Highly Efficient Techniques for Mitigating the Effects of Multipath Propagation in DS-CDMA Delay Estimation

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***Abstract*—Delay estimation in direct sequence CDMA (DS-CDMA) systems is necessary for accurate code synchronization and for applications such as mobile phone positioning. Multipath propagation is among the main sources of error in the DS-CDMA delay estimation process, together with multiple access interference and Non Line-Of-Sight (NLOS) propagation. This paper provides a review of main delay estimation techniques, existing in the literature so far, which are able to cope with multipath propagation, together with our novel delay estimation techniques proposed in the context of DS-CDMA systems. The performance of all these techniques is compared through analysis and simulations, considering also their relative computational complexity and required prior information. Starting from the traditional delay locked loops (DLL) and their improved variants, we discuss several recently introduced delay estimation techniques able to cope with multipath propagation. The characterization of these methods is given in a unified framework, suited for both rectangular and root raised cosine pulse shapes. The main focus in the performance comparison of the algorithms is on the closely-spaced multipath scenario, since this situation is the most challenging for achieving diversity gain with low delay spreads and for estimating Line-Of-Sight (LOS) component with high accuracy in positioning applications.**

***Index Terms*—Closely-spaced multipaths, code synchronization, Line-Of-Sight (LOS) estimation, mobile phone positioning, multipath delay estimation, DS-CDMA, wideband CDMA (WCDMA).**

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I. Introduction

DS-CDMA receivers have to cope with various channel imperfections, such as multipath propagation, time dispersion, and interference. One important task in the channel estimation process is the multipath delay estimation, which, together with the estimation of the complex channel coefficients. allows a good reception of DS-CDMA signal [1]-[3]. A relatively new and particularly challenging problem in DS-CDMA systems is the situation of closely-spaced paths [4]-[9], where different replicas of the transmitted signal arrive at the receiver within less than one chip interval. When Root Raised Cosine (RRC) pulse shape is used, there is still some interference between neighboring paths, even for more distant paths (due to the sidelobes in the correlation function of the RRC pulses). Many techniques exist in the literature to cope with multipath propagation, but they have been mainly studied for the rectangular (RECT) pulse shape [4]-[16] and no comprehensive comparison between existing techniques has been reported in the literature so far.

The purpose of this paper is to give a comprehensive review of the DS-CDMA delay estimation techniques proposed in the literature so far, together with our novel delay estimation techniques, which are able to cope with multipath propagation under various environments and transmission techniques (e.g., bandlimited versus non-bandlimited systems, Rayleigh and Rician channels, closely-spaced and distant paths). Many of the DS-CDMA delay estimation techniques have been initially proposed for systems with RECT pulse shape and short codes (i.e., the code period is equal to the symbol period). However, current DS-CDMA based standards, such as WCDMA standard [17] and satellite CDMA proposals [18], employ RRC pulse shape and long codes (long code means that the symbol period is shorter than the code period), and these features might affect the behavior of the delay estimation algorithms. Here, we develop a unified framework for analyzing and comparing different techniques for a generic DS-CDMA receiver with arbitrary pulse shape and arbitrary pseudo-random codes. The main emphasis is on the closely-spaced multipath scenario (i.e., successive paths are at most one chip apart), because this is one of the most challenging situations in the delay estimation process. Solving closely-spaced paths leads to applications such as maximum ratio combining (MRC) with increased diversity via subchip-spaced components [19] and accurate mobile positioning, when LOS is obstructed by closely-spaced NLOS components.

We start from the classical delay locked loop (DLL) concepts [2], [20]-[22]. We present improved DLL techniques that cope with multipath interference [22]-[24]. We review and extend the concepts of Extended Kalman Filter (EKF) to deal with CDMA delay estimation [25]-[29] and we show via simulations how the EKF behaves in multipath fading channels. Then we present a subclass of Maximum Likelihood (ML) delay estimators, which estimate the multipath delays via interference cancellation based on a reference correlation function. These algorithms are the Multipath Estimating DLL (MEDLL) [4], [5], the Pulse Subtraction algorithm (PS) [6], [30] and the improved PS (iPS) [31]. The delay estimation via deconvolution [7]-[9], [32]-[34] and subspace-based methods [10]-[14], [35] are also discussed and tested in the context of DS-CDMA systems with multipath interference. Finally, we present the delay estimation via Quadratic Programming (QP) [15] and by using the Teager-Kaiser (TK) operator [30], [35]-[37]. The performance of these delay estimation algorithms is compared through analysis and simulations, as well as through consideration of their relative computational complexity and required prior information. Our simulation system is based on WCDMA specifications [17]. However, the presented techniques are valid for any DS-CDMA system.

Three of the algorithms presented here have been introduced by the authors, namely PS [6], [30], iPS [31] and TK [30], [36], [37]. Also, the use of deconvolution and subspace-based delay estimation algorithms in the context of DS-CDMA systems with long codes and RRC pulse shapes has been initially proposed by the authors [35], [38]. The performance of MEDLL has not been reported until now in the literature in the context of DS-CDMA systems with RRC pulse shapes. We also show here the close connection between different ML based approaches (namely MEDLL, PS and iPS). The original idea of the QP approach can be found in [15], but we re-formulate here the quadratic program in a manner more suitable for DS-CDMA receivers. The main novelty of the paper comes from the fact that all the algorithms are defined in a unified framework, which allows us to use them with any type of pulse shape and for any channel profile (for closely-spaced as well as for distant paths), and from the analysis and comparison of these algorithms with the same multipath, multicell, and multiuser scenarios. Moreover, the best solutions for the delay estimation problem are emphasized, and the steps for a practical receiver design are given.

The paper is structured as follows: Section II presents the signal and channel models. Section III describes the delay estimation techniques and their ability to deal with multipath propagation. Section IV shows the simulation comparison between different delay estimation algorithms, for a downlink asynchronous WCDMA system. We also show the comparison between the different algorithms in terms of their relative complexity and needed a priori information. In Section V, we discuss the steps to be taken for a practical CDMA receiver design. Conclusions are drawn in Section VI, with perspectives on new research areas and remaining open issues.

II. SIGNAL AND CHANNEL MODEL

The signal received via a multipath fading channel with additive white Gaussian noise (AWGN) and multiuser interference can be modeled as [39]

$$r(t) = \sum_{v=1}^{N_u} \sqrt{E_b^{(v)}} \sum_{m=-\infty}^{\infty} b_m^{(v)} \sum_{k=1}^{S_F^{(v)}} c_{k,m}^{(v)} \sum_{l=1}^{L^{(v)}} \alpha_{l,m}^{(v)} g(t - mT_{sym}^{(v)} - kT_c - \tau_{l,m}^{(v)}) + \eta_{AWGN}(t). \quad (1)$$

Here, $E_b^{(v)}$ is the bit energy of v-th user, N_u is the number of users, $b_m^{(v)}$ is the m-th transmitted data symbol of the v-th user, $c_{k,m}^{(v)}$ is the code value for the k-th chip of the v-th user, during the m-th symbol (for long codes, the codes are different from one symbol to another), $T_{sym}^{(v)}$ is the symbol interval of v-th user, T_c is the chip interval, $g(\cdot)$ is the chip pulse shape after the matched filtering ($g(t) = g_T(t) \circledast g_R(t)$, $g_T(\cdot)$ is the transmit pulse shape, $g_R(\cdot)$ is the filter matched to the transmitter pulse shape), $S_F^{(v)}$ is the spreading factor of the v-th user ($S_F^{(v)} = T_{sym}^{(v)}/T_c$), and $L^{(v)}$ is the number of channel paths of the v-th user. The standard wide sense uncorrelated scattering (WSSUS) fading model [1] is assumed to hold, which means that the channel paths of a particular user are uncorrelated and the channel autocorrelation function $\phi_l^{(v)}(n,m) = \mathbf{E}[\alpha_{l,n}^{(v)}\alpha_{l,m}^{(v)}]$ is a function of the time difference $n-m$ only ($\mathbf{E}$ is the expectation operator). Here, $\alpha_{l,m}^{(v)}$ is the complex fading coefficient of the l-th path of the v-th user, corresponding to the m-th symbol and $\tau_{l,m}^{(v)}$ is the delay introduced by the l-th multipath component of the v-th user, corresponding to the m-th symbol. The envelope of $\alpha_{l,m}^{(v)}$ coefficients is assumed to be Rayleigh or Rician distributed, i.e., $\alpha_{l,m}^{(v)}$ are complex Gaussian variables with complex means $\mu_l^{(v)}$ ($\mu_l^{(v)} = 0$ for Rayleigh distribution). The code chips are normalized in such a way that $\sum_{k=1}^{S_F^{(v)}} |c_{k,m}^{(v)}|^2 = 1, \ \forall v, m$. The data symbols satisfy the equality $|b_m^{(v)}|^2 = b, \ \forall v, m$, where $b = 1$ if BPSK modulation is used, and $b = 2$ if QPSK modulation is used. In Eq. (1), $\eta_{AWGN}(\cdot)$ is a filtered AWGN noise, of power spectral density $bN_0|G_R(f)|^2$, where $G_R(\cdot)$ is the transfer function of the receiver matched filter. The signal model of Eq. (1) is valid for both uplink and downlink channels. In what follows, we assume that the user of interest is user 1 and we drop the superscript for convenience. The term ‘user’ is employed here in a more general way, including also the channels from the interfering base stations in the case of an asynchronous downlink scenario, such as for WCDMA standard [17]. Due to the typical high number of interfering ’users’, by virtue of central limit theorem, we assume in what follows that the Multiple Access Interference (MAI) is seen as Gaussian noise (standard Gaussian approximation [40]). Hence, the received signal of Eq. (1) can be rewritten as:

$$r(t) = \sqrt{E_b} \sum_{m=-\infty}^{\infty} b_m \sum_{k=1}^{S_F} c_{k,m} \sum_{l=1}^{L} \alpha_{l,m} g(t - mT_{sym} - kT_c - \tau_{l,m}) + \eta_{AWGN\_MAI}(t), \quad (2)$$

where $\eta_{AWGN\_MAI}(\cdot)$ is a noise process incorporating MAI and the filtered AWGN. We remark that the Gaussian approximation does not limit the applicability of the algorithms described here; it only increases the noise level when testing the performance of the algorithms (as shown in Section IV, all the algorithms are tested in multicell and multiuser scenarios). The delay estimation algorithms presented in this paper may be further used in conjunction with any advanced multiuser detection structure [3] such as, for example, the interference cancellation methods (an example of using EKF with interference cancellation has recently been presented in [28]).

Traditional delay estimation techniques are based on the correlation [1]. Because the trend nowadays is towards all-digital receivers [30], in what follows we assume that the correlation and all further processing are done after sampling (in digital domain). However, most of the algorithms described in this paper can also be formulated in the continuous-time domain. The received signal $r(\cdot)$ is downsampled at N_s samples per chip, where $N_s = T_c/T_s$ and T_s is the sampling interval. The samples of the received signal are denoted by $r(iT_s),\ i = 0, 1, \ldots$. The correlator output during the n-th symbol for a discrete time lag τ is given by:

$$y_n(\tau) = \frac{1}{N_C S_F N_s} \sum_{i=nS_F N_s}^{(n+N_C)S_F N_s} r(iT_s) s_n^{ref}(iT_s, \tau), \qquad (3)$$

where N_C is the number of symbols used in the coherent integration and $s_n^{ref}(iT_s, \tau)$ is the reference code sequence (at sample level) with discrete-time lag τ: $s_n^{ref}(iT_s, \tau) = d_n \sum_{k_1=1}^{S_F} c_{k_1,n} g_1(iT_s - nT_{sym} - k_1 T_c - \tau)$. The following cases might be distinguished: a) Non-data aided (NDA): $d_n = 1$ and $N_C = 1$; b) Decision directed (DD): $d_n = \hat{b}_n$ and $N_C > 1$, and c) Data aided (DA): $d_n = b_n$ and $N_C > 1$. Here, d_n is the reference data symbol and $\hat{b}_n$ stands for the estimate of the data symbol b_n. For coherent integration cases, there is a trade-off between the coherent integration gain and the loss of performance because of the changes in the channel coefficients due to fading [33]. Above, $g_1(\cdot)$ denotes the pulse shape of the replica code at the receiver. It follows from Eq. (3) after simple manipulations that

$$\begin{aligned} y_n(\tau) = \sqrt{E_b} \sum_{m=-\infty}^{\infty} b_m d_n^* \sum_{k=1}^{S_F} \sum_{k_1=1}^{S_F} c_{k_1,m} c_{k,n} \sum_{l=1}^{L} \alpha_{l,m} \\ \times \mathcal{R}\Big((n-m)T_{sym} + (k-k_1)T_c + \tau - \tau_{l,m} \Big) + \tilde{\eta}(\tau), \\ \approx \xi_n \sum_{l=1}^{L} \alpha_{l,n} \mathcal{R}(\tau - \tau_{l,n}) + \eta_{ICI}(\tau) + \eta_{ISI}(\tau) \\ + \tilde{\eta}_{AWGN\_MAI}(\tau), \qquad (4) \end{aligned}$$

where ξ_n is a notation standing for $\xi_n = \sqrt{E_b} b_n d_n^*$, $\eta_{ICI}(\cdot)$ is the noise due to inter-chip interference (ICI), $\eta_{ISI}(\cdot)$ is the noise due to inter-symbol interference (ISI), $\tilde{\eta}_{AWGN\_MAI}(\cdot)$ is the filtered and sampled AWGN plus MAI noise, and $\mathcal{R}(\cdot)$ is the discrete pulse shape cross-correlation function:

$$\mathcal{R}(\tau) = \frac{1}{N_C S_F N_s} \sum_{nS_F N_s}^{(n+N_C)S_F N_s} g(iT_s) g_1(iT_s - \tau). \qquad (5)$$

III. Delay estimation techniques

A. Delay Locked Loops (DLLs)

Several types of DLLs have been proposed in the literature for delay tracking. Mainly, there are two types of delay locked loops: non-coherent and coherent [2], [20]-[22]. Non-coherent DLLs use non-linear devices (such as squaring or absolute value) in order to remove the effect of data modulation and channel variations. But standard non-coherent DLLs seem to be adequate only in single user, single path channels [20]. Coherent DLLs do not suffer of squaring losses and gain imbalances. However, in spread spectrum applications, the spread energy-to-noise ratio is typically too low to obtain amplitude and phase recovery before despreading. Therefore, some “quasi-coherent” solutions have been proposed [20], where the phase is corrected after despreading. Since a part of the signal power will be lost, the solution is a quasi-coherent one. Usually, a DLL is based on correlating the incoming signal with a delayed and an early version of a reference signal [2], [20], [22]. In a coherent or quasi-coherent case, the data is removed by complex multiplication of the early and late correlations with the complex conjugates of the estimated data symbols. The effect of the time-varying phases of the channel should also be removed, i.e., multiplication with the complex conjugate of the estimated fading channel complex coefficients is needed. The error signal is passed through a loop filter, which can be an Integrate-and-Dump (I&D) filter [20]. The performance of a DLL is characterized by the so-called S-curve, which presents the error signal as a function of the reference parameter error (the code mismatch in our case). Assuming that the parameter to be estimated is the delay $\hat{\tau}_l$ of the l-th multipath, the S-curve of a coherent DLL is given by [22]:

$$z(\hat{\tau}_l) = \mathbf{E}\left[\hat{\alpha}_{l,n}^* \Big(y_n(\hat{\tau}_l - \Delta) - y_n(\hat{\tau}_l + \Delta) \Big) \right], \qquad (6)$$

where $\hat{\alpha}_{l,n}$ is the estimated complex coefficient of the l-th tap during the n-th symbol, $y_n(\cdot)$ is the output of the correlator from Eq. (3) (in DA or DD mode in order to remove the data modulation), and the expectation operator is taken with respect to the symbol index. The parameter Δ is half of the spacing between the early and late correlations of the DLL (usually $\Delta = T_c/2$). Based on Eq. (4) and (6), the S-curve of a coherent DLL in the presence of multiple paths can be written as:

$$z_{CDLL}(\hat{\tau}_l) \approx \sqrt{E_b} \hat{\alpha}_{l,n}^* \sum_{l_1=1}^{L} \hat{\alpha}_{l_1,n} \mathcal{S}_{DLL}(\hat{\tau}_l - \tau_{l_1}), \qquad (7)$$

where $\mathcal{S}_{DLL}$ is the ideal S-curve for a coherent DLL in single path propagation: $\mathcal{S}_{DLL}(\tau) = \mathcal{R}(\tau - \Delta) - \mathcal{R}(\tau + \Delta)$. The zero-crossings from below of the S-curve announce the presence of a multipath. The classical DLL fails to operate in the presence of multipath interference when successive multipaths are spaced at less than one chip distance [41], [42]. In order to cope with multipath interference, more recent approaches make use of Interference Cancellation (IC) [24], [41], or Interference Minimization (IM) techniques [22], [23] when employing DLLs.

The DLL with IC subtracts the contribution of interfering paths from the output of the finger tracking the path of interest l. The S-curve of a DLL with IC becomes [24], [41]:

$$z_{CDLL\_IC}(\hat{\tau}_l) = z_{CDLL}(\hat{\tau}_l) - \sqrt{E_b}\hat{\alpha}^*_{l,n} \sum_{\substack{l_1=1 \\ l_1 \neq l}}^{L} \hat{\alpha}_{l_1,n} \mathcal{S}_{DLL}(\hat{\tau}_l - \tau_{l_1}). \quad (8)$$

In order to perform the IC of Eq. (8), we need to know the estimates of the delays (relative to the first path delay) and complex coefficients of the interfering paths. In [41], it was assumed that *a priori* knowledge about the channel profile was available, while in [24], the channel coefficients were computed via a ML algorithm, and the initial delay estimates were assumed to be equal to the true path delays. Fig. 1 shows an example of S-curve of a coherent DLL with IC for 2 paths spaced at $0.75\ T_c$ distance. The situation without IC is also shown for comparison and we note that DLL without IC fails to estimate the correct delays if the paths are closely-spaced. For the DLL with IC, if the complex coefficient and the delay of the second path are correctly estimated in advance, we obtain a very good estimate of the first path delay. However, if the delay of the second path is estimated with a small delay error, this error also affects the estimate of the first path delay. This is one of the main drawbacks of the DLLs with IC, because this accurate knowledge about channel coefficients and delays of the interfering paths is hard to be obtained, especially in the case of closely-spaced paths. Besides, ML-based solutions are rather complex for practical implementations.

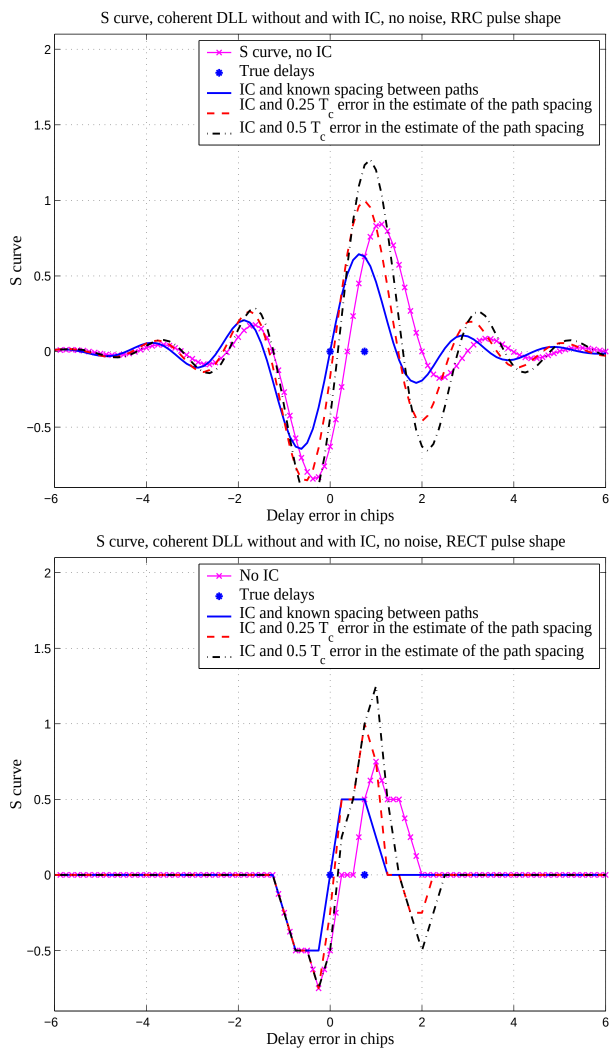

Fig. 1. S curves of coherent DLL without and with interference cancellation on the second path; RRC pulse shape (upper plot) and RECT pulse shape (lower plot), equal average tap powers, $0.75\ T_c$ path spacing, $\Delta = 0.5\ T_c$.

The second approach is to filter the output of the matched filter with some adaptive FIR coefficients [22], [23]. This is the principle of DLL with interference minimization (IM). It has been reported in [22] that these IC and IM techniques have similar results in terms of delay estimation accuracy. We notice that estimates of the interfering paths coefficients and delays are also needed for the DLL with IM scheme and the slowly convergence problem remains, so DLLs with IM suffer from the same disadvantages as the DLLs with IC scheme.

B. *Extended Kalman Filter (EKF)*

The use of Kalman filtering for tracking the time delays in CDMA environment was first proposed in [25]. Later, Kalman filter-based solutions were proved to be near-far resistant in multiuser environments [26]. Due to the fact that the received signal is not a linear function of the multipath delays, an Extended Kalman Filter (EKF) was proposed in [25]. The EKF is a practical approximation to the minimum variance estimator when the observation sequence is nonlinear in the state variables. The state variables are the multipath complex coefficients $\alpha_{l,n}$, the multipath delays $\tau_{l,n}$, and, possibly, the Doppler shift [29]. For clarity reasons, we explain here only the model for channel coefficients estimation and delay estimation, the extension to the Doppler shift estimation being straightforward [29]. The main assumption of EKF algorithm is that the complex channel coefficients and time delays can be modeled as a first-order Gauss-Markov process:

$$\begin{cases} \alpha_{l,n+1} = & \beta_l \alpha_{l,n} + w_{\alpha_{l,n}} \\ \tau_{l,n+1} = & \gamma_l \tau_{l,n} + w_{\tau_{l,n}}, \end{cases} \quad (9)$$

where n is the symbol index (we remark that the estimation can be also done at chip level, before despreading, but in this case the signal-to-noise ratio will be very low), $w_{\alpha_{l,n}}$ and $w_{\tau_{l,n}}$ are mutually independent AWGN processes, β_l is a coefficient accounting for the Doppler spread of path l [43], and γ_l is a coefficient accounting for the delay variation of path l (it is close to unity if the Doppler shift is negligible, as it is the case in terrestrial communications) [25]. This model corresponds to a Rayleigh WSSUS fading model [25].

We remark that the channel models of [25], [26], [29] are different from our model of Eq. (9) in the sense that earlier, the paths were assumed to be uniformly spaced at T_c distance, and the only delay to be estimated was the delay of the first path. Here, we extend the EKF algorithm to model all the path delays; it will increase the number of parameters to be estimated, and hence the complexity. However, it allows for more flexibility as the path delays may be also closely-spaced.

With the assumption of Eq. (9), the Kalman filter equations can be written as

$$\begin{cases} \text{state model:} & \mathbf{x}_{n+1} = \mathbf{F}\mathbf{x}_n + \mathbf{w}_n \\ \text{measurement model:} & y_n(\tau) = \mathcal{H}(\mathbf{x}_n, \tau) + \nu_n, \\ & \text{with } \tau = 0, T_s, \ldots, \tau_{max} T_s. \end{cases} \quad (10)$$

Here, $\mathbf{x}_n$ is the $2L \times 1$ state vector which we want to estimate $\mathbf{x}_n = [\alpha_{1,n}, \ldots \alpha_{L,n}, \tau_{1,n}, \ldots \tau_{L,n}]^T \triangleq [x_{1,n}, \ldots, x_{2L,n}]$, $\mathbf{F} = diag(\beta_1, \ldots \beta_L, \gamma_1, \ldots \gamma_L) \in \mathbb{R}^{2L \times 2L}$, $\mathbf{w}_n$ and ν_n are white Gaussian noise processes, and $y_n(\tau)$ is the scalar observation (despread symbol n with the time lag τ), τ_{max} is the maximum channel delay spread in samples, and $\mathcal{H}(\cdot)$ is the non-linear transform derived from Eq. (4) (the ISI and ICI are included in the ν_n process):

$$\mathcal{H}(\mathbf{x}_n, \tau) = \xi_n \sum_{l=1}^{L} \alpha_{l,n} \mathcal{R}(\tau - \tau_{l,n}) \triangleq \xi_n \sum_{l=1}^{L} x_{l,n} \mathcal{R}(\tau - x_{l+L,n}). \quad (11)$$

The matrix $\mathbf{F}$ need to be estimated in order to solve the EKF system. In our simulations, we assumed that the delay variation coefficients are $\gamma_l = 1,\ l = 1, \ldots, L$, similar to [25], [27], and that the coefficients β_l are equal to $J_0(2\pi \widehat{f_D} T_{sym})$ [43], where $J_0(\cdot)$ is the Bessel function of the first kind, and $\widehat{f_D}$ is the estimated maximum channel Doppler spread (assumed to be equal to the true maximum Doppler spread in our simulations). The EKF algorithm requires the linearization of the transform $\mathcal{H}(\cdot)$. The most commonly used linearization is the first order Taylor expansion [25], [26], [29]. We notice that the multipath delays are real valued parameters, while the measurements are complex-valued. In [29], an improved EKF algorithm was proposed with the Taylor series expansion with respect to the real part of the state vector. An alternative solution is to keep only the real part of delay estimates after each iteration [25], [27]. The prediction and estimation equations for solving the EKF system are found in [25]-[29].

Fig. 2 shows an example of EKF delay estimates in Rayleigh fading channels with closely-spaced paths (similar results have been obtained for distant paths). It can be seen that the initialization parameters for the state vector have a major effect on the convergence of the EKF: we have found via simulations that, when the initial delay error was less than one chip, the EKF always converged (within 1 sample error). But when the initial delay error is greater or equal to one chip, the convergence can be improper (first path converges to the second and vice versa) or the algorithm may diverge. Also, it can be seen in Fig. 2 that, when the algorithm converges, the convergence is achieved in less than 20 symbols. However, the estimated paths will not lock exactly to the right delays, but they might oscillate within 1 sample error around the true path delays (this was the case for oversampling factors of 4 and 8).

The reasons that make the EKF algorithm attractive in the context of multipath delay estimation in the DS-CDMA environments include the good behavior in fading channels, the joint estimation of multipath delays and complex coefficients, the suitability for differentiable pulse shapes (condition satisfied by the RRC pulses), the good convergence if the initial estimates for multipath delays are accurate enough (i.e., errors less than one chip), and the ability to track closely-spaced multipaths.

On the other hand, there are also problems associated with an EKF-based solution, such as its sensitivity to the parameters of the model (e.g., erroneous estimate $\widehat{f_D}$ might affect the EKF performance quite significantly), and the errors introduced by the linearization procedure. Also, the Gauss-Markov channel model might be inadequate (e.g., for Rician fading channels). EKF is solved iteratively, which induces an increased complexity with the number of symbols (or samples [27]). Moreover, if the error in the initial delay estimates are higher than one chip, it is likely that the EKF algorithm does not converge to the right solution. The continuous-time derivatives of the signal are needed for the linearization process. In practice, the discretization process might introduce some errors. Efficient solutions to implement continuous-time derivatives could be based on interpolation and Farrow-based structures [44], [45].

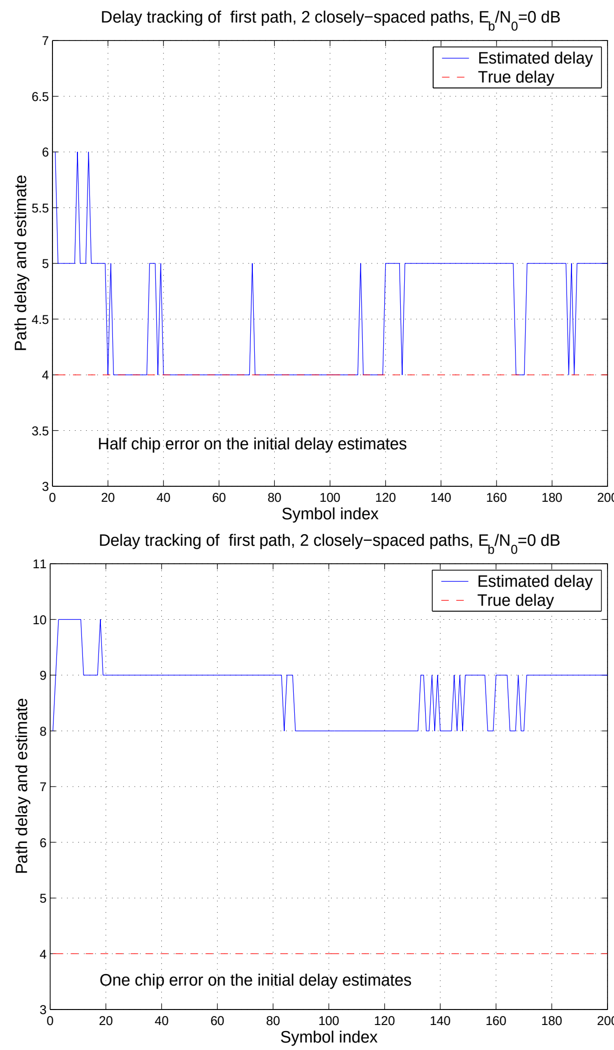

Fig. 2. Symbol level EKF delay estimation for 2 closely-spaced path Rayleigh fading channel, $E_b/N_0 = 0$ dB, $S_F = 64$, $N_s = 4$, average tap gains before normalization are 0 dB and -1 dB, RRC pulse shape: initial delay estimates are equal to half chip error (upper plot) and to one chip error (lower plot). Similar results are obtained for tracking the second path, too.

C. ML-based algorithms

The algorithms grouped here under the generic name of Maximum Likelihood (ML)-based algorithms refer to the algorithms that try to cancel the multipath interference via subtraction from the correlation function of a certain reference pulse, as it is explained in the next two subsections.

C.1 Multipath Estimating Delay Locked Loop (MEDLL)

One of the first algorithms proposed to diminish the effect of multipath propagation in the DS-CDMA delay estimation was introduced by Van Nee [4], [5] and called the MEDLL. We notice that MEDLL is not a 'true' tracking loop in the sense of classical DLLs. MEDLL was developed for GPS positioning applications and its performance was reported only for the RECT pulse shape case. Here, we explain the MEDLL concepts and we extend this approach to DS-CDMA systems employing RRC pulse shape. MEDLL idea is based on the ML theory. It tries to estimate jointly the delays $\tau_{l,n}$, phases $\theta_{l,n}$ and amplitudes $a_{l,n}$ of all the multipaths:

$$\hat{\tau}_{l,n} = \arg\max_{\tau} \Bigg[Re\Bigg\{ \Bigg[y_n(\tau) - \sum_{\substack{l_1=1\\ l_1\neq l}}^{L} \hat{a}_{l_1,n} R_{ref}(\tau - \hat{\tau}_{l_1,n}) \times e^{j\hat{\theta}_{l_1,n}} \Bigg] e^{-j\hat{\theta}_{l,n}} \Bigg\} \Bigg] \tag{12}$$

$$\hat{\theta}_{l,n} = \arg \Bigg[y_n(\hat{\tau}_{l,n}) - \sum_{\substack{l_1=1\\ l_1\neq l}}^{L} \hat{a}_{l_1,n} R_{ref}(\hat{\tau}_{l_1,n} - \hat{\tau}_{l,n}) e^{j\hat{\theta}_{l_1,n}} \Bigg] \tag{13}$$

$$\hat{a}_{l,n} = Re\Bigg\{ \Bigg[y_n(\hat{\tau}_{l,n}) - \sum_{\substack{l_1=1\\ l_1\neq l}}^{L} \hat{a}_{l_1,n} R_{ref}(\hat{\tau}_{l_1,n} - \hat{\tau}_{l,n}) \times e^{j\hat{\theta}_{l,n}} \Bigg] e^{-j\hat{\theta}_{l,n}} \Bigg\}. \tag{14}$$

The correlator output $y_n(\tau)$ should be computed in DA or DD mode for increased performance. Above, $R_{ref}(\cdot)$ is a reference correlation function, $\hat{\tau}_{l,n}$, $\hat{a}_{l,n}$ and $\hat{\theta}_{l,n}$ are the respective estimates of the delay, amplitude and phase of the l-th path during the n-th symbol. The system of Eq. (12) to (14) can be solved by iterative matrix calculation, but in practice, this approach is too costly. Van Nee proposed an IC technique which reduces the complexity [5]. The steps of this IC technique form the MEDLL algorithm and they can be summarized as follows:

Step 1. Calculate the correlation function $y_n(\tau)$. Find the maximum peak (peak 1) of the correlation function $y_n(\cdot)$, and its corresponding delay, amplitude and phase: $\hat{\tau}_1$, $\hat{a}_{1,n}$ and $\hat{\theta}_{1,n}$.

Step 2. Subtract the contribution of the calculated peak to yield a new approximation of the correlation function $y_n^{(1)}(\tau) = y_n(\tau) - \hat{a}_{1,n} R_{ref}(\tau - \hat{\tau}_{1,n}) e^{j\hat{\theta}_{1,n}}$ and find the new peak (peak 2) of the residual correlation function $y_n^{(1)}(\cdot)$, and its corresponding delay, amplitude and phase: $\hat{\tau}_{2,n}$, $\hat{a}_{2,n}$ and $\hat{\theta}_{2,n}$. Subtract the contribution of peak 2 from $y_n(\cdot)$, and find a new estimate of peak 1.

Step 3. Repeat step 2 , until a certain criterion of convergence is met (e.g., until the calculated delays of peaks 1 and 2 vary with less than $0.1\ T_c$ at each new iteration). For more than 2 peaks, continue the above procedure, until all the desired peaks are estimated.

One important issue in implementing the MEDLL algorithm is the choice of an adequate reference function $R_{ref}(\tau)$. In [5], it was proposed that the reference correlation function is measured in the absence of any noise or multipath, using a signal simulator, and that its shape is stored in the memory of the receiver. Furthermore, it was assumed that the reference function covered only the main correlation peak (of length $2T_c$). One drawback of MEDLL in the context of WCDMA applications seems to be its sensitivity to the choice of the reference correlation function. For good results, as those reported for GPS [4], the reference correlation function should take into account not only the pulse shape, but also the spreading code correlation properties. In DS-CDMA applications, computing a reference correlation function for each base station and keeping it in the memory of the mobile receiver is practically infeasible. Instead, we propose to use a generic reference correlation function such as the Autocorrelation Function (ACF) of the transmitter pulse shape.

C.2 Pulse subtraction algorithms

The MEDLL algorithm suffers of increased complexity with the number of paths. A lower complexity method based on pulse subtraction (PS) has been proposed in [6] (a somehow similar approach is also found in [46]). This method is similar to the MEDLL algorithm using only one step (after one subtraction is performed, no improvement in the previous estimates is made). The search for the multipaths is performed by trying to approximate $y_n(\tau)$ by a superposition of reference pulses $\mathcal{R}_{ideal}(\cdot)$ with different amplitudes, where the reference pulse is taken equal to the ACF of a RECT pulse shape, based on the ideas developed in [6], [31]. Each multipath delay is estimated after the contribution of previous estimated multipaths has been subtracted from the correlation function $y_n(\tau)$. The PS algorithm terminates when a pre-defined number of delays is reached (if knowledge about the number of channel paths exist) or until the maximum of the residual correlation function is less than a certain threshold. The PS algorithm has the advantage of simplicity, but its capability to solve closely-spaced multipaths remains rather moderate, especially when RRC pulse shapes are employed. An improved pulse subtraction algorithm (iPS) has been introduced by the authors in [31] to deal with RRC pulse shapes and closely-spaced multipaths. The iPS stages are:

Step 1. Calculate the correlation $y_n(\tau)$ between the received signal and the reference signal. Find the global maximum $\widehat{\tau}_g = \arg\max_\tau |y_n(\tau)|$ (the algorithm is repeated for each symbol, but, for clarity reasons, we dropped the subscript n from the delay indices).

Step 2. For each value $\widehat{\tau}_1^{(m)}$ taking M discrete values in the interval around the global maximum $[\widehat{\tau}_g - \Delta\tau; \widehat{\tau}_g + \Delta\tau]$, estimate successively:

$$\widehat{\tau}_l^{(m)} = \arg\max_{\tau} \left| y_n(\tau) - \sum_{l_1=1}^{l-1} y_n(\widehat{\tau}_{l_1}^{(m)}) \mathcal{R}_{ideal}(\tau - \widehat{\tau}_{l_1}^{(m)}) \right|,$$
$$l = 2,3,\ldots,L,\ m = 1,\ldots,M.$$

Here, $\Delta\tau$ is the variation interval around the global maximum. We recall that all the delays here are given in samples (we considered the discrete-time domain correlation).

Step 3. Compute

$$y_{est}(\widehat{\tau}_1^{(m)},\tau) = \sum_{l=1}^{L} y_n(\widehat{\tau}_l^{(m)}) \mathcal{R}_{ideal}(\tau - \widehat{\tau}_l^{(m)}),$$
$$m = 1,2,\ldots,M.$$

and choose the m^{th} set $\{\widehat{\tau}_l^{(m)}\}_{l=1,2,\ldots,L}$ which minimizes the mean square error between $y_n(\tau)$ and $y_{est}(\widehat{\tau}_1^{(m)},\tau)$. This set represents the iPS-estimates of the multipath delays. Empirically, the best $\Delta\tau$ is $0.25T_c$, i.e., we assume that due to the merging between multipaths, an error of maximum $0.25T_c$ can occur in the delay of the strongest multipath.

D. Deconvolution methods

Deconvolution methods are means of inverse filtering. One of the adverse effects of inverse filtering, when noise is present, is the noise enhancement. The noise enhancement effect can be reduced by using the so-called constrained inverse filtering methods. These methods are *constrained* in the sense that they do not allow the output values to lie outside some predefined set or in the sense that the inverse operator is never completely formed, but only approximated iteratively. Among the constrained inverse filtering methods, the best known ones are the Least-Squares (LS) techniques [7], [32], [33] and the Projection Onto Convex Sets (POCS) algorithm [8], [9], [34], [38].

If we group in a vector the correlation samples $y_n(\tau)$ at different time lags between 0 and maximum channel delay spread $\tau_{max}T_s$, we can define the following vector of correlation outputs: $\mathbf{y}_n = [y_n(0), y_n(T_s), \ldots, y_n(\tau_{max}T_s)]^T \in \mathbb{C}^{(\tau_{max}+1)\times 1}$. Eq. (4) can therefore be rewritten as

$$\mathbf{y}_n = \xi_n \mathbf{G}\mathbf{h}_n + \mathbf{v}_n, \tag{15}$$

where $\mathbf{G}$ is the pulse shape deconvolution matrix with elements $g_{i,j} = \mathcal{R}(i-j)$, $i,j = 0\ldots,\tau_{max}$, and $\mathcal{R}(\cdot)$ given by Eq. (5), $\mathbf{v}_n$ is the superposition of ICI, ISI, MAI and AWGN noises after the despreading operation (which may be assumed Gaussian by virtue of central limit theorem), $\mathbf{h}_n$ is a $(\tau_{max}+1)\times 1$ vector with elements $h_{l,n} = 0$ if no multipath is present at the time delay l, and $h_{l,n} = \alpha_{l,n}$ if index l corresponds to a true path location. In the least-squares sense, we are looking for the vector $\mathbf{h}_n$ which minimizes the l2-norm $||\mathbf{y}_n - \xi_n\mathbf{G}\mathbf{h}_n||_2$. Resolving multipath components refers to the problem of estimating the non-zero elements of the unknown gain vector $\mathbf{h}_n$. The LS solution for $\mathbf{h}_n$ is given by

$$\tilde{\mathbf{h}}_n^{LS} = (\xi_n \mathbf{G}^H\mathbf{G})^{-1}\mathbf{G}^H\mathbf{y}_n. \tag{16}$$

For more accurate estimates, non-coherent averaging may be used [33]

$$\hat{\mathbf{h}} = \frac{1}{N_{NC}} \sum_{n=1}^{N_{NC}} \left| \tilde{\mathbf{h}}_n^{LS} \right|^2, \tag{17}$$

where N_{NC} is the non-coherent integration length. Now, the multipath delays are obtained by taking the maximum peaks of the estimated vector $\hat{\mathbf{h}}$ of Eq. (17).

The noise enhancement is a known drawback of least-squares based methods. To better cope with noise and to solve the problem of closely-spaced paths, a constrained iterative deconvolution technique called projection onto convex sets (POCS), was introduced in [8], [9] for a Rake receiver with RECT pulse shapes, and later applied for narrowband signals with Welsh-Costas pulses as well [34]. The POCS estimation takes place in several stages. At stage $k+1$, the POCS estimate can be written as [34]:

$$\tilde{\mathbf{h}}_n^{POCS,(k+1)} = \tilde{\mathbf{h}}_n^{POCS,(k)} + \left(\frac{1}{\lambda} I + |\xi_n|^2 \mathbf{G}^H\mathbf{G}\right)^{-1} \times \xi_n^* \mathbf{G}^H \left(\mathbf{y}_n - \xi_n \mathbf{G}\tilde{\mathbf{h}}_n^{POCS,(k)}\right), \tag{18}$$

where λ is a constant determining the convergence speed and I is the unity matrix. Based on simulations, about $N_{iter} = 5$ iterations in POCS algorithm seemed to be sufficient and $\lambda = 0.5$ proves to be a good choice. Similarly with LS algorithm, the estimates can be further improved by coherent and non-coherent averaging (see Eq. (17)).

Generally, the deconvolution methods have been shown to exhibit good subchip resolution in multipath static channels with RECT pulse shapes [8], [9] and with Welsh-Costas pulses [34]. In our simulations, we evaluate the performance of these algorithms in fading channels for both RECT and RRC pulses. The main drawbacks associated with the convolution methods are their increased complexity and memory requirements (mainly due to matrix inversions) and their lack of robustness at high noise levels.

E. Subspace-based algorithms

The subspace-based algorithms involve the decomposition of the space spanned by the observation vector (i.e., the vector formed by the received signal samples) into several subspaces, usually a noise subspace and a signal subspace. Furthermore, these algorithms use the orthogonality property between noise subspace and signal subspace in order to estimate the channel parameters. The subspace approaches involve eigenvector decomposition of high-order matrices, and they remain usually very complex for practical applications. The main advantage of the subspace-based methods is their increased resolution in the parameter estimates. The most known subspace-based methods which have been employed in delay estimation applications are the Multiple Signal Classification (MUSIC) [10]-[14], [35] and Estimation of Signal Parameters via Rotational Invariance techniques (ESPRIT) [10]. In [10] it was shown that these two methods are quite similar, both from the point of view of their performance and complexity. This is the reason why we consider only the MUSIC algorithm. We remark that, when using long codes in the CDMA system (such as the WCDMA standard), the noise subspace usually cannot be extracted directly from the observation vector space because the noise and signal spaces vary from symbol to symbol. This means that in long code systems we usually expect some extra errors in the signal and noise space estimates (and hence, in the delay estimates) due to the aperiodicity of the code. The performance of subspace-based algorithms in the context of CDMA delay estimation has traditionally been asserted in the literature only for short code systems [11], [12]-[14]. Recently, the performance of MUSIC delay estimation with long codes has been studied by the authors [35].

Among the advantages of MUSIC-based delay estimation, we have the high resolution of the estimates and the near far resistance in multiuser environments [14]. However, the high complexity associated with the matrix eigen-decomposition makes the MUSIC algorithm quite difficult to be used in practical applications. Moreover, the subspace based algorithms are blind algorithms in the sense that they do not use information about the transmitted data. However, this information is sometimes available, e.g., via pilot channels in WCDMA. Improved schemes should also take into consideration the knowledge about the transmitted pilot data.

F. Quadratic optimization methods

Keeping in mind the matrix form of the correlator output from Eq. (15), the delay estimation problem can be reformulated as a quadratic program (QP). The criterion proposed in [15] for solving closely-spaced echoes in static channels is the $l1$-minimization criterion:

$$\tilde{\mathbf{h}}_n^{QP} = \min_{\mathbf{h}_n} ||\mathbf{h}_n||_1, \text{ under a quadratic constraint:} \quad \left(||\mathbf{y}_n - \xi_n \mathbf{G}\mathbf{h}_n||_2 \right)^2 < \rho. \tag{19}$$

Here, $||\cdot||_1$ is the $l1$-norm (i.e., if $\mathbf{x}$ is a vector with N elements, then $||\mathbf{x}||_1 \triangleq \sum_{i=1}^{N} |x_i|$), and $||\cdot||_2$ is the $l2$-norm (i.e., $||\mathbf{x}||_2 \triangleq \sqrt{\sum_{i=1}^{N} |x_i|^2}$), and ρ is a noise tolerance index.

Alternatively, we suggest here an $l2$-minimization criterion, under a linear constraint, in order to reduce the implementation complexity. The parameters to be estimated are the envelopes $\zeta_n = [\zeta_{1,n}, \ldots, \zeta_{L,n}]^T$ of the complex channel coefficients ($\zeta_n = |\mathbf{h}_n|$):

$$\tilde{\zeta}_n^{QP} = \min_{\zeta_n} \left(\Big|\Big| |\mathbf{y}_n| - \xi_n \mathbf{G}\zeta_n \Big|\Big|_2 \right)^2, \text{ under the } \tau_{max} \text{ constraints:} \quad \zeta_{l,n} \leq \rho_l, \; l = 0, 1, ..., \tau_{max}. \tag{20}$$

The averaging given in Eq. (17) can be used to improve the QP-estimates.

Among the advantages of the QP approach, the followings have been reported in the literature so far: the capacity to solve closely-spaced paths and the good convergence even if only a low number of symbols are available for delay estimation [15]. The drawbacks of QP-based delay estimation are its high complexity associated with solving the minimization program (it can be somehow reduced if we reduce the search space, e.g., by looking for the peaks within 1 or 2 chip intervals) and its sensitivity to noise. Results with QP delay estimation in fading channel environments have not been reported in the literature before.

G. Teager-Kaiser (TK) operator-based algorithm

The nonlinear quadratic TK operator was first introduced for measuring the real physical energy of a system [47]. It was found that this nonlinear operator is simple, efficient and able to track instantaneously-varying spatial modulation patterns [48]. Since its introduction, several other applications have been found for TK operator, one of the most recent being the estimation of closely-spaced paths in DS-CDMA systems, introduced by the authors for GPS and WCDMA systems [30], [35]-[37]. The discrete-time TK operator for a complex valued signal $x(n)$ is given by [49]

$$\Psi_d[x(n)] = x(n)x^*(n) - \frac{1}{2}[x(n-1)x^*(n+1) + x(n+1)x^*(n-1)]. \tag{21}$$

Now, if we apply this operator to the ACF $\mathcal{R}_{ideal}(\cdot)$ of a RECT pulse shape of duration N_s, it is straightforward to show that the TK energy of this function will exhibit a peak at 0 lag. In the presence of multipaths and multi-user interference, the correlation function $y_n(\tau)$ can be seen as a superposition of shifted and distorted versions of pulse shape ACFs. We showed in [37] that, when the RECT pulse shape is used, the TK operator applied to the output of the correlator in the presence of multipath interference provides clear time-aligned peaks located at the closely-spaced paths positions, in the presence of a certain noise floor. The TK algorithm can be extended to the band-limiting pulse shapes, such as RRC pulse shapes. However, for RRC pulse shape, the performance is not as good as for the RECT pulse shape, due to the fact that the ACF of RRC is not piecewise linear (as the ACF of a RECT pulse) and due to the presence of side-lobes, which increase the inter-path interference. The output of the TK operator can be further averaged to improve the estimates, and the decision variable becomes similar to Eq. (17): $J^{TK}(\tau) = \frac{1}{N_{NC}} \sum_{n=1}^{N_{NC}} |\Psi_d[y_n(\tau)]|^2$.

The generic block diagram of the deconvolution and TK-based delay estimation algorithms is illustrated in Fig. 3. In order to increase the delay estimation accuracy, one of the proposed signal processing algorithms (e.g., LS, POCS, TK) is applied at the output of the correlator. After the square envelope detection, we perform non-coherent averaging over several symbols in order to reduce the effect of noise. The search for the maximum peaks is done on the averaged correlation function (also referred to as cost function).

We remark that the block diagram for ML-based algorithms (i.e., MEDLL, PS and iPS) is similar to that of Fig. 3 with two modifications: firstly, the pulse subtraction may be applied either before the square envelope detection (i.e., we also need the phase estimates of the channel paths), or after the square envelope detection and non-coherent averaging (no phase estimates are needed, therefore this variant has lower complexity than the previous one, and it was selected for our simulations). Secondly, the search for maximum peaks is different compared to the deconvolution and TK-based methods (i.e., it is done via pulse subtraction, as explained in Section III.C).

IV. Performance comparison

The performance of the different delay estimation algorithms was studied in a multi-cell downlink WCDMA scenario. The

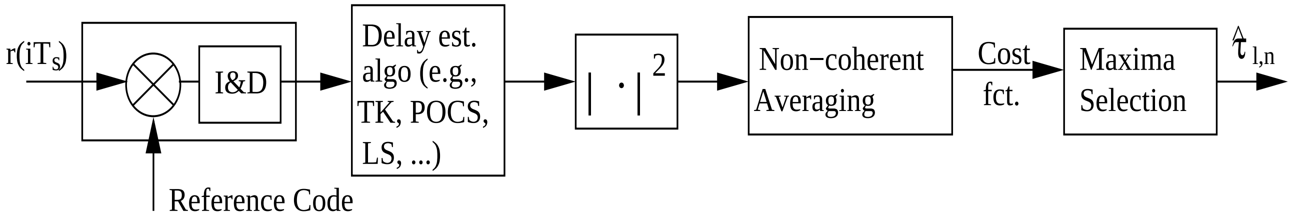

Fig. 3. Generic block diagram of the DS-CDMA feedforward delay estimation (various algorithms, such as TK or POCS, may be used to increase the delay estimation accuracy.

parameters of the simulations are given in the captions of each figure. The correlation measurements were done on the Common Pilot Channel (CPICH)[17], which was assumed to be 5 dB stronger than the sum of the Dedicated Physical Data Channels (DPDCH) in each cell [17]. The Dedicated Physical Channels (DPCH)[1] within one cell are transmitted synchronously, but in different cells they are not synchronized between them [17]. Therefore, we have asynchronous intercell interference and synchronous intracell interference (with the exception of CPICH channel which is also transmitted asynchronously compared to the DPCH channels), as specified in the current WCDMA standard [17]. The channel was modeled as Rayleigh or Rician distributed with L paths in the desired cell and random number of paths in the cells of the interfering BS. The channel delays were modeled as constant over $N_{NC} * N_C$ symbols. At each $N_{NC} * N_C$ symbols, a set of L delays were generated according to the uniform distribution in the interval $[\epsilon_l, \epsilon_{l+1})$, $l = 1, \ldots, L$, where $\epsilon_l = N_s + (l-1)N_s D_{max}/2$ and D_{max} is the maximum separation between two consecutive path delays. The oversampling factor was $N_s = 8$. The delays were rounded to the nearest integer, such that only integer multiples of the sampling interval were allowed. The near far ratio (NFR) between different BSs is defined as the ratio between the CPICH powers of interfering BSs and the CPICH power of the desired BS, similar to [14]. CPICH channels have $S_F = 256$, except for the case when the MUSIC algorithm is also included in the comparison. MUSIC algorithm is only shown for $S_F = 64$ due to its prohibitive complexity. We assumed that the number of estimated paths was equal to the true number of paths L. The comparison criterion is the probability that all the paths are estimated with at most one sample error: $P_{all} = \prod_{l=1}^{L} Proba(|\hat{\tau}_l - \tau_l| \leq 1)$, where τ_l and $\hat{\tau}_l$ are given in samples. The mapping between the estimated delays and the true path delays is done as explained in [46].

Figs. 4 to 6 show the algorithm comparison for Rayleigh fading channels with closely-spaced and distant paths (similar results were obtained for Rician channels). MUSIC algorithm is only shown for the situation with $S_F = 64$ and coherent integration over one symbol only, which corresponds to an NDA approach (Fig. 4). The curves in the legend are shown in decreasing order of their performance at NFR$= -10$ dB. Both RECT and RRC pulse shapes were used in the simulations, in order to see the deterioration of performance when RRC pulses were

[1] One DPCH channel in downlink WCDMA contains the dedicated pilot (or control) channel (DPCCH) time-multiplexed with the data symbols (i.e., DPDCH) [17].

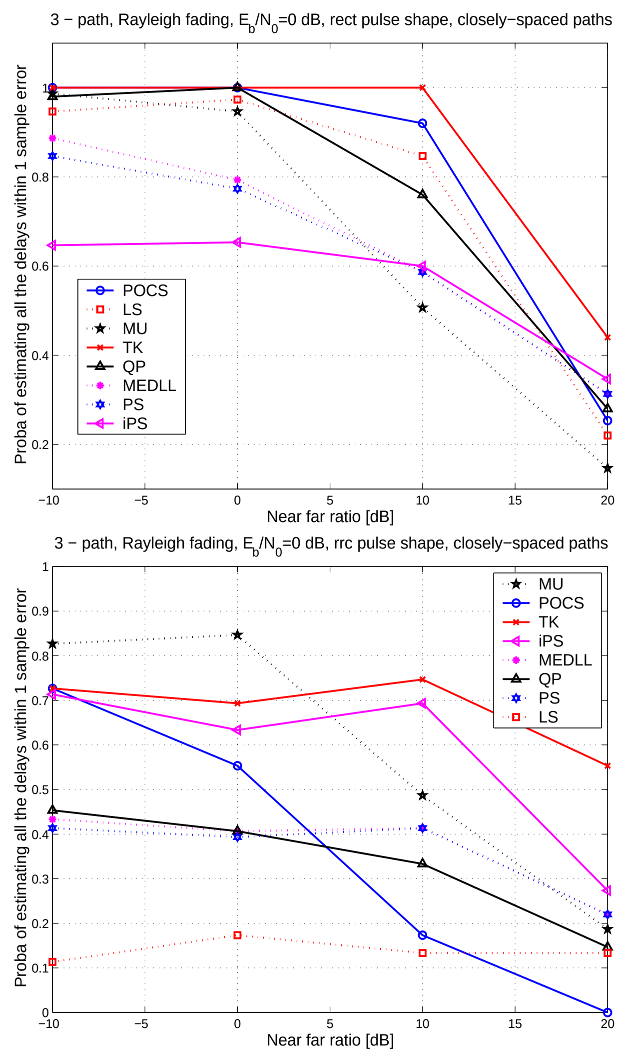

Fig. 4. Comparison of delay estimation algorithms for Rayleigh fading channels with 3 *closely-spaced* paths ($D_{max} = 1$ chip), for RECT and RRC pulse shape. The average tap powers are 0, -2, and -4 dB, there are 2 BTS with 32 users per BTS; CPICH spreading factors are $S_F = 64$, DPDCH spreading factors are 128, $E_b/N_0 = 0$, $N_s = 8$, $N_C = 1$, and $N_{NC} = 40$.

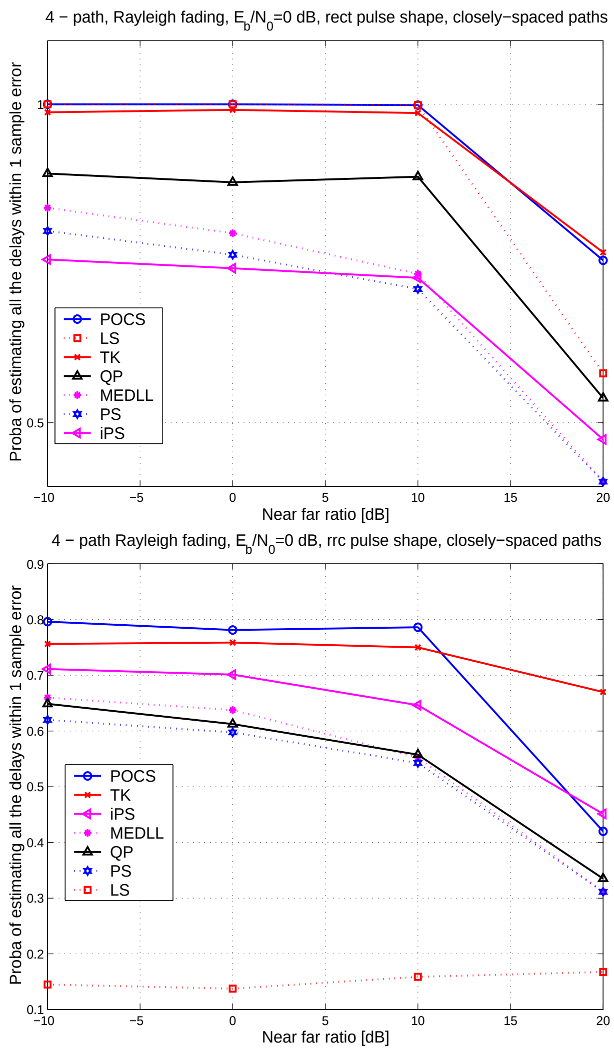

Fig. 5. Comparison of delay estimation algorithms for Rayleigh fading channels with 4 *closely-spaced* paths ($D_{max} = 1$ chip), for RECT and RRC pulse shape. The average tap powers are -2, 0, -1 and -4 dB, there are 3 BTS with 64 users per BTS; CPICH spreading factors are $S_F = 256$, DPDCH spreading factors are 128, $E_b/N_0 = 0$, $N_s = 8$, $N_C = 10$, and $N_{NC} = 4$.

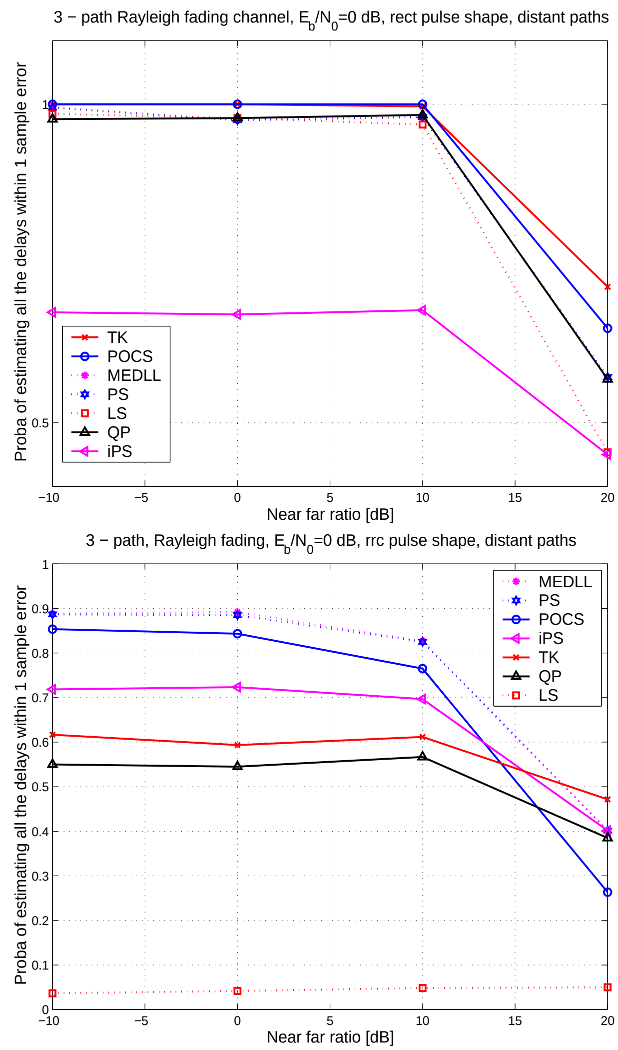

Fig. 6. Comparison of delay estimation algorithms for Rayleigh fading channels with 3 *distant* paths ($D_{max} = 3$ chips), for RECT and RRC pulse shape. The average tap powers are -2, 0, and -1 dB, there are 2 BTS with 32 users per BTS; CPICH spreading factors are $S_F = 256$, DPDCH spreading factors are 128, $E_b/N_0 = 0$, $N_s = 8$, $N_C = 5$, and $N_{NC} = 4$.

employed. We notice, by comparing the upper and lower plots of each figure, how the different algorithms are affected by the pulse shapes. Clearly, the LS algorithm is the most heavily affected by the pulse shape: it has a very good performance for RECT pulse shape (at low NFR), but it fails completely in the RRC case. For closely-spaced paths, POCS, MUSIC, and TK algorithms have good performance at low NFR for both RECT and RRC cases. POCS algorithm is very sensitive to the coherent integration length, as seen from comparing the NDA approach of Fig. 4 with DA approaches of Fig. 5 and 6. This is due to the increased noise level in the NDA approach and from the fact that deconvolution algorithms are known to be quite sensitive to noise. For NFR ratios higher than 10 dB, all the algorithm suffer from fast degradation of performance. TK algorithm is the most near-far resistant, followed by the POCS algorithm in the DA cases. For Rayleigh fading channels, MEDLL and PS algorithms have quite similar performance, but if the channel has a strong LOS component (i.e., Rician fading), MEDLL algorithm becomes significantly better. The iPS algorithm performs quite well for RRC pulses, but it is not suited for the RECT case. For RECT pulses, the QP algorithm has a good performance, but for RRC pulses it has only moderate performance in both closely-spaced and distant scenarios. The table given in Fig. 7 shows the comparison between the different discussed algorithms in terms of complexity, a priori information needed as input, and performance in closely-spaced and distant paths scenarios. This comparison is based on the algorithm descriptions of Section III and on the simulation results.

V. Receiver Design Issues

The delay estimation algorithms presented here are targeting two main CDMA applications, namely, the code synchronization block design in DS-CDMA receivers, and the mobile positioning applications. For both applications, the estimation of the multipath delays in fading channels is a very challenging task at the receiver, especially when closely-spaced paths are present.

	A priori information	Complexity	Ability to solve closely spaced paths	Ability to solve distant paths
DLL	Initial coarse estimates of the path delays	Moderate	Low	Moderate (dependent on initial conditions)
DLL with IC or IM	Initial coarse estimates of the path delays Knowledge of the spacing between paths Knowledge of the channel coefficients of the interfering paths	Moderate	Moderate (dependent on initial conditions)	Moderate (dependent on initial conditions)
EKF	Initial coarse estimates Accurate state space model	Moderate to high (dependent on the size of the state vector)	Moderate (dependent on the initial conditions)	Good
MEDLL	Reference correlation function	Low to moderate (dependent on the number of iterations)	Moderate	Good
PS	Reference correlation function	Low	Moderate	Good
iPS	Reference correlation function	Low to moderate (dependent on the number of iterations)	Good for RRC pulse Moderate for RECT pulse	Moderate for RRC pulse Low for RECT pulse
LS	None	Moderate	Good for RECT pulse and for good SNR, very low otherwise	Good for RECT pulse and for good SNR, very low otherwise
POCS	None	Moderate to high (dependent on the number of iterations)	Good	Good
MUSIC (MU)	None	High	Good	Moderate
QP	None	High	Good for RECT pulse Moderate for RRC pulse	Moderate
TK	None	Low	Good	Good for RECT pulse Moderate for RRC pulse

Fig. 7. Comparative behavior of delay estimation algorithms

When considering the design of a CDMA receiver, the performance of the delay estimators is defined by the error statistics (such as the acquisition probability, as it was considered in Figs. 4 to 6, the mean or the variance of the error, a.s.o.), and by the Bit Error Rates (BER) obtained under various multipath propagation scenarios and with various channel coefficient estimation and detection structures [3]. However, for mobile positioning purpose, a very high accuracy of the delay estimation of LOS component is desired, and the acquisition probability is the most illustrative measure of performance in this context. For illustration purpose, we selected the acquisition probabilities as significant measures of performance for both applications. The joint root mean square error (RMSE) curves were also studied during the simulations and they mainly gave similar information as the acquisition probabilities.

Targeting one of these two applications, the main steps to be considered in a CDMA receiver design process related to the delay estimation part are:

1. Choose the general structure of the delay estimator, i.e. feedback versus feedforward implementation. As seen above, the feedback structures (e.g., DLLs and EKF) suffer of feedback error propagation and need very good delay estimates (within one chip error) available from an initial acquisition phase[2]. Moreover, the performance of the feedback loops is rather limited in the presence of closely-spaced paths. Therefore, the authors would opt for a feedforward implementation, as that illustrated in Fig. 3. In a feedforward structure, no assumption about the initial delay estimation error is made (i.e., the error can be higher than one chip), and therefore, there are less requirements for the acquisition stage. The search range is from 0 to $(\Delta t)_K \triangleq \widehat{\tau}_{max} T_s$, where $\widehat{\tau}_{max}$ is an estimate of the maximum delay spread of the channel in samples (and may be taken in such a way to cover all the possible propagation scenarios[3]).
2. Once the feedforward architecture is selected, there are several signal processing algorithms that can be used to increase the delay estimation accuracy (they were described in Sections III.C to III.G):
   (a) From the point of view of the Rake receiver design, the lowest complexity solutions (namely PS and TK) would naturally seem the best choice. The reason is as follows: RMSE simulation results (not included here from lack of space) showed that most of the discussed algorithms have a RMSE between $0.1\ T_c$ and $0.2\ T_c$ for NFR up to 0 dB and, usually, delay error differences which are less than $0.1\ T_c$ do not have a significant impact on BER. Moreover, when designing the Rake receiver, the complexity issue is of utmost importance and typically prevails over the performance, if there is no significant degradation in performance. Both PS and TK algorithms have at least moderate performance in all studied multipath scenarios, and both have reduced complexity, therefore they are viable choices for a low complexity Rake receiver.
   (b) From the point of view of the mobile positioning applications, very good accuracy is needed, even at the expense of a slightly increased complexity. In this context, the best options are TK and POCS delay estimation algorithms (their implementation structure is shown in the block diagram of Fig. 3). POCS algorithm has the best performance under various pulse shapes and closely-spaced as well as distant paths; however, its complexity (i.e., the number of required multiplications per delay estimate) is of the order of $\mathcal{O}(N_{iter}\widehat{\tau}_{max}^3)$, where N_{iter} is the number of POCS iterations (e.g., equal to 5) and $\widehat{\tau}_{max}$, as explained above, is an integer number related to the maximum delay spread of the channel. The high complexity of POCS is mainly due to the matrix inversion operations. However, for indoor channels (i.e., with low delay spreads) and for oversampling factors up to $N_s = 4$, the number of samples $\widehat{\tau}_{max}$ is in the range of 6 to 25, and therefore the complexity of POCS is not very high. Moreover, under the assumption of perfect power control, the matrix to be inverted in Eq. (18) of POCS can take only a finite number of values (according to the transmitted bit energy[4] E_b. Therefore, it can be computed only once (for all the possible E_b values), at the beginning of the operations, and stored in the memory of the receiver for further use, thus decreasing the complexity.

   The TK algorithm has lower complexity, of the order of $\mathcal{O}(\widehat{\tau}_{max})$, but its performance is not as good as POCS performance for RRC pulse shaping and distant paths at low NFR. From the design point of view, the best trade-off between accuracy and complexity should be chosen according to the estimated maximum delay spread of the channel.
3. Another issue in the design process is the choice of an oversampling versus interpolation-based architecture. For a digital implementation, increased accuracy of the delay estimates can be achieved either by using a sufficiently high oversampling factor, or by using a moderate oversampling (e.g., for RRC pulse shape, 2 samples per chip may be used, which are enough to satisfy the Nyquist criterion), followed by interpolation. Details of an interpolation based architecture and its advantage over the oversampling based architecture (in terms of number of operations) can be found in [50].

VI. Conclusions

This paper gave an overview of the delay estimation techniques able to cope with multipath propagation in DS-CDMA environments. Existing approaches and novel techniques were described in a unified framework suitable for DS-CDMA systems. These techniques were compared from the point of view of their performance, complexity, and a priori information needed. The simulation results were carried out for a WCDMA downlink system, but the techniques presented here are valid for any DS-CDMA system. We noticed that most of the pre-

[2] For the distinction between delay acquisition and delay tracking in CDMA environments a good description can be found in [2].

[3] E.g., an estimated maximum delay spread $(\Delta t)_K = 35\ \mu s$ would cover most of the WCDMA propagation scenarios [17].

[4] We recall that the matrix to be inverted in Eq. (18) depends on $|\xi_n|^2$, which is equal to $b^2 E_b$ for coherent integration cases, and to bE_b for non-coherent integration cases.

sented techniques had reasonable performance if the multipaths were distant enough (at least one chip apart). The challenge appears for closely-spaced paths, where traditional methods such as DLLs failed to work. The algorithms based on feedback loops such as DLL with IC, DLL with IM, and EKF may be used up to a certain extent to solve the problem of closely-spaced paths, but they are usually quite complex and require some a priori information which is not easy to obtain. Good performance for closely-spaced paths was obtained via TK, POCS, and MUSIC algorithms independently of the pulse shape, and also via QP and LS algorithms for RECT pulse shapes and via iPS algorithm for RRC pulses. Moderate performance, independently of the pulse shape, is also achieved with the very-low complexity PS algorithm. We saw that among the presented techniques, the methods with a significant potential, from the point of view of their accuracy, in DS-CDMA applications are POCS and TK. POCS algorithm performs very well in good noise conditions, but tends to loose its performance rather quickly at high NFR or low coherent integration times. Low cost implementations may successfully use PS and TK algorithms. TK algorithm has also the advantage of simplicity and efficiency compared to the other presented algorithms. As open issues and future research topics, we mention the use of combined techniques to improve estimation (especially under strong interferences), the optimization of TK and POCS algorithms under various noise conditions, and the use of more advanced IC methods to deal with multiuser interference. We also remark that for DS-CDMA positioning applications, finding optimal thresholds to estimate the number of significant paths is crucial for LOS estimation with high accuracy.

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