Scrambling–Exact Channel Estimation in W–CDMA Systems

An Algorithm Beyond Correlation

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***Abstract*— Multiple–input Multiple–output (MIMO) antenna systems require one pilot signal for each transmitting antenna. As scrambling is a quasi–orthogonal operation a small portion of the pilot energy is scattered into other pilot signals even with ideal conditions apart from that small non–orthogonality. In this paper the correlation approach of estimating the channel impulse response is extended analytically such as to take into account the full spreading sequence, improving thereby significantly the performance of high–speed downlink packet access (HSDPA) MIMO systems especially for very high data rates when dual stream transmission is used in combination with cancellation of the interference between the data streams.**

Keywords-W-CDMA; HSDPA; MIMO; channel estimation; scrambling

I. Introduction

Channel estimation plays an important role in wideband code division multiple access (W–CDMA) networks, especially for very high data rates as they are achieved with MIMO [1]. It has therefore been topic of different investigations which go beyond the conventional correlation approach. In [2], a novel type of linear minimum mean square equalizer (LMMSE) was presented which is able to properly take into account all types of interference without being too complex for implementation in real systems, whereas [6,7] investigate smoothening the primary channel estimates by appropriate filtering and cancelling the inter–antenna interference, respectively.

In MIMO systems it is necessary to measure all physical channels from each transmitting to each receiving antenna. Therefore, each transmitting antenna has to be fed with a separate pilot sequence. For HSDPA systems [5], there are defined primary and secondary common pilot channels which are orthogonal either by using orthogonal patterns or by applying orthogonal spreading codes. In any case, the coded pilot sequences are scrambled.

As scrambling is a quasi–orthogonal but not an exact orthogonal operation the channel information derived by the receiver remains distorted even under conditions being ideal beyond these distortions, i.e., absence of receiver noise and additive white Gaussian noise (AWGN) channels. If only one pilot channel is present, the effect is usually negligible: Although a certain amount of energy is scattered by the spreading operation into other codes the general shape persists as the energy reduction is more or less proportional to the available energy. If there are however two or more pilot sequences the relative error strongly increases as the scattering of energy from other codes also happens at time (slot) positions where the respective pilot signal is weak or even vanishes.

If MIMO is configured for more than one data stream, these data streams will interfere with each other with strength depending on the correlation of the different physical channels. But this interference can effectively be reduced by interference cancellation as all data streams have the same origin and therefore are subject of the same channel conditions. Furthermore, the receiver has to decode all data streams anyway so that the effort of cancellation is limited. As interference cancellations is re–applying the channel transformation to the already detected data, any error in the estimated channel information takes effect three times, namely at first equalization, at re–application of the channel operation in the course of the interference cancellation and at re–equalization of the equivalent single stream data. Any error in channel estimation therefore acts non–linearly on the overall performance of the system.

The scrambling distortion is mainly significant for very high data rates where the respective UE is served with all available resources. This means that intra–cell interference (originating from other users) is small. Furthermore, these mentioned high data rates are achievable under quite good radio conditions only, i.e., also the inter–cell interference must be small. Therefore, a correlation–based estimator might be sufficient in that range of interest but the impact of scrambling induced distortion should be removed.

As the receiver knows the scrambled pilot symbol patterns as they are fed to the transmitting antennas it is possible to calculate the distortion induced by scrambling analytically as a function of the channel impulse response. Inverting this function allows to extend the correlation algorithm such as to take into account the distortion, leading to an estimated channel impulse response being exact with respect to scrambling effects. Although a matrix operation is

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required to solve the resulting equation system the corresponding matrix elements depend only on the scrambling vector and can hence be calculated once at the initial phase. As long as the scrambling code is not changed these elements remain stable.

The remainder of this paper is organized as follows: In Section II we introduce the signal model with the basic equations. They are required to describe the data transfer through the channel and will be used in Section III for formulating the analytical solution for scrambling–exact channel estimation. Section IV then demonstrates the gain of the proposed algorithm with some results from link level simulation. Finally, we draw our conclusion in Section V.

II. BASIC CHANNEL PILOT EQUATIONS

In the following, we assume that two transmitting antennas are present, each fed with a separate pilot sequence, namely $s_1(n)$ and $s_2(n)$ with n counting the oversampled time steps. The signal $r(l)$ at oversampled chip position l is for a single antenna then given by (see e.g., [4], Section 1.3):

$$r(l)=\sum_{k=0}^{N_{\mathrm{OSF}}\cdot M} c\left(\frac{l-k}{N_{\mathrm{OSF}}}\right)\left[h_1(k)s_1\left(\frac{l-k}{N_{\mathrm{OSF}}}\right)+h_2(k)s_2\left(\frac{l-k}{N_{\mathrm{OSF}}}\right)\right]. \tag{1}$$

With the conventional correlation approach of channel estimation, the received signal is de–spread ignoring the error made by this quasi–orthogonal operation (see, e.g., [1]):

$$\begin{aligned}&\sum_{l=l_0+1}^{N_{\mathrm{SF}}} c^*\left(\frac{l}{N_{\mathrm{OSF}}}\right)r(l+d)=\\ &=\sum_{l=l_0+1}^{N_{\mathrm{SF}}}\sum_{k=0}^{N_{\mathrm{OSF}}\cdot M} c^*\left(\frac{l}{N_{\mathrm{OSF}}}\right)c\left(\frac{l-k+d}{N_{\mathrm{OSF}}}\right)\cdot\\ &\cdot\left[h_1(k)s_1\left(\frac{l-k+d}{N_{\mathrm{OSF}}}\right)+h_2(k)s_2\left(\frac{l-k+d}{N_{\mathrm{OSF}}}\right)\right]\approx\\ &\approx h_1(d)s_1(l_0)+h_2(d)s_2(l_0)\end{aligned} \tag{2}$$

In (2), we introduced a start index l_0 for summation over chips in order to cope with handling of more than one symbol. For the first symbol, $l_0=0$, for the second one $l_0=N_{SF}$, and so on. If we now define

$$\begin{aligned}&\kappa_{l_0,j}(d,k)\equiv\kappa_{l_0,j}(d-k)=\\ &\quad\sum_{l=l_0+1}^{N_{\mathrm{OSF}}\cdot N_{\mathrm{SF}}} c^*\left(\frac{l}{N_{\mathrm{OSF}}}\right)c\left(\frac{l-k+d}{N_{\mathrm{OSF}}}\right)s_j\left(\frac{l-k+d}{N_{\mathrm{OSF}}}\right)\\ &\kappa_{l_0,j}(d-k)=0\quad\forall\ d-k\notin Z\\ &\rho_{l_0}(d)=\sum_{l=l_0+1}^{N_{\mathrm{SF}}} c^*\left(\frac{l}{N_{\mathrm{OSF}}}\right)r(l+d)\end{aligned} \tag{3}$$

eq. (2) can be written as

TABLE 1: NAMING CONVENTIONS FOR IMPORTANT VARIABLES

Received signal (single antenna)	$r(l)$
Sent signal (index i for Tx antenna)	$s_i(l)$
Channel impulse response (index i for Tx antenna)	$h_i(l)$
Scrambling code element at integer chip index n	$c(n)$
Channel length	$M+1$
Spreading sequence length	N_{SF}
Oversampling Factor	N_{OSF}
Number of symbols required for CPICH orthogonality	N_{ortho}
Number of symbols required for CPICH orthogonality, measured in chips	$N=\frac{\langle\text{chips per slot}\rangle}{N_{\mathrm{ortho}}}$

$$\rho_{l_0}(d)=\sum_{k=0}^{N_{\mathrm{OSF}}\cdot M}\left[h_1(k)\kappa_{l_0,1}(d-k)+h_2(k)\kappa_{l_0,2}(d-k)\right]. \tag{4}$$

We then are able to split the summation into two parts, namely one counting the chips and the other stepping through the oversampling within each chip. With the definitions

$$\begin{aligned}&d=N_{\mathrm{OSF}}\left\lfloor\frac{d}{N_{\mathrm{OSF}}}\right\rfloor+d\bmod N_{\mathrm{OSF}}=\hat{d}+\breve{d}\\ &\hat{d}=N_{\mathrm{OSF}}\left\lfloor\frac{d}{N_{\mathrm{OSF}}}\right\rfloor\\ &\breve{d}=d\bmod N_{\mathrm{OSF}}\geq 0\end{aligned} \tag{5}$$

and assuming that l is an integer multiple of N_{OSF} (generally spoken, oversampling can be accounted for in d) we can write finally

$$\begin{aligned}&\rho_{l_0}\left(\hat{d}+\breve{d}\right)=\\ &=\sum_{k=0}^{M}\sum_{j=0}^{N_{\mathrm{OSF}}-1}\left[h_1(N_{\mathrm{OSF}}k+j)\kappa_{l_0,1}\left(\hat{d}+\breve{d}-N_{\mathrm{OSF}}k-j\right)+\right.\\ &\qquad\left.h_2(N_{\mathrm{OSF}}k+j)\kappa_{l_0,2}\left(\hat{d}+\breve{d}-N_{\mathrm{OSF}}k-j\right)\right]=\\ &=\sum_{k=0}^{M}\left[h_1\left(N_{\mathrm{OSF}}k+\breve{d}\right)\kappa_{l_0,1}\left(\hat{d}-N_{\mathrm{OSF}}k\right)+\right.\\ &\qquad\left.h_2\left(N_{\mathrm{OSF}}k+\breve{d}\right)\kappa_{l_0,2}\left(\hat{d}-N_{\mathrm{OSF}}k\right)\right]\end{aligned} \tag{6}$$

In (6), we have separated integer and oversampled parts on the right hand side of (4) by splitting

$$k=N_{\mathrm{OSF}}\left\lfloor\frac{k}{N_{\mathrm{OSF}}}\right\rfloor+j\rightarrow N_{\mathrm{OSF}}k+j\,. \tag{7}$$

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III. SCRAMBLING–EXACT CHANNEL ESTIMATION

In [5], there are defined different configurations for the common pilot channel (CPICH) in W–CDMA systems. For the usage in combination with MIMO, two of them are relevant, namely either using only the primary CPICH spread with code c_{SF1} of length 256 and orthogonal bit sequences on the two transmitting antennas, or using the primary CPICH on the first and a secondary CPICH with code c_{SF2} on the second transmitting antenna. The secondary CPICH than must use the same bit sequence as the primary CPICH on antenna 1. The same scrambling code for both cases is used in any case.

If two adjacent symbols are merged both options can formally be described by a virtual spreading with length of 512 where the effective spreading codes can be defined in the first case as [c_{SF1} c_{SF1}] and [c_{SF1} inv(c_{SF1})], and in the second case as [c_{SF1} c_{SF1}] and [c_{SF2} c_{SF2}]. As long as the channel estimation algorithm finally uses any filtering procedure over an even number of pilot symbols, both options are equivalent with respect to the influence of scrambling. We will restrict therefore the following investigations to the configuration with primary CPICH only and orthogonal bit sequences on the two transmitting antennas.

A. Single Symbol Channel Estimation

If we want to calculate $M+1$ coefficients for both channels, we need $2\cdot(M+1)$ equations, i.e.,

$$\begin{pmatrix} \rho_{l_0}\left(N_{OSF}\cdot 0+\breve{d}\right) \\ \vdots \\ \rho_{l_0}\left(N_{OSF}\cdot M+\breve{d}\right) \\ \vdots \\ \rho_{l_0}\left(N_{OSF}\cdot(2M+1)+\breve{d}\right) \end{pmatrix} = K_{l_0} \begin{pmatrix} h_1\left(N_{OSF}\cdot 0+\breve{d}\right) \\ h_2\left(N_{OSF}\cdot 0+\breve{d}\right) \\ \vdots \\ h_1\left(N_{OSF}\cdot M+\breve{d}\right) \\ h_2\left(N_{OSF}\cdot M+\breve{d}\right) \end{pmatrix}, \quad (8)$$

where the pilot scrambling matrix $\mathbf{K}_{l_0}$ is defined as

$$\mathbf{K}_{l_0} = \begin{pmatrix} \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) & \cdots & \kappa_{l_0,1}(-M) & \kappa_{l_0,2}(-M) \\ \vdots & \ddots & & \iddots & \vdots \\ \kappa_{l_0,1}(M) & \kappa_{l_0,2}(M) & \cdots & \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) \\ \vdots & \iddots & & \ddots & \vdots \\ \kappa_{l_0,1}(2M+1) & \kappa_{l_0,2}(2M+1) & \cdots & \kappa_{l_0,1}(M+1) & \kappa_{l_0,2}(M+1) \end{pmatrix}. \quad (9)$$

If the scrambling sequences would be completely orthogonal, the pilot scrambling matrix (9) would contain non–zero elements ($\kappa_j(0)$) in the first M rows only, i.e., rows $M+1\ldots 2M+1$, wouldn't contribute at all to the solution and **K** wouldn't be invertible. The accuracy of these equations is therefore small even if the non–orthogonality of spreading is taken into account. To overcome this limitation one has therefore either to average over at least as many symbols as there are pilot sequences (see Section B), or the approach is extended such as to obtain equations of each pilot symbol (see Section C). This becomes obvious when looking to the pilot scrambling matrix for a single–tap channel including root raised cosine filtering as displayed in Fig. 1.

The channel coefficients (at sampling positions) are finally given by

$$\begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix} = \mathrm{K}^{-1}_{\mathrm{single},l_0} \begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \vdots \\ \rho_{l_0}(2M+1) \end{pmatrix}. \quad (10)$$

The upper half of **K** shows large real entries on the main diagonal, whereas the other elements are dominated by the scrambling correlation coefficients. As expected, the calculation of the channel impulse response fails at least partly even if afterwards an averaging over all symbols of a frame is performed. This is shown in Fig. 2 for an AWGN channel with raised cosine filter.

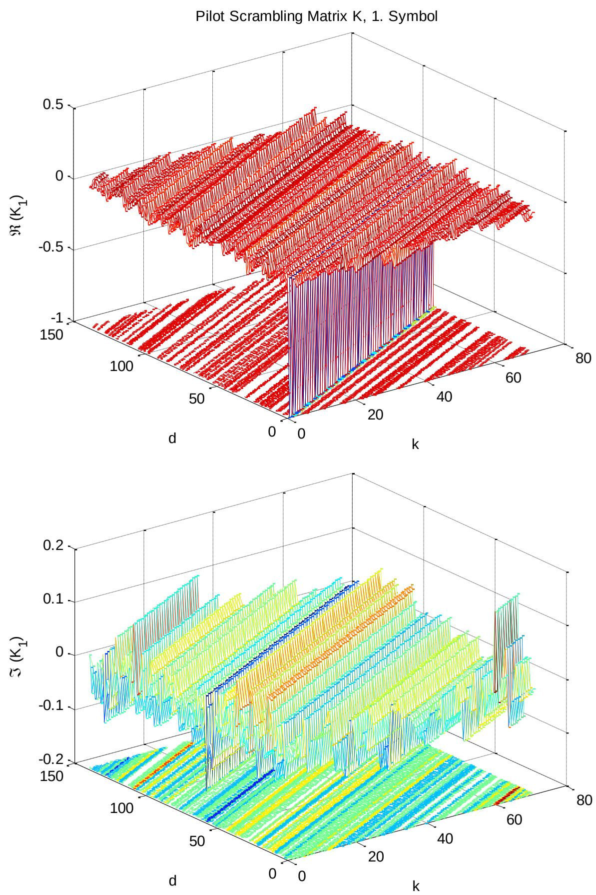

Figure 1. Pilot scrambling matrix for a single tap channel including raised cosine filtering.

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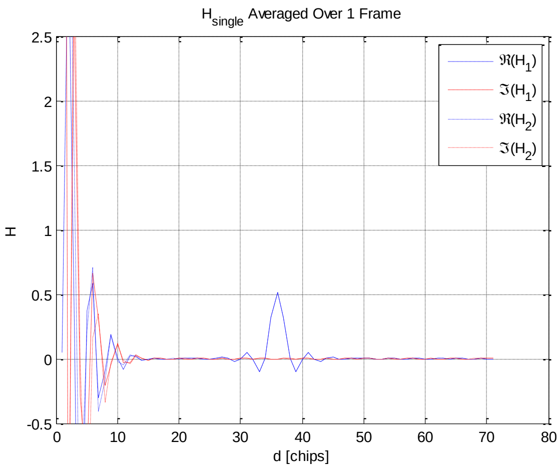

Figure 2. Channel Impulse Response averaged over all symbols of a frame.

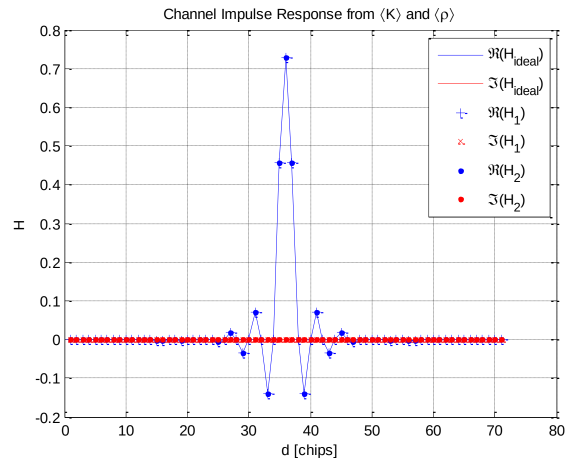

Figure 3. Channel Impulse Response calculated with averaged K.

B. Averaging Pilot Scrambling Matrix

The inaccuracy of the channel impulse response calculated for a single symbol is caused by $\mathbf{K}$ being badly conditioned. This can be improved by averaging both $\mathbf{K} \rightarrow \langle \mathbf{K} \rangle_{frame}$ and $\rho \rightarrow \langle \rho \rangle_{\text{frame}}$. Although the general shape of the pilot scrambling matrix is preserved by this operation the conditioning is now about 10 orders of magnitude better and $\langle \mathbf{K} \rangle_{frame}$ becomes invertible. For an AWGN channel with raised cosine filter it provides a perfectly estimated channel impulse response, see Fig. 3.

C. Twin Symbol Channel Estimation

The averaging procedure discussed in Section B assumes that K and ρ are statistically independent and the averaging therefore factorizes and can be executed independently. Inversion is then executed with the averaged pilot scrambling matrix. The shown result implies that this assumption is correct at least for a single tap channel including root raised cosine filter.

There is, however, an alternative approach which doesn't need the above restriction: Both pilot scrambling matrix and de–spread received signal vector are constructed using two symbols with different pilot sequences,

$$\begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \rho_{l_1}(0) \\ \vdots \\ \rho_{l_1}(M) \end{pmatrix} = \begin{pmatrix} \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) & \cdots & \kappa_{l_0,1}(-M) & \kappa_{l_0,2}(-M) \\ \vdots & \ddots & & \iddots & \vdots \\ \kappa_{l_0,1}(M) & \kappa_{l_0,2}(M) & \cdots & \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) \\ \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) & \cdots & \kappa_{l_1,1}(-M) & \kappa_{l_1,2}(-M) \\ \vdots & \iddots & & \ddots & \vdots \\ \kappa_{l_1,1}(M) & \kappa_{l_1,2}(M) & \cdots & \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) \end{pmatrix} \begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix}, \quad (11)$$

where the pilot scrambling matrix $\mathbf{K}_{\text{twin}}$ now is defined as

$$\mathrm{K}_{\text{twin},l_{0,1}} = \begin{pmatrix} \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) & \cdots & \kappa_{l_0,1}(-M) & \kappa_{l_0,2}(-M) \\ \vdots & \ddots & & \iddots & \vdots \\ \kappa_{l_0,1}(M) & \kappa_{l_0,2}(M) & \cdots & \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) \\ \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) & \cdots & \kappa_{l_1,1}(-M) & \kappa_{l_1,2}(-M) \\ \vdots & \iddots & & \ddots & \vdots \\ \kappa_{l_1,1}(M) & \kappa_{l_1,2}(M) & \cdots & \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) \end{pmatrix}. \quad (12)$$

In the following, these symbols will be called *twin symbols* if the two symbols are adjacent. The channel coefficients are then given by

$$\begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix} = \mathrm{K}_{\text{twin},l_{0,1}}^{-1} \begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \rho_{l_1}(0) \\ \vdots \\ \rho_{l_1}(M) \end{pmatrix}. \quad (13)$$

Now, all rows and columns of the pilot scrambling matrix contain (at least) one main element as displayed in Fig. 4.

The real part shows two bars, one completely in the lower half volume, the other in both half volumes. Whereas the former belongs to equal pilot symbols for both channels, the latter results from inverse symbols.

The AWGN channel impulse response including raised cosine filter calculated from the twin pilot correlation matrix again matches perfectly with the ideal channel impulse response, not only when averaged over all symbols of a frame (see Fig. 5), but also for each twin symbol. This means that no averaging is required in the absence of noise.

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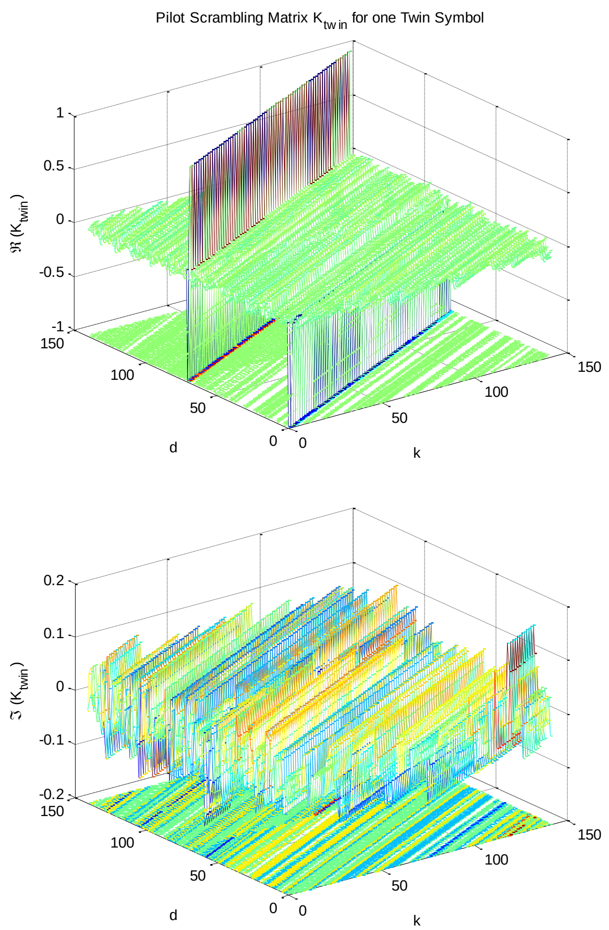

Figure 4. Pilot scrambling matrix for one twin symbol.

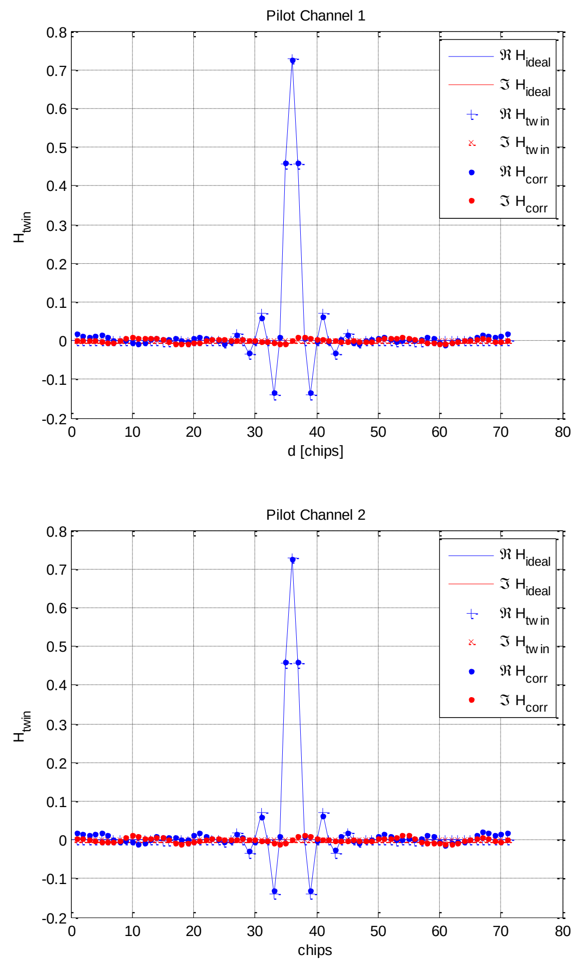

Figure 5. Twin Channel Impulse Response averaged over all symbols, compared with conventional correlation result and ideal case.

IV. Link Level Simulations

In order to demonstrate the benefit of the scrambling–exact channel estimation we executed link level simulations for a single UE moving with 3km/h and assuming a channel of type Pedestrian A. Two data streams were transmitted in MIMO mode and for both the modulation and coding scheme corresponding to entry 10 in the CQI mapping table K [6] was used. The transmitting antennas were assumed to be mounted crosswise diagonal (X), the receiving antennas crosswise perpendicular and horizontal (+) and crosstalk between the polarization directions was allowed. In Fig. 6, the block error rate (BLER) is displayed for each data stream as a function of SNR. The SNRs for a working point at 10% BLER are given in the legend for the first (solid line) and second (dashed line) data stream.

As expected, there is an appreciable gain of performance notable by a shift of the working point of up to 2dB for the second data stream which mainly benefits from interference cancellation, but even the working point of the first data stream is improved by 0.5dB.

V. Conclusion

The usual procedure for estimating the radio channels discards the non–orthogonality of the spreading sequence which causes an error of several percents in case of more than one pilot sequence is required as is valid for using spatial diversity and MIMO systems. It has to be emphasized that this error is present even in case of pilot channels being fully separated, e.g., by cross polarized wave fronts with each front bearing exactly one pilot and receiving antennas oriented along the polarization directions, as long as this fact is not known a priori on receiving side: The estimator has to assume that signals of both pilots are present at each receiving antenna and hence misinterprets fractions scattered from the ‘real’ pilot by scrambling as contributions of the other pilot. Merely a single pilot being present combined with the knowledge about this fact on receiving side avoids this complication.

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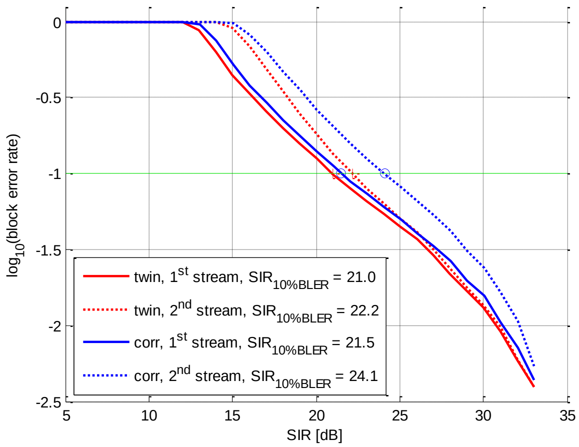

Figure 6. Link level simulation results with simple correlation (upper) and scrambling–exact (lower) channel estimation.

Due to the importance of channel estimation for reaching high data rates, there are published many proposals for improving the basic correlation approach, e.g., by filtering the primary channel estimates [3], cancelling inter‒antenna interference [4] or extending the RAKE correlation approach to a LMMSE one [2] and thus taking into account all kinds of interference. All of these suggestions can basically be extended by scrambling‒exact dealing of pilot patterns as proposed in this paper even if we have investigated here only the simple RAKE correlation approach, including appropriate filtering as described in [3], in the simulator.

Although the spreading induced error is negligible in many cases, high data rates in combination with MIMO and transmission of two data streams require a higher accuracy. This can be reached by taking into account the spreading sequence exactly in the channel estimation algorithm. As in the correlation case, two symbols (in case of two pilot signals) with different pilot code elements are required to resolve the channels from both transmitting antennas. Therefore, a twin symbol pilot scrambling matrix is introduced with its inverse providing the channel impulse response from the received data.

The numerical effort of this approach is for sure higher than the conventional correlation method. This effort however can at least partly be shifted to an initialization phase because the pilot scrambling matrix only depends on the scrambling sequence, the pilot patterns and the maximum length of the channel but not on any quantities varying with time. Mainly the second (weaker) data stream benefits from this improved channel estimation if cancellation of inter–stream interference is applied. In link level simulation it was shown that its working point as function of SNR can be improved by 2dB for 10% block error rate.

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