where γ_c is the SNR/chip. The missing probability of the search mode P_M is the probability that all samples are $< R_K$. This probability is given by

$$P_M = \sum_{n=0}^{\theta/\Delta-1} (-1)^n \binom{\theta/\Delta - 1}{n} \times \left[\left(\frac{K}{n+K}\right)^{M/\Delta} - \left(\frac{K(1+\gamma_c W/M)}{1+\gamma_c+(n+K)(1+\gamma_c W/M)}\right)^{M/\Delta} \right] \quad (4)$$

The false alarm probability of the search mode can be obtained from P_M and P_D as $P_F = 1-P_M-P_D$. For the considered parallel acquisition scheme with the single dwell, the mean acquisition time $E[T_{acq}]$ can be modified by $(1+JP_M)MT_c/P_D$ [1]. J represents the penalty factor for returning a false alarm. In our analysis, the assumption that all samples of the I-Q MF are independent is approximately true when $k \ll M$ and the correlation coefficients are small.

Numerical results and discussions: In evaluating the performance of the proposed acquisition system, we used the following parameters: (i) the uncertainty region of 1024 chips is considered (not in a practical situation), (ii) MF length M of 128 chips is taken (or equivalently, $N = 8$), (iii) $\Delta = 1/2$, (iv) the penalty factor J is taken to be 1000, and (v) the correlation coefficients ρ_i for the fading process are taken to be ρ_i [1].

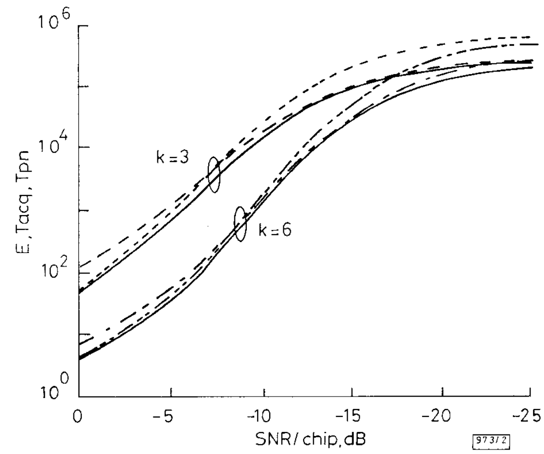

Fig. 2 *Mean acquisition time for proposed and convntional parallel acquisition schemes for the Rayleigh fading channel* ($\rho = 0.2$)

——— conventional parallel scheme (optimum)
- - - - - proposed parallel scheme ($k = 3$, $K = 16$)
— — proposed parallel scheme ($k = 3$, $K = 20$)
-·- - - proposed parallel scheme ($k = 6$, $K = 16$)
— - — proposed parallel scheme ($k = 6$, $K = 18$)

The performance comparison between the conventional parallel system described in [1] and the proposed parallel system is given for the Rayleigh fading channel. Fig. 2 shows the mean acquisition time for proposed and conventional parallel acquisition systems, where for the conventional parallel system, the threshold value for the search is selected numerically to minimise the mean acquisition time for each value of the SNR/chip. For the results shown, the mean acquisition time has been normalised by T_{pn}, where $T_{pn} = \theta T_c$. In the considered fading channel ($k = 3$ and $k = 6$), it is observed that: (i) The proposed acquisition system provides threshold value to give a suboptimal mean acquisition time performance about the theoretical optimum mean acquisition time of the conventional acquisition system; (ii) The mean acquisition time is less sensitivite to changes in K, and the mean acquisition time performance comparability to the conventional parallel system is maintained for a wide range of SNR/chip. This insensitivity is more significant when k becomes smaller, or equivalently, when the channel fades faster, which makes the proposed acquisition system a good candidate for slotted-mode preamble search in the CDMA reverse link under fast fading environments.

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Performance of component interleaved signal sets for fading channels

G. Taricco and E. Viterbo

Indexing terms: Fading, Rayleigh channels, Error statistics

The authors calculate the exact (pairwise) error probability for an arbitrary component-interleaved 2-D constellation over the Rayleigh fading channel. Using this result, they improve the accuracy of the performance analysis at high error rates. For illustration, this is applied to signal sets derived from 4-PSK and 16-QAM over the Rayleigh fading channel.

Introduction: The constant increase in the use of portable wireless communication systems has attracted considerable attention to digital transmission over fading channels. One of the problems with fading channels is that error probabilities decrease slowly compared to in the Gaussian channel case. A common solution is to resort to diversity techniques such as space, time, frequency, and code diversity [5].

Recently, a new type of diversity, known as modulation diversity, has been considered [1, 3]. It applies to any modulation scheme and affects its performance over fading channels in conjunction with component interleaving.

The modulation diversity of a signal set is defined as the minimum number of different components between any two distinct points of the set. As an example, Fig. 1 shows how rotation increases the diversity of 4-PSK from the figure shown in Fig. 1*a* to that in Fig. 1*b*.

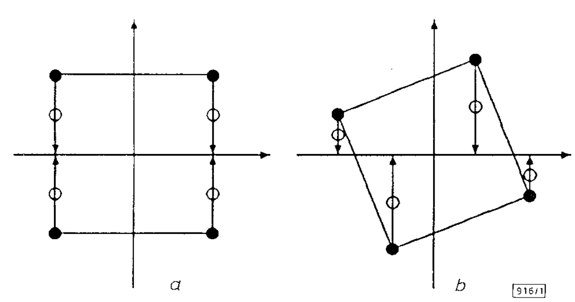

Fig. 1 *Example of modulation diversity*
a 4 PSK, $L = 1$
b 4 RPSK, $L = 2$

Performance enhancement on the fading channel can be explained as follows. Assume that a component interleaver is used before transmission. Then, in-phase and quadrature components of deinterleaved samples are affected by independent fading coefficients, which are supposed to be known at the receiver (perfect channel state information, CSI). If a deep fade affects one component of the transmitted signal, it is likely that the other one is not similarly impaired. It can easily be seen from Fig. 1 that the minimum distance of the 'compressed' constellation (empty circles) is lower in Fig. 1*a* than in Fig. 1*b*. In the case of a very deep fade, the points collapse together in Fig. 1*a*, but not in Fig. 1*b*. Optimally rotated 4-PSK achieves 8dB of gain in signal-to-noise ratio at error probability 10^{-3} against standard 4-PSK [3].

We study the performance of 2-D modulation over a Rayleigh fading channel.

System model: We consider a communication system with ideal (infinite-depth) component interleaving over a Rayleigh fading channel, unaffected by intersymbol interference or any other impairment than additive white Gaussian noise. Let $\mathbf{x} = (x_1, x_2)$ denote a transmitted signal sample from a given constellation S. Received signal samples are then given by

$$y_i = g_i x_i + n_i \qquad \text{for } i = 1, 2$$

Here, g_i are independent Rayleigh-distributed random variables with unit second moment (i.e. $E[g_i^2] = 1$) representing the fading coefficients, and n_i are Gaussian random variables with mean zero and variance $N_0/2$ representing the additive noise. We also denote $\mathbf{g} = (g_1, g_2)$, $\mathbf{n} = (n_1, n_2)$, and $\mathbf{y} = (y_1, y_2)$ and write $\mathbf{y} = \mathbf{g} \odot \mathbf{x} + \mathbf{n}$.

Perfect phase tracking and CSI is assumed at the receiver, which performs maximum likelihood (ML) detection. It computes the sample metrics

$$m(\mathbf{x}|\mathbf{y}, \mathbf{g}) = \|\mathbf{y} - \mathbf{g} \odot \mathbf{x}\|^2 \quad \forall \mathbf{x} \in S \qquad (1)$$

where $\|\cdot\|^2$ is the standard Euclidean norm, and decides for the signal $\hat{\mathbf{x}}$ attaining the minimum value of $m(\mathbf{x}|\mathbf{y}, \mathbf{g})$. [1 – 4] are based on the Chernoff bound to the PEP of a multidimensional signal set. They show that PEP is determined asymptotically by the modulation diversity L and the minimum L product distance $d^{(L)}_{p,min} = \Pi_{x_i \neq \hat{x}_i} |x_i - \hat{x}_i|$ of the constellation. Optimal signal sets maximise L and $d^{(L)}_{p,min}$.

Performance analysis: A standard approach of evaluating the symbol error probability of a signal set S consists in using the union bound

$$P(e) \leq \frac{1}{|S|} \sum_{\mathbf{x} \in S} \sum_{\mathbf{x} \neq \hat{\mathbf{x}}} P(\mathbf{x} \to \hat{\mathbf{x}}) \qquad (2)$$

Each PEP $P(\mathbf{x} \to \hat{\mathbf{x}})$ can be approximated by using the Chernoff bound or other bounds, but we show in this Letter that it can be calculated exactly over the Rayleigh fading channel with component interleaving.

Let $\mathbf{x} = (x_1, x_2)$, $\hat{\mathbf{x}} = (\hat{x}_1, \hat{x}_2)$, and $\mathbf{x} \to \hat{\mathbf{x}}$ represent a pairwise error event. According to eqn. 1, the PEP is given by

$$P(\mathbf{x} \to \hat{\mathbf{x}}) = P(\|n\|^2 > \|\mathbf{y} - \mathbf{g} \odot \hat{\mathbf{x}}\|^2) = E_{\mathbf{g}}\left[Q\left(\sqrt{\frac{g_1^2\Delta_1^2 + g_2^2\Delta_2^2}{2N_0}}\right)\right] \qquad (3)$$

where $\Delta_i = x_i - \hat{x}_i$ for $i = 1, 2$ and $Q(x) = (1/\sqrt{[2\pi]})\int_x^\infty e^{-u^2/2}\, du$ is the Gaussian tail function. The expectation in eqn. 3 can be evaluated as follows:

$$P(\mathbf{x} \to \hat{\mathbf{x}}) = \int_0^\infty \int_0^\infty \int_{\sqrt{(g_1^2\Delta_1^2+g_2^2\Delta_2^2)/2N_0}}^\infty 2g_1 e^{-g_1^2} 2g_2 e^{-g_2^2} \frac{1}{\sqrt{2\pi}} e^{-z^2/2} dz\, dg_1\, dg_2 \qquad (4)$$

Substituting $g_1 = \rho\cos\phi$, $g_2 = \rho\sin\phi$, and $z = \rho u$, we obtain

$$P(\mathbf{x} \to \hat{\mathbf{x}}) = \int_0^\infty \int_0^{\pi/2} \int_{\sqrt{(\Delta_1^2\cos^2\phi+\Delta_2^2\sin^2\phi)/2N_0}}^\infty \frac{4}{\sqrt{2\pi}} \rho^4 \cos\phi \sin\phi\, e^{-\rho^2(1+u^2/2)} du\, d\rho\, d\phi$$

$$= \int_0^{\pi/2} \int_{\sqrt{(\Delta_1^2\cos^2\phi+\Delta_2^2\sin^2\phi)/2N_0}}^\infty \frac{6}{(2+u^2)^{5/2}} \cos\phi \sin\phi\, du\, d\phi$$

$$= \int_0^1 \int_{\sqrt{(\Delta_1^2(1-v)+\Delta_2^2 v)/2N_0}}^\infty \frac{3}{(2+u^2)^{5/2}} du\, dv$$

$$= \frac{1}{2}\left[1 - \frac{1}{\Delta_1^2 - \Delta_2^2}\left(\frac{\Delta_1^3}{\sqrt{4N_0 + \Delta_1^2}} - \frac{\Delta_2^3}{\sqrt{4N_0 + \Delta_2^2}}\right)\right] \qquad (5)$$

where: (i) we integrated over ρ, (ii) substituted $v = \sin^2\phi$, and (iii) integrated to obtain the final result. Series expansions of the PEP are given by

$$P(\mathbf{x} \to \hat{\mathbf{x}}) = \begin{cases} \frac{3}{\Delta_1^2\Delta_2^2} N_0^2 + O(N_0^3) & N_0 \to 0 \\ \frac{1}{2} - \frac{1}{4}\frac{\Delta_1^3-\Delta_2^3}{\Delta_1^2-\Delta_2^2} N_0^{-1/2} + O(N_0^{-3/2}) & N_0 \to \infty \end{cases} \qquad (6)$$

Note that the first expression, valid for high SNR, is a significant improvement on the corresponding Chernoff bound [1]

$$P(\mathbf{x} \to \hat{\mathbf{x}}) \leq \frac{1}{(1 + \Delta_1^2/4N_0)(1 + \Delta_2^2/4N_0)} = \frac{16}{\Delta_1^2\Delta_2^2} N_0^2 + O(N_0^3) \qquad (7)$$

which is asymptotically looser by $10\log_{10} \sqrt{(16/3)} = 3.6$dB.

Results: In this section we consider the error performance of rotated 4-PSK and 16-QAM (4-RPSK and 16-RQAM, respectively). Results are shown as symbol error rate curves against E_s/N_0 in dB, where E_s is the average signal energy. The rotation angle θ is chosen to maximise the minimum product distance of the rotated constellation. Straightforward analysis gives

$$\frac{d^{(2)}_{p,min}}{E_s} = \begin{cases} \min\{2|c|, |s|\} & \text{4-RPSK} \\ \min\{0.4|c|, 0.2|s|, |0.8c \pm 0.6s|, \\ |1.2c \pm 1.6s|, |2.4c \pm s|\} & \text{16-RQAM} \end{cases}$$

where we set $c = \cos(2\theta)$ and $s = \sin(2\theta)$. The optimal rotation angle is $\theta = {}^1/_2 \arctan 2 = 31.7°$, in both cases, corresponding to $d^{(2)}_{p,min} = 0.894E_s$ (4-RPSK) and $0.179E_s$ (16-RQAM). Asymptotic performance within 2dB from the optimum is obtained with θ in the range from 17.2 to 36.8° (for 4-RPSK) and from 29.8 to 32.4° (for 16-RQAM).

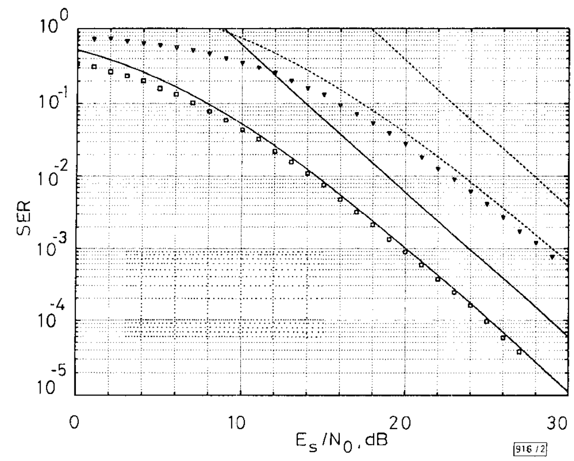

Fig. 2 *SER of 4 RPSK and 16 RQAM with optimal rotation angle θ = 31.7°*

——— analytic bounds for 4 RPSK
- - - - analytic bounds for 16 RQAM
□ 4 RPSK, simulation
▽ 16 RQAM, simulation
Upper curves are based on Chernoff bounds, lower curves on exact PEP

Fig. 2 shows the symbol error rate (SER) of 4-RPSK and 16-RQAM with optimal rotation angle θ = 31.7°. The curves report simulation and analytic results based on the Chernoff bound and the exact PEP. They show that the union bound based on exact

PEP is within 0.2 and 0.7dB from simulation results for 4-RPSK and 16-RQAM, respectively, and the same bound based on Chernoff bounds on PEP is ~4dB looser.

The accuracy of the proposed bound suggests it can be used as a design criterion for the choice of 2-D constellations for use on the Rayleigh fading channel with component interleaving.

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Resynchronising delay-locked loop

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Indexing terms: Spread spectrum communication, Synchronisation

The delay-locked loop (DLL) is a device that is often used for PN-code tracking to synchronise direct sequence spread spectrum (DS-SS) systems. A DLL with resynchronising capability, which is more robust against loss of lock, is presented. The new scheme has two modes: a normal tracking mode and a resynchronising mode. The structure of the new loop and its function are described.

Introduction: The delay-locked loop (DLL) [1] generates a local reference PN-sequence in the receiver which is time-aligned to the received direct sequence spread spectrum (DS-SS) signal. This is achieved by correlating the received signal with a local PN-sequence in order to estimate the delay between those two signals. The delay detector output controls the reference signal generator in a closed loop operation in order to minimise the delay error in the presence of noise and Doppler effects. The delay error detector characteristic is called the S-curve $S(\varepsilon)$, where ε is the normalised delay error. The shape of the S-curve is very important for the DLL performance. The range where the error detector provides a useful control signal is known as the overall tracking range (TR_{ov}) of the DLL. Other important parameters of the DLL are the loop order and bandwidth, the number of correlators used and the operation mode, i.e. coherent or noncoherent. Important performance criteria are the RMS tracking jitter and the mean time to lose lock (MTLL). A coherent DLL configuration for non data-modulated signals disturbed by white Gaussian noise is considered here.

Resynchronising DLL: The DLL is designed to function at specified operating conditions. These operating conditions include the noise level and the input dynamics of the received signal. The DLL operates well under these specified conditions. However, if higher noise levels or higher input dynamics occur, then the DLL will easily lose lock and a time consuming reacquisition process has to be started. During reacquisition no data demodulation or ranging is possible because there is no ontime reference signal available. The DLL can be made more robust against loss of lock if the DLL is adapted to the operating conditions [2].

A loss of lock occurs when the DLL leaves TR_{ov} because the error detector produces a zero control signal outside the TR_{ov}. The use of extended tracking range DLLs [1] extends TR_{ov}, depending on the number of correlators used. However, each additional correlator increases the noise level in the loop: this limits the number of correlators that can be used in practical applications. This can be improved by selecting only the correlators with a useful control signal, as has been proposed in [3]. Ideally this will extend TR_{ov} while maintaining the noise level of the conventional DLL. Owing to decision errors in the selection device the actual performance is not ideal.

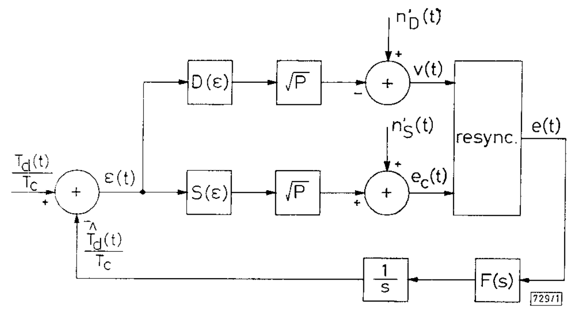

Fig. 1 *Baseband equivalent model*

A solution for keeping the DLL locked is to detect an imminent loss of lock and to use an extra control signal to bring the DLL back into the overall tracking range. This is known as a resynchronising DLL, and its function is proposed in this Letter. The new DLL structure is a classical DLL enhanced by a resynchronising unit. The baseband equivalent model is shown in Fig. 1. During normal operation in the vicinity of the operating point there is no difference between this DLL and the conventional DLL. When the DLL leaves TR_{ov}, then a loss of lock is detected and the DLL is switched into the resynchronising mode. The detection of an imminent loss of lock has to be determined with high probability and in enough time. The detection is based on a conventional lock detector [1], eventually with enlarged lock detector characteristic $D(\varepsilon)$. The threshold on the detector output should be optimised to keep the false alarm probability low. The filtering of the detector output is very important and must be smooth enough.

When an imminent loss of lock has been detected, the algorithm has to determine on which side of the error detector characteristic the actual delay error is. This is crucial to generate the correct resynchronising control signal. The correct side can be determined by considering the error detector output some time before the loss of lock occurs. If the DLL was in a late condition before the lock of loss a positive static control signal is applied to speed the loop up. If it was in an early condition a negative static control signal is applied to slow the loop down. A filtering of the error detector output is used to consider the past signal characteristic. The resynchronising control signal is added, with the correct sign to the conventional DLL control signal. The amplitude of the resynchronising control signal must be chosen carefully in order not to oversteer the DLL. The use of an extended DLL is useful because the DLL has a longer interval with the same error detector signal sign before losing lock.

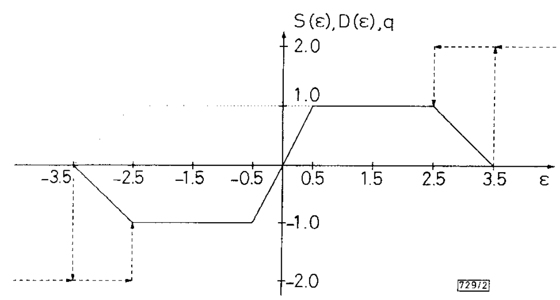

Fig. 2 *Detector curves for resynchronising DLL*
—— error detetor $S(\varepsilon)$, – – – resync signal q, ······ lock detector $D(\varepsilon)$