Initial Synchronization of W-CDMA Systems using a Power-Scaled Detector with Antenna Diversity in Frequency-Selective Rayleigh Fading Channels

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***Abstract*— An analytical evaluation of the performance in terms of detection probability and mean detection time of a noncoherent detector at the base station is presented for a wideband CDMA system, where the mobile terminal transmits a pilot signal in the form of bursts of modulated chips, which are transmitted periodically and separated by long silent intervals. The performance analysis is carried out for a power-scaled detector employing temporal and spatial noncoherent averaging in frequency-selective fading channels taking channel dynamics and initial frequency offsets into account. With the presented analysis, which is verified by means of simulations, it is possible to conveniently trade off temporal noncoherent versus coherent averaging depending on the length of the pilot bursts and channel dynamics and to examine the improvement of using multiple diversity antennas at the base station.**

I. Introduction

In common DS-CDMA systems like UMTS [1], random access in the uplink is realized by means of a periodically transmitted pilot preamble. In particular, the terminal transmits bursts of known pilot chips separated by a long silent interval. After the pilot preamble has been detected by the base station, an acquisition indicator is sent back to the terminal and a dedicated communication link is established. The received preamble samples are also used for an initial timing acquisition of the channel taps. This initial timing information is passed on to the succeeding Rake structure [3] used for processing the data symbols of the dedicated communication link and is further refined using timing tracking structures as described in [2].

Usually, the detection of a known signal in a noisy environment is performed by means of binary hypothesis tests. Therefore, with a spacing of one chip period or a fraction of it, a certain region of the delay domain, the search window, is searched for the pilot signal. This search can either be done serially [4] or in parallel [5].

The binary tests are usually performed by correlating the received signal with the known pilot sequence and comparing the squared magnitude of the correlator output with a fixed threshold, see e.g. [6].

An important aspect is the impact of channel dynamics and an initial frequency offset. Due to the time-varying nature of the channel, not the whole pilot sequence may be used for coherent accumulation without undergoing degradation in performance. Hence, in order to exploit the temporal diversity of the fading channel, successive correlator outputs have to be accumulated noncoherently, as also proposed in [6].

The detection performance can be further improved by exploiting spatial diversity through the use of multiple antenna elements at the base station receiver. Before initial acquisition occurs, parameters like the amplitudes and phases of the fading processes are not available and cannot be used for maximum ratio combining. Hence, the antenna elements are noncoherently combined as proposed in [7].

Since the received interference level is unknown, some form of adaptive threshold setting is required to guarantee a constant false alarm rate (CFAR). Equivalently, the test statistic may be scaled by an estimate of the total received power as proposed in [8].

Thus, by employing the concepts of temporal and spatial noncoherent averaging and scaling the test statistic by the estimated total received power, the basic decision device used to scan the search window for the pilot signal is shown in Fig. 1.

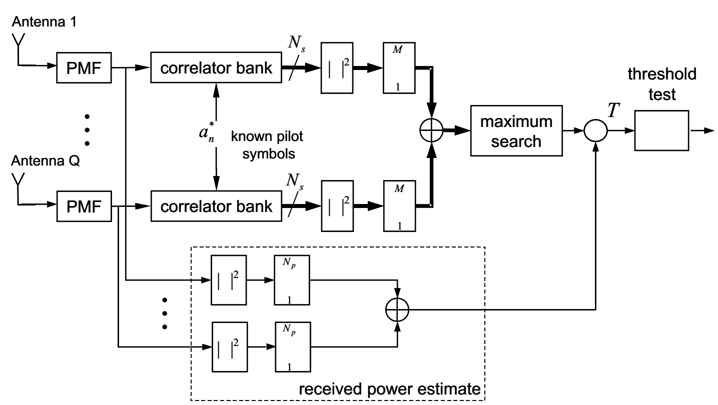

Fig. 1. Block diagram of power-scaled decision device employing temporal and spatial noncoherent averaging.

The focus of this paper is to analytically assess the performance of the decision device in Fig. 1 in terms of detection probability and mean detection time taking both temporal and spatial averaging and scaling of the test statistic by the estimated total received power into account. The analysis is carried out for frequency selective Rayleigh fading channels. In previous publications the detection performance is assessed either not taking noncoherent averaging into account [5], taking only spatial noncoherent averaging

into account [7], [9], or taking only temporal noncoherent averaging based on a flat block fading channel assumption into account [6].
Based on the results of the analysis the performance improvement using multiple antenna elements is examined and the trade-off between coherent and noncoherent temporal averaging can be carried out easily.

II. Received Signal Model

By employing Q antenna elements at the receiver, it is assumed that, due to the small size of the antenna array, the long-term channel statistics are identical at each antenna element, but that the spatial separation of the antenna elements is sufficiently large to yield mutually uncorrelated signals.
Like in [5] the multipath fading channel is modelled as a tapped delay line with a tap spacing of one chip period T_c, see Fig. 2. Each tap is multiplied by an independent time-varying complex zero-mean Gaussian random variable. Due to the long silent interval it is assumed that the whole delay region of uncertainty is searched for a signal between two successive pilot bursts. Hence, without loss of generality, it is assumed that there are L channel taps with non-zero variance σ_l^2, $l = 0, \dots, L-1$ located at a delay $\tau_l = lT_c$. In addition, the total channel power is normalized to unity.
In common DS-CDMA systems, like in UMTS [1], the possible starting points of the pilot bursts are known to the base station and hence, the total delay region of uncertainty can be constrained to N_s chips. The size N_s of the search window depends on the physical cell radii, which are usually in the range up to several kilometers.

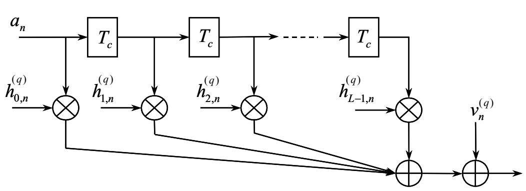

Fig. 2. Effective frequency selective fading channel model.

Assuming like in [5], that the receiver is chip synchronized with the user of interest, the T_c-sampled version of the received signal after chip-matched filtering at the q-th antenna element can be written as

$$z_n^{(q)} = z^{(q)}(t = nT_c) = e^{j2\pi f_e nT_c + \phi} \sum_{l=0}^{L-1} h_{l,n}^{(q)} a_{n-l} + v_n^{(q)}, \tag{1}$$

for $n = 0, \dots, N_p - 1$. The transmit and receive pulses are modelled within the effective channel coefficients $h_{l,n}^{(q)}$. The additive white Gaussian noise at each antenna element, which also models other user interference, is denoted by $v_n^{(q)}$ with variance σ_v^2, and a_n denotes the complex chips of a pilot burst of length N_p chips consisting of a periodically repeated signature scrambled by a long scrambling code like in [1]. A possible initial frequency and phase offset is denoted by f_e and ϕ, respectively.

III. Performance Analysis

A. Test Statistic

The correlator bank in Fig. 1 consists of N_s sequence correlators, which partition the pilot sequence into $M = N_p/N_c$ nonoverlapping consecutive subsequences of length N_c and therefore, perform for each subsequence a coherent accumulation over N_c chips by correlating the received signal with the known pilot sequence corresponding to each tap within the search window. However, as mentioned above, due to the long silent interval between two successive pilot bursts it is not necessary to perform a completely parallel processing. The test statistic T (see Fig. 1) can be written as

$$T = \max_{d \in \{0,\dots,N_s-1\}} \frac{\sum_{q=0}^{Q-1} \sum_{m=0}^{M-1} \left| \boldsymbol{a}_d^{(m)^H} \boldsymbol{z}^{(q,m)} \right|^2}{\sum_{q=0}^{Q-1} \sum_{n=0}^{MN_c-1} \left| z_n^{(q)} \right|^2}, \tag{2}$$

where $\boldsymbol{z}^{(q,m)} = [z_{mNc}^{(q)}, \dots, z_{(m+1)Nc-1}^{(q)}]^T$ denotes the m-th received signal block at the q-th antenna element. The m-th block of the pilot sequence used for coherent accumulation with respect to a delay of d chips is denoted by $\boldsymbol{a}_d^{(m)} = [a_{mN_c-d}, \dots, a_{(m+1)N_c-1-d}]^T$ setting $a_n = 0$ for $n < 0$ or $n > N_p - 1$. By defining the correlator output of the m-th signal block at antenna q corresponding to a delay d

$$x_d^{(q,m)} = \boldsymbol{a}_d^{(m)^H} \boldsymbol{z}^{(q,m)}, \tag{3}$$

the numerator of the test statistic can be written as the noncoherent accumulation of the squared magnitude of the correlator outputs

$$T = \max_d \frac{\sum_{q=0}^{Q-1} \sum_{m=0}^{M-1} \left| x_d^{(q,m)} \right|^2}{\sum_{q=0}^{Q-1} \sum_{n=0}^{MN_c-1} \left| z_n^{(q)} \right|^2} = \max_d T_d. \tag{4}$$

B. Detection Probability

In this section an expression for the probability, that the test statistic T exceeds a certain threshold in case a pilot burst has been transmitted, is derived.
Let the temporal consecutive correlator outputs at the q-th antenna element be contained in the vector

$$\boldsymbol{x}_d^{(q)} = [x_d^{(q,0)}, \dots, x_d^{(q,M-1)}]^T. \tag{5}$$

Since the received signal is zero-mean complex Gaussian distributed, the correlator outputs $x_d^{(q,m)}$ are also zero-mean jointly Gaussian distributed and hence, $\boldsymbol{x}_d^{(q)}$ is completely characterized by its covariance matrix

$$\breve{\boldsymbol{K}}_d = \mathrm{E}\left\{ \boldsymbol{x}_d^{(q)} \boldsymbol{x}_d^{(q)^H} \right\}, \tag{6}$$

which is independent of the respective antenna element, since by assumption, the temporal long-term statistics at each antenna element are the same.
By assuming good autocorrelation properties of the pilot scrambling sequence and using the autocorrelation properties of the independent Rayleigh fading channel taps, see [10],

$$\mathrm{E}\{h_{l,k}^{(q)} h_{l,n}^{(q)^*}\} = \sigma_l^2 J_0(2\pi f_D [k-n] T_c) \tag{7}$$

the elements of the covariance matrix can be easily calculated as follows

$$\breve{\boldsymbol{K}}_d(k,n) = \mathrm{E}\{x_d^{(q,k)} x_d^{(q,n)^*}\} = \begin{cases} N_c\sigma_d^2 e^{j2\pi f_e N_c T_c(k-n)} \sum_{u=-N_c+1}^{N_c-1}\left(1-\frac{|u|}{N_c}\right) \\ \times J_0(2\pi f_D[N_c(k-n)+u]T_c)\cos(2\pi f_e u T_c) \quad 0\le d\le L-1 \\ +N_c\sigma_v^2\delta(k-n) \\ N_c\sigma_v^2\delta(k-n) \qquad L\le d\le N_s-1 \end{cases} \tag{8}$$

where J_0 denotes the zero-order Bessel function of the first kind and $\delta(n)$ is the Kronecker-delta. The maximum Doppler frequency is denoted by f_D which is related to the terminal velocity v and the carrier frequency f_t via $f_D = \frac{v}{c} f_t$, where c is the speed of light.

Now, let all correlator outputs corresponding to a delay d be comprised in the vector

$$\boldsymbol{x}_d = [\boldsymbol{x}_d^{(0)^T}, \ldots, \boldsymbol{x}_d^{(Q-1)^T}]^T. \tag{9}$$

Per assumption, the signals at the antenna elements are mutually uncorrelated and hence, the $(QM \times QM)$ covariance matrix of $\boldsymbol{x}_d$ is block diagonal

$$\boldsymbol{K}_d = \mathrm{E}\{\boldsymbol{x}_d \boldsymbol{x}_d^H\} = \boldsymbol{I}_Q \otimes \breve{\boldsymbol{K}}_d \tag{10}$$

where $\otimes$ denotes the Kronecker product and $\boldsymbol{I}_Q$ is the $(Q\times Q)$ identity matrix.

Thus, for $d = 0, \ldots, L-1$, the numerator of the test statistic T_d in (4) is the sum of the squared magnitude of MQ correlated Gaussian random variables, whereas for $d = L, \ldots, N_s - 1$ the $x_d^{(q,m)}$ are independent. But also for $d = 0, \ldots, L-1$ the numerator can be written as the sum of independent random variables by means of an eigendecomposition of the hermitian positive semidefinite matrix $\breve{\boldsymbol{K}}_d$. According to the block diagonal structure $\boldsymbol{K}_d$ can be written as

$$\boldsymbol{K}_d = \boldsymbol{I}_Q \otimes \boldsymbol{V}_d \boldsymbol{\Lambda}_d \boldsymbol{V}_d^H, \tag{11}$$

where $\boldsymbol{V}_d$ is an unitary matrix and $\boldsymbol{\Lambda}_d$ is a diagonal matrix with diagonal elements $\lambda_d^{(0)}, \ldots, \lambda_d^{(M-1)}$ denoting the M eigenvalues of $\breve{\boldsymbol{K}}_d$. By using $\boldsymbol{V}_d\boldsymbol{V}_d^H = \boldsymbol{I}$, the numerator of T_d in (4) can be equivalently written as $\sum_{q=0}^{Q-1} \boldsymbol{x}_d^{(q)^H}\boldsymbol{x}_d^{(q)} = \sum_{q=0}^{Q-1} \boldsymbol{x}_d^{(q)^H}\boldsymbol{V}_d\boldsymbol{V}_d^H\boldsymbol{x}_d^{(q)} = \sum_{q=0}^{Q-1}\tilde{\boldsymbol{x}}_d^{(q)^H}\tilde{\boldsymbol{x}}_d^{(q)}$, where $\tilde{\boldsymbol{x}}_d^{(q)} = \boldsymbol{V}_d^H\boldsymbol{x}_d^{(q)}$. Since $\mathrm{E}\{\tilde{\boldsymbol{x}}_d^{(q)}\tilde{\boldsymbol{x}}_d^{(q)^H}\} = \mathrm{E}\{\boldsymbol{V}_d^H\boldsymbol{x}_d^{(q)}\boldsymbol{x}_d^{(q)^H}\boldsymbol{V}_d\} = \boldsymbol{V}_d^H\breve{\boldsymbol{K}}_d\boldsymbol{V}_d = \boldsymbol{\Lambda}_d$ is a diagonal matrix, the numerator of T_d is the sum of the squared magnitude of independent complex zero-mean Gaussian random variables with Q-fold identical variance $\lambda_d^{(0)}, \ldots, \lambda_d^{(M-1)}$, i.e the sum of M nonidentically distributed chi-square random variables, each having $2Q$ degrees of freedom.

The denominator in (4) is also a sum of chi-square distributed random variables, but due to the large number of degrees of freedom ($2QMN_c$) and since the chip signal-to-noise ratio $1/\sigma_v^2$ is very small it can be very well approximated by

$$\sum_{q=0}^{Q-1}\sum_{n=0}^{MN_c-1}\left|z_n^{(q)}\right|^2 \approx QMN_c(1+\sigma_v^2) \approx QMN_c\sigma_v^2. \tag{12}$$

Now, let the cumulative density function (cdf) of T_d in case a signal is present (hypothesis $\mathcal{H}_1$) be denoted by $F_{T_d|\mathcal{H}_1}(\gamma)$. A convenient way to calculate the cdf of T_d is by means of the characteristic function of T_d. Since T_d can be expressed as the sum of the squared magnitude of QM uncorrelated complex Gaussian random variables with different variances

$$\widetilde{\lambda}_d^{(m)} = \frac{\lambda_d^{(m)}}{QMN_c\sigma_v^2} \quad ,m = 0, \ldots, M-1 \tag{13}$$

the characteristic function of T_d is given by

$$\Phi_{T_d|\mathcal{H}_1}(\omega) = \prod_{m=0}^{M-1}\frac{1}{(1-j\omega\widetilde{\lambda}_d^{(m)})^Q}. \tag{14}$$

Now, a common method is to perform a partial fraction expansion of the characteristic function resulting in closed form expression of the pdf and the cdf by inverse Fourier transform of the partial fractions. A problem with this approach is the numerical evaluation of the resulting closed-form expressions, especially if the eigenvalues in (13) are nearly identical. For $Q > 1$ there are indeed Q-fold eigenvalues and it is difficult to analytically perform the partial fraction expansion, if in addition it is $M > 2$.

Hence, a numerically more stable way is to calculate the cdf of T_d directly via the characteristic function. According to the lemma of Gil-Pelaez (see e.g. [11]), the cdf $F_X(x)$ of a random variable X can be calculated via the characteristic function $\Phi_X(\omega)$ of X as follows

$$F_X(x) = \frac{1}{2} - \frac{1}{\pi}\int_0^\infty \frac{\mathrm{Im}\left\{\Phi_X(\omega)e^{-jx\omega}\right\}}{\omega}\, d\omega. \tag{15}$$

Therefore, the cdf of T_d can be expressed in terms of a single integral

$$\begin{aligned} F_{T_d|\mathcal{H}_1}(\gamma) &= \frac{1}{2} - \frac{1}{\pi}\int_0^\infty \frac{\mathrm{Im}\left\{\Phi_{T_d|\mathcal{H}_1}(\omega)e^{-j\gamma\omega}\right\}}{\omega}\, d\omega \\ &= \frac{1}{2} - \frac{1}{\pi}\int_0^\infty \mathrm{Im}\left\{\frac{e^{-j\gamma\omega}}{\omega\prod_{m=0}^{M-1}(1-j\omega\widetilde{\lambda}_d^{(m)})^Q}\right\} d\omega, \end{aligned} \tag{16}$$

which can be evaluated numerically. Note, that the integrand is real valued and decays very fast for increasing ω. The limit of the integrand for $\omega \to 0$ exists and can easily be calculated

$$\lim_{\omega\to 0} \; \mathrm{Im}\left\{\frac{e^{-j\gamma\omega}}{\omega\prod_{m=0}^{M-1}(1-j\omega\widetilde{\lambda}_d^{(m)})^Q}\right\} = -\gamma + Q\sum_{m=0}^{M-1}\widetilde{\lambda}_d^{(m)}. \tag{17}$$

With (16), the probability of detecting the pilot signal within the search window, and hence, achieving initial timing synchronization, can be calculated as

$$\begin{aligned} P_D &= P\left\{\max_{d\in\{0,\ldots,L-1\}} T_d > \gamma|\mathcal{H}_1\right\} &(18) \\ &= 1 - P\left\{\bigcap_{d=0}^{L-1} T_d \le \gamma|\mathcal{H}_1\right\} &(19) \\ &= 1 - \prod_{d=0}^{L-1} F_{T_d|\mathcal{H}_1}(\gamma), &(20) \end{aligned}$$

where for the last equality the fact, that the events $T_d \le \gamma$ are independent, has been used.

Note, that the derivation can be easily extended to the case of spatial fading correlations, by replacing $\boldsymbol{I}_Q$ in (11) with a spatial correlation matrix and performing the eigendecomposition on $\boldsymbol{K}_d$.

C. False Alarm Probability

In case no signal is present (hypothesis $\mathcal{H}_0$), a false alarm event is very costly, since it results in the initiation of further signalling and processing procedures at the base station receiver. Hence, only a certain level of false alarm should be allowed.

Obviously, in case no signal is present, all correlator outputs $x_d^{(q,m)}$ are independent identical complex Gaussian distributed with zero-mean and variance $N_c\sigma_v^2$. Approximating the estimated received power by a constant, like in (12), the test statistic T_d can be expressed as the sum of the squared magnitude of QM uncorrelated complex identical distributed Gaussian random variables with variance $\frac{1}{QM}$, or equivalently, QMT_d is chi-square distributed with $2QM$ degrees of freedom.

Using the result for the right-tail probability of a chi-square distributed random variable with an even number of degrees of freedom (see e.g. [12]), the probability, that T_d exceeds a threshold γ in case no signal is present, is for all $d = 0, \ldots, N_s - 1$ given by

$$P\{T_d > \gamma|\mathcal{H}_0\} = e^{-QM\gamma} \sum_{k=0}^{QM-1} \frac{(QM\gamma)^k}{k!}. \qquad (21)$$

The total probability of false alarm taking the search window of length N_s chips into account, is given by

$$P_F = P\left\{ \max_{d\in\{0,\ldots,N_s-1\}} T_d > \gamma|\mathcal{H}_0 \right\} \qquad (22)$$

$$= 1 - \left[1 - e^{-QM\gamma} \sum_{k=0}^{QM-1} \frac{(QM\gamma)^k}{k!} \right]^{N_s} \qquad (23)$$

In order to calculate the value of the threshold corresponding to a specific false alarm probability, (23) has to be numerically solved for γ. Note, that the false alarm probability does not depend on the amount of interference σ_v^2, since the test statistic is scaled by the estimated total received power.

D. Mean Detection Time

Let the mean detection time $\overline{T}_D$ be defined as the average time it takes to achieve initial synchronization in terms of detecting a random access event. Usually, like in [1] the pilot bursts are repeated with increasing power. Let ρ_0 be the chip-to-interference ratio of the initial pilot burst, which is incremented by $\Delta\rho$ after each transmission. Furthermore, let T_p denote the time between the starting points of two consecutively transmitted bursts and $T_{D_i} = iT_p$ the time it takes that the pilot sequence is detected after i successive transmissions, then the mean detection time can be calculated as follows

$$\begin{aligned}\overline{T}_D &= \mathrm{E}\{T_{D_i}\} \\ &= \sum_{i=1}^{\infty} iT_p \ \Pr\{\text{preamble first detected at } i\text{-th transmission}\} \\ &= \sum_{i=1}^{\infty} iT_p P_D(\rho_0 + (i–1)\Delta\rho) \prod_{k=1}^{i-1} (1–P_D(\rho_0 + (i–1)\Delta\rho)),\end{aligned} \qquad (24)$$

where $P_D(\rho)$ is given by (20) corresponding to a chip-to-interference ratio of $\rho = 1/\sigma_v^2$. Of course, for the evaluation of $\overline{T}_D$, the sum does not have to be taken up to ∞, since the probability that a preamble has not been detected until the i-th retransmission vanishes very rapidly with increasing i.

IV. Results

First of all, the theoretical results are verified by means of Monte-Carlo simulations[1]. The setup for the simulation of the DS-CDMA system is as follows. According to the 3GPP standard [13], the QPSK modulated pilot preamble consists of a 4096 chips long fraction of a long scrambling Gold sequence, i.e. $N_p = 4096$ chips, and is transmitted through a selective Rayleigh fading channel after being pulse-shaped by a root-raised cosine transmit filter with roll-off factor 0.22. The channel consists of 4 taps located at delays of $\{0, 1, 2, 3\}$ chips in addition to an initial delay of 10 chips. The average tap powers are $\{0, -3, -6, -9\}$ dB, according to the *Case 3* channel from [14]. The total channel power is normalized to 1 and the channel dynamic corresponds to a terminal velocity of $v = 500$ km/h. The search window is set to $N_s = 64$ chips, corresponding to a cell radius of 2.5 km.

For a partitioning of the received sequence into $M = 1$ and $M = 2$ blocks ($N_c = 4096$ and $N_c = 2048$) the detection miss probability ($P_M = 1 - P_D$), with P_D from (20), is plotted versus the chip signal-to-interference ratio (SIR) in Fig. 3, whereas the false alarm probability is set to $P_F = 1\%$. Obviously, the theoretical results match the simulative results very well. It can be observed, that using the whole pilot preamble for coherent accumulation yields a degradation in performance, as expected for such high channel dynamics. Adding spatial diversity to the system by using $Q = 2$ instead of $Q = 1$ antenna element at the receiver results in a significant performance gain (Fig. 3b).

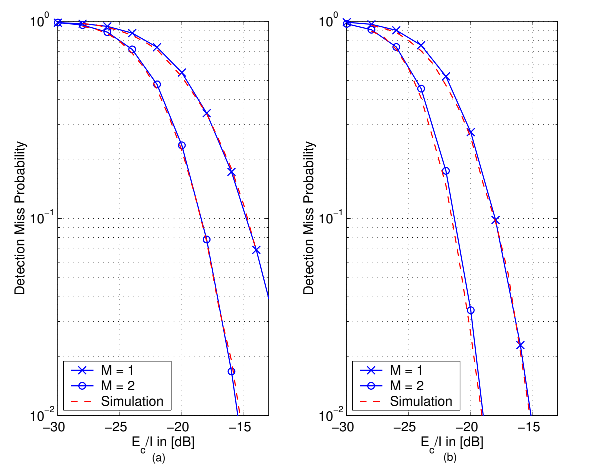

Fig. 3. Detection miss probability versus E_c/I for a frequency selective fading channel, $P_F = 0.01$, $N_s = 64$ chips, $N_p = 4096$ chips employing (a) one Antenna (b) two Antennas at the receiver.

In order to illustrate the effect of channel dynamics and frequency error on the optimal partitioning of the pilot sequence, Fig. 4 depicts the optimal partitioning of the pilot preamble into M blocks of length $N_c = \lfloor N_p/M \rfloor$ chips for a flat fading channel. As expected, large frequency errors and high channel dynamics restrict the correlator length. Obviously, the optimal

[1]The results were obtained using CoCentric System Studio from Synopsys.

partitioning also strongly depends on the operating point in terms of chip SIR. For a high chip SIR, the optimum correlator length is expected to be smaller than for a low chip SIR, since the priority shifts from further increasing the chip SIR by means of coherent averaging to exploiting the temporal diversity by means of noncoherent averaging.

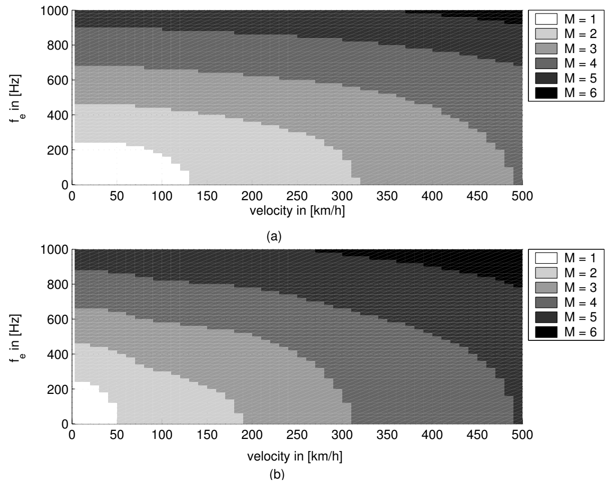

Fig. 4. (a) Optimal partitioning of preamble sequence (a) $E_c/I = -23$ dB (b) $E_c/I = -14$ dB, $P_F = 0.01$, $N_s = 64$ chips, $N_p = 4096$ chips.

For an initial access scheme as specified in [1] with periodically power-ramped pilot bursts, an appropriate performance measure is the mean time it takes to acquire initial synchronization, rather than just the performance of detecting one pilot burst. Hence, by evaluating (24), the mean detection time is plotted in Fig. 5 versus the partitioning of the pilot sequence for various scenarios. Usually, in order to minimize interference, the initial chip SIR of the first pilot burst is very low. Here, it is assumed, that the first pilot burst has a chip SIR of $\rho_0 = -30$ dB and that the power increment between two successive pilot bursts is $\Delta\rho = 1$ dB. According to [1] the time between the starting points of two consecutively transmitted bursts is $T_p = 15360$ chips, whereas the length of one pilot burst is $N_p = 4096$ chips, as above. Obviously, for low and moderate channel dynamics it is optimal to use the whole pilot burst for coherent accumulation. For high channel dynamics, the pilot sequence should be split up into 3 parts which are noncoherently accumulated.

As expected, exploiting spatial diversity through the use of two antenna elements, the mean detection time can be lowered significantly. But interestingly, the presence of multipath diversity (Case 3 Channel) yields a degradation of performance. This has two reasons. First, the total channel power is normalized to 1, and hence, an additional path always results in decreased absolute powers of the remaining channel paths, and in addition, the chip SIR is very low. A multipath diversity gain can only be achieved for a larger chip SIR, but the mean acquisition time is lower than 30 ms, i.e. the chip SIR is in the average lower than -23 dB at the point of detection, for all scenarios presented in Fig. 5.

V. Conclusion

An analytical framework for the performance in terms of detection probability and mean detection time of a power-scaled detector employing temporal and spatial noncoherent averaging has been derived for the uplink of a W-CDMA system, where the initial synchronization is facilitated by the use of a periodically repeated pilot preamble. The performance analysis has been carried out for frequency selective fading channels taking an initial frequency offset and channel dynamics into account. It has been shown, that the mean detection time can be significantly lowered by the use of multiple diversity antennas and an optimal partitioning of the pilot preamble at the receiver.

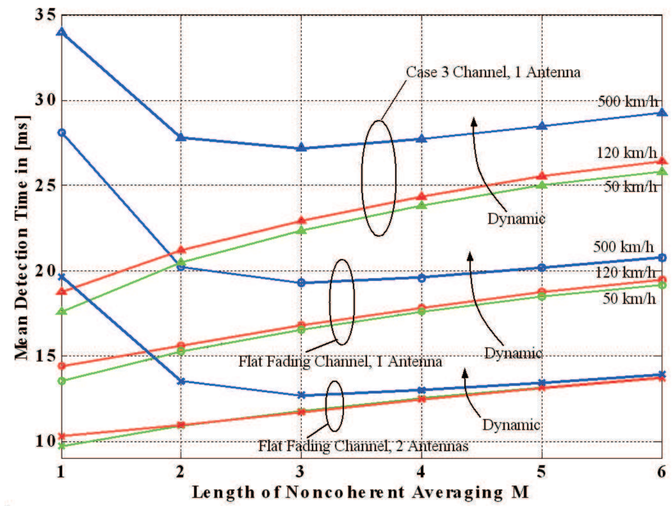

Fig. 5. Mean detection time versus partitioning of the pilot preamble, $P_F = 0.01$, $\rho_0 = -30$ dB, $\Delta\rho = 1$ dB, $N_s = 64$ chips, $N_p = 4096$ chips.

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