Channel Estimation for DS-CDMA with Transmit Diversity over Frequency Selective Fading Channels

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Abstract

RAKE receivers for DS-CDMA communications over fading channels require the knowledge of instantaneous complex channel phasors for each of the multipath diversity branches (or RAKE fingers). If transmit diversity is employed by using N transmit antennas in the basestation, N such fading channels need to be estimated in the mobile receiver. In this paper, a channel estimation algorithm which estimates two fading channels in the case of dual TX diversity is presented, making use of known pilot symbols. The proposed scheme performs remarkably well in Rayleigh fading multipath environments with little additional complexity if compared to a scheme where only one channel needs to be estimated (i.e. no TX or RX diversity). The performance of the algorithm is assessed by means of computer simulations.

1. Introduction

The RAKE receiver is a low-complexity structure for the reception of code-division multiple access signals over multipath fading channels [2] and will be used in the first 3G handsets. It requires knowledge of the instantaneous channel state (timing and complex-valued attenuation) for each multipath. In absence of perfect channel state knowledge, the receiver must attempt to estimate the necessary channel parameters and compensate for their effects on the received signal, a task commonly referred to as *synchronization*. The concept of estimating those parameters and then using them as if they were the true values is called *synchronized detection* [3] and is used in virtually every digital receiver implementation.

The focus of this work lies on the estimation of the time-varying complex-valued channel phasors, which are needed in the RAKE receiver for maximum-ratio combining. In [6], the authors presented phasor estimation algorithms based on postprocessing maximum-likelihood (ML) estimates of the channel phasors with a Wiener filter. In this paper, we shall consider a system employing dual transmit diversity, i.e. using two transmit and one receive antenna. Assuming equal power-delay profiles but uncorrelated fading on the two channels, two uncorrelated phasors need to be estimated in each Rake finger. If pilot symbols are transmitted on both transmit antennas in an alternate fashion, ML estimates can be generated for both phasors in each finger. Those can then be postprocessed in a similar way as in [6], taking into consideration the irregular pilot pattern.

The paper is organized as follows. In Section 2, the system model is introduced for the RAKE receiver and the transmit diversity scheme. In Section 3, a short review of the phasor estimation technique from [6] is given and the new phasor estimation for the system with transmit diversity is presented. Simulation results and a conclusion are presented in Section 4.

2. System Model

2.1 DS-CDMA system without antenna diversity

In CDMA communications, the user data symbols $\{a_k\}$ are oversampled by the spreading factor $N = T/T_c$ and then multiplied by a user spreading sequence **d**, which in the case of the third-generation standard draft from 3GPP [5] consists of length N user-specific OVSF spreading sequences as well as much longer basestation-specific scrambling sequences. T and T_c are the symbol and chip duration, respectively. Throughout this paper we shall consider baseband-equivalent signal representation only. The transmitted CDMA signal for one user can then be expressed as

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$$s(t) = \sum_{k=-\infty}^{\infty} a_{\lfloor \frac{k}{N} \rfloor} d_{k \bmod N_s} g(t - kT_c), \quad (1)$$

where N_s is the length of the scrambling sequence and $g(t)$ is the transmit pulse, for the 3GPP draft a root-raised cosine pulse with rolloff factor 0.22. We will assume a wide-sense stationary multipath fading channel model with N_p uncorrelated scatterers (paths):

$$h(\tau;t) = \sum_{l=0}^{N_p-1} c^{(l)}(t)\,\delta(\tau - \tau^{(l)}) \quad (2)$$

with $c^{(l)}$ being the l-th complex-valued path tap and $\tau^{(l)}$ the respective path delay. The received signal is pulse-matched filtered with the receive filter $g^*(-t)$ and is given by

$$z(t) = \sum_{l=0}^{N_p-1} c^{(l)}(t) \sum_{k=-\infty}^{\infty} a_{\lfloor \frac{k}{N} \rfloor} d_{k \bmod N_s} R_g(t - kT_c - \tau^{(l)}) + \tilde{n}(t) \quad (3)$$

$\tilde{n}(t)$ is complex valued coloured noise and includes additive white gaussian channel noise as well as other-user interference. $R_g(t)$ is the pulse filter autocorrelation function:

$$R_g(t) = \int_{-\infty}^{\infty} g^*(\tau) g(t+\tau)\, d\tau \quad (4)$$

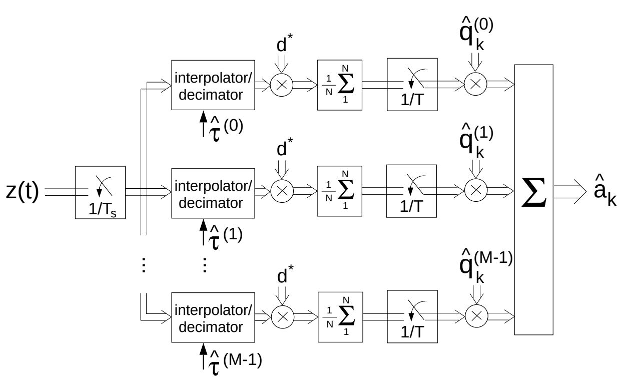

Figure 1. RAKE receiver model

The RAKE receiver is depicted in Figure 1. The received signal is first sampled at the sampling rate $1/T_s$, which must be at least equal to $2/T_c$ to guarantee sufficient statistics for detection and synchronization [3], and subsequently fed to all M RAKE fingers. In each finger l, an interpolation/decimation unit provides a data stream on chiprate $1/T_c$ which is sampled at the estimated timing instant $\hat{\tau}^{(l)}$. The latter is provided by the code tracking loop, one for each finger.

After correlation with the complex conjugate of the spreading sequence and summation over one symbol interval, symbol estimates are available in each finger:

$$\begin{aligned} r_k^{(l)} &= \frac{1}{N} \sum_{n=kN}^{(k+1)N-1} z(nT_c + \hat{\tau}^{(l)})\, d^*_{n \bmod N_s} + \tilde{\tilde{n}}_k^{(l)} \\ &= a_k c_k^{(l)} + \sum_{n=kN}^{(k+1)N-1} \sum_{\substack{p=0 \\ p \neq l}}^{N_p-1} c_k^{(p)} \sum_{m=-\infty}^{\infty} a_{\lfloor \frac{m}{N} \rfloor} d_{m \bmod N_s} \\ &\quad \cdot d^*_{n \bmod N_s} R_g\left((n-m)T_c - \tau^{(p)} + \hat{\tau}^{(l)}\right) + \tilde{\tilde{n}}_k^{(l)} \end{aligned} \quad (5)$$

The second equation is justified if the timing estimates are accurate, i.e. $\tau^{(l)} = \hat{\tau}^{(l)}$. Furthermore, the channel phasors are assumed to be essentially constant for the duration of one symbol. The signal component in equation (5) thus consist of two terms: one desired term which depends only on the multipath component l, and an interference term which results from the other channel paths. If the assumption is made that the unwanted components are negligibly small, e.g. due to very good crosscorrelation properties of the scrambling sequences, the correlator output reduces to

$$r_k^{(l)} = a_k c_k^{(l)} + \tilde{\tilde{n}}_k^{(l)} \quad (6)$$

Those estimates are then weighted according to the maximum-ratio criterion, where each weighting coefficient is given by

$$\hat{q}_k^{(l)} = \frac{c_k^{(l)*}}{\sum_{n=0}^{N_p-1} |c_k^{(n)}|^2} \quad (7)$$

2.2 System with antenna diversity

In this paper we are concerned with a system which employs space-time transmit diversity (STTD) as described in [1]. There, the information symbols a_k are space-time coded with a code of length 2 prior to transmission over two antennas. The coding scheme is depicted in Figure 2.

Taking into account the assumptions made for equation (6), two consecutively received symbols in RAKE finger l can then be described by

$$\begin{aligned} \mathbf{r}_k^{(l)} &= \mathbf{H}^{(l)} \cdot \mathbf{a}_k + \mathbf{n_k} \quad (8) \\ &= \begin{pmatrix} c^{(l,0)} & -c^{(l,1)} \\ c^{(l,1)*} & c^{(l,0)*} \end{pmatrix} \cdot \begin{pmatrix} a_{2k} \\ a^*_{2k+1} \end{pmatrix} + \begin{pmatrix} n_{2k} \\ n_{2k+1} \end{pmatrix} \end{aligned}$$

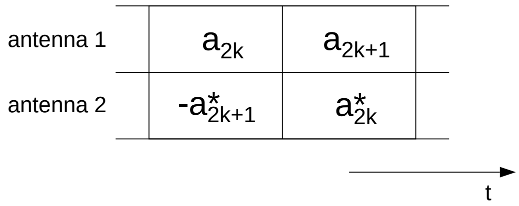

Figure 2. Space-Time Coding Scheme

Here, $c^{(l,0)}$ and $c^{(l,1)}$ denote the channel phasors for channels 1 and 2 in finger l which are assumed to be constant over two symbol intervals, and a_{2k} and a_{2k+1} denote two consecutively transmitted symbols which are space-time coded and transmitted over both antennas. For the channel phasors, the time index k has been dropped for notational convenience. The two channels are assumed to have equal power delay profiles but uncorrelated fading.

With the channel matrix $\mathbf{H}^{(l)}$ being orthogonal, a simple linear operation leads to the desired maximum-ratio combining (MRC) of the two diversity branches:

$$\begin{aligned} \tilde{\mathbf{r}}_{\mathbf{k}}^{(\mathbf{l})} &= \mathbf{H}^{(l)^H} \cdot \mathbf{r}_k^{(l)} \\ &= \rho \cdot \mathbf{I} \cdot \mathbf{a}_k + \mathbf{n}_{\mathbf{k}}, \end{aligned} \quad (9)$$

where $\rho = |c^{(l,0)}|^2 + |c^{(l,1)}|^2$ and $\mathbf{I}$ is the 2x2 identity matrix. The superscript ^{H} denotes complex conjugate transpose. The joint estimation of $c^{(l,0)}$ and $c^{(l,1)}$ is now addressed in the following section.

3 Wiener Channel Interpolation

It is shown in [3] that the optimal channel estimator for known pilots and channel statistics can be generated by postprocessing maximum-likelihood channel estimates with a Wiener filter, i.e. $\hat{\mathbf{c}} = \mathbf{w}^H \cdot \hat{\mathbf{c}}_{ML}$, where the latter are generated by complex-conjugately multiplying the observation with known symbols, i.e. $\hat{\mathbf{c}}_{ML} = a_k^* \cdot \mathbf{r}_n$. The Wiener filter coefficients depend on the mobile velocity and on the Signal to Noise ratio. In contrast to [6], however, we will make use of the common pilot channel defined in [4], where for one transmission link, i.e. without any TX or RX diversity, pilot symbols are transmitted continuously on a fixed rate. Phasor estimates on higher rates can be generated by linear interpolation and used for MRC in the RAKE in a feedforward fashion, making a delay of the data stream necessary.

For the transmit diversity mode defined in [4], however, pilot symbols are transmitted over both antennas according to the scheme depicted in Figure 3.

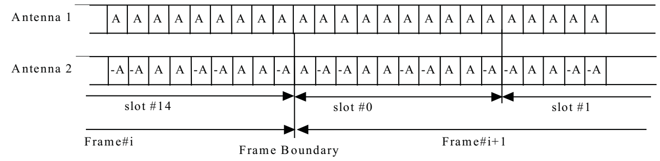

Figure 3. Pilot scheme for STTD

It is seen that the received effective pilot symbols are proportional to the sum and difference of the actual channel phasors in an alternating way (neglecting multipath interference):

$$\begin{aligned} r_\Sigma^{(l)} &= A \cdot (c^{(l,0)} + c^{(l,1)}) = A \cdot c^\Sigma \\ r_\Delta^{(l)} &= A \cdot (c^{(l,0)} - c^{(l,1)}) = A \cdot c^\Delta \end{aligned} \quad (10)$$

$A = 1 + j$ denotes the QPSK pilot symbol. With the channel phasors $c^{(l,0)}$ and $c^{(l,1)}$ assumed to be independent complex gaussian variables, the sum and difference channel phasors will also be independently gaussian and thus can be estimated using Wiener interpolation of ML estimates. From there, the original phasors can be reconstructed by addition and subtraction of c^Σ and c^Δ.

With the pilot scheme from Figure 3, 4 sets of Wiener filters need to be used for the estimation of c^Σ and c^Δ. They are shown in Figure 5 for a filter symbol memory $N_{Wiener} = 24$.

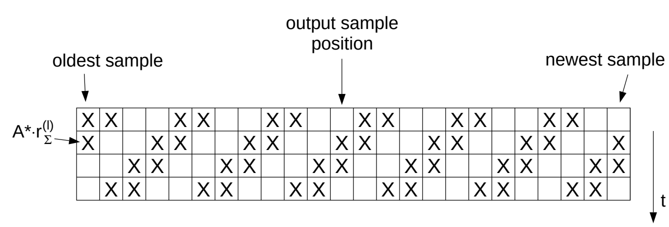

Figure 5. Layout of Wiener coefficients

Only half of the incoming ML estimates are used in the Wiener filtering operation per output sample for each of the channels. With the pilot pattern being symmetric in c^Σ and c^Δ, both Wiener filters use the same coefficient sets. A control unit must switch between the respective Wiener filter sets for each FIR output, reflecting the current position of the $r_\Sigma^{(l)}$ (or $r_\Delta^{(l)}$) in the data stream. An equivalent transmission model for the pilot symbols is depicted in Figure 4 for multipath l.

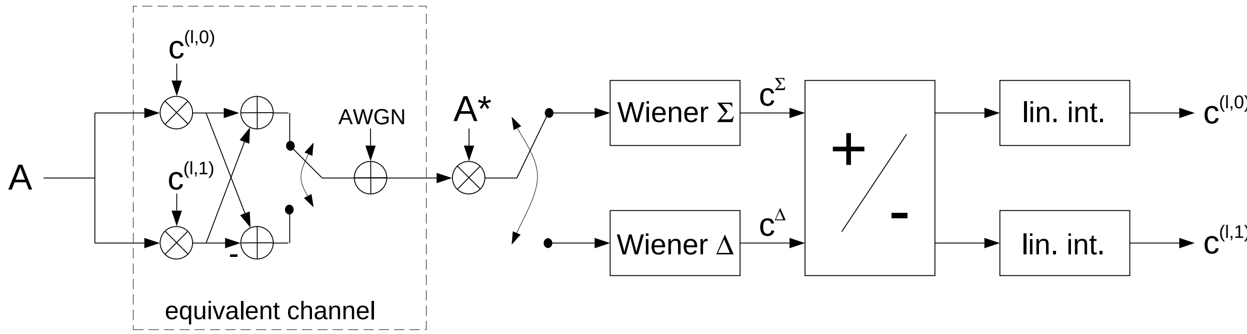

Figure 4. Equivalent transmission model for multipath l

It is noted that in order to account for the irregular pilot pattern at the frame boundaries (Figure 3), additional Wiener filter sets have to be used. They can however be precomputed and stored in memory. A complete transition over a frame boundary is shown in Figure 6 for $N_{Wiener} = 16$.

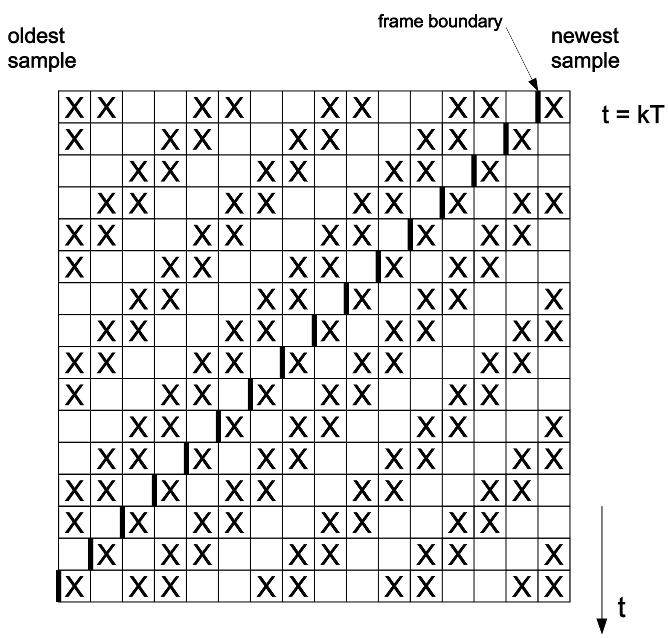

Figure 6. Layout of Wiener coefficients at frame boundaries

4 Simulation Results and Conclusion

Figures 7 through 9 show simulation results with a 2-tap Rayleigh fading channel model with tap powers of 0 and -10 dB and a tap spacing of $3.74 \cdot T_c$, for mobile velocities of 10, 120 and 500 km/h. The solid curves represent perfect channel state information (CSI), whereas the dotted curves were produced using Wiener channel estimation. In Figures 7 and 8, the upper dotted curves represent Wiener filters designed for a speed of 500 km/h, whereas for the lower dotted curves the Wiener filters were designed for the actual mobile speed (10 km/h and 120 km/h, respectively). For the curves on the right side (no STTD), a Wiener filter of length 16 was assumed. The results with STTD were generated with $N_{Wiener} = 16$. The complexity of the estimation algorithms with and without STTD are thus comparable: in the former case, only half as many effective Wiener filter taps are used per channel. However, linear interpolation between Wiener filter outputs must be performed twice as often as in the latter case.

It is seen that on one hand, using STTD yields a significant SNR gain, which is however dependent on the target BER. On the other hand, the synchronization losses due to imperfect CSI are small with a properly designed estimator and only slightly larger than those for the case without STTD, i.e. when only one channel needs to be estimated. Furthermore, the phasor estimation is not very sensitive to the design speed of the Wiener coefficients, expressed for example by the small loss in Figure 7 when the maximum speed of 500 km/h is assumed in the Wiener filter design but the mobile speed is only 10 km/h. This allows for the use of only one Wiener coefficient set, designed for 500 km/h, reducing memory requirement. A significant dependence on the SNR could not be verified.

It can thus be stated that in order to effectively utilize the SNR advantage of transmit diversity schemes such as STTD, accurate channel phasor estimates are required in the receiver. One method of estimating channel phasors in a transmit diversity scenario for a DS-CDMA system was presented in this paper. The synchronization losses due to imperfect channel knowledge were shown to be in the same order of magnitude as for a scheme without diversity, at comparable computational complexity. Furthermore, the quality of the estimates and thus the overall BER performance was shown to be robust with respect to the Wiener design parameters.

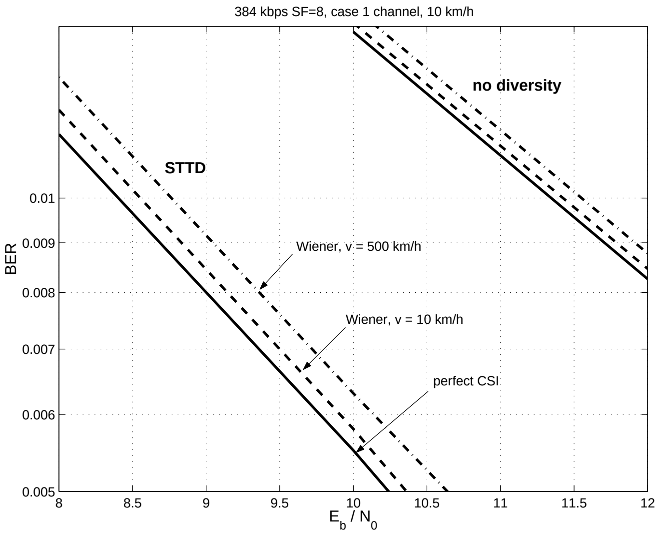

Figure 7. BER, STTD vs. no diversity, 10 km/h

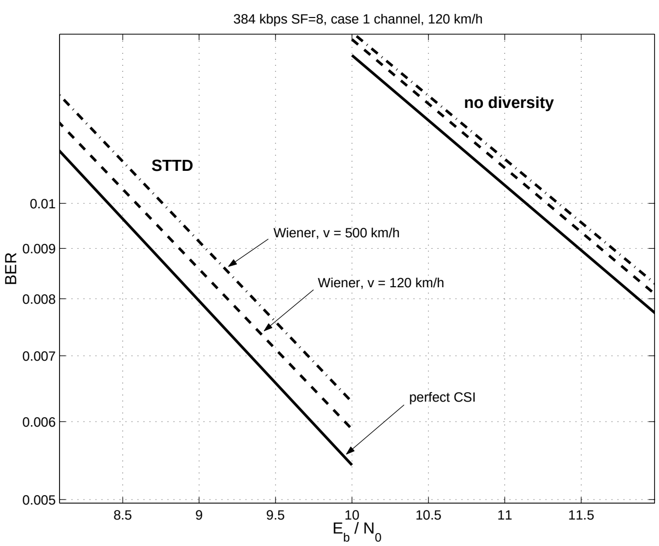

Figure 8. BER, STTD vs. no diversity, 120 km/h

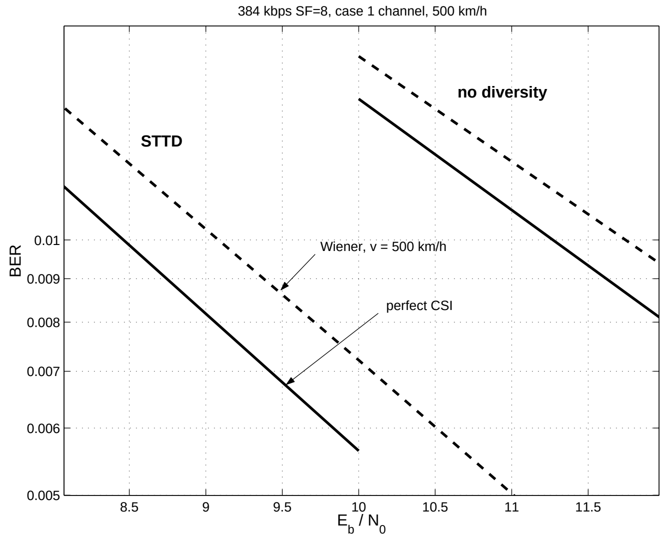

Figure 9. BER, STTD vs. no diversity, 500 km/h

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