection of M-ary PSK on frequency-nonselective fading channels. The major advantage of differentially coherent detection is that its implementation is very simple and that no information about the channel is needed. However if the channel characteristics are known at least approximately and if diversity techniques or some kind of channel coding is employed, then coherent detection is superior, in particular when high level modulation schemes are used and in the presence of cochannel interference. Thus the major criterion for selecting an appropriate detection scheme is not the rapidity of the fading but rather the aspects mentioned above.

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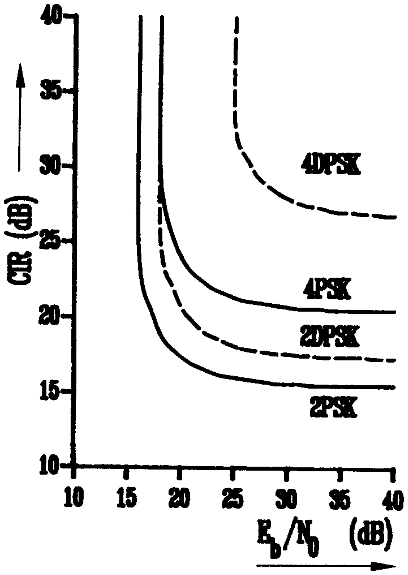

Fig.4: Required CIR to achieve a bit error rate of 10^{-2} versus E_b/N_0

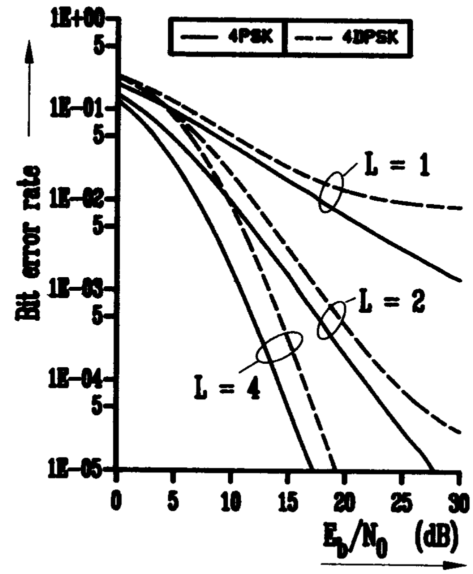

Fig.5: Bit error rate of 4PSK (coherent and differentially coherent) with diversity degree L as parameter (CIR → ∞)