Impact of Scrambling-Exact Channel Estimation on HSDPA MIMO

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***Abstract*—Multiple-input Multiple-output (MIMO) antenna systems require a separate pilot signal for each transmitting antenna for enabling the receiver to measure all physical channels separately. As scrambling operates with pseudorandom sequences the auto-correlation is quasi-orthogonal with small deviations from exact orthogonality. In this paper the impact of these deviations is investigated, which is important especially for estimating the channel impulse response. Therefore, an analytical approach beyond simple correlation is proposed, taking into account the full scrambling sequence. Link level simulations show that the performance of high-speed downlink packet access (HSDPA) MIMO systems is significantly improved, 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; crosstalk; interference cancellation;**

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], [2]. It has therefore been topic of different investigations, which go beyond the conventional correlation approach. In [3], 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 [4] and [5] investigate smoothening the primary channel estimates by appropriate filtering and cancelling the inter-antenna interference, respectively. An adaptive channel estimator was proposed in [6] for system utilizing space time transmit diversity.

In order to benefit from diversity of transmitted signals achieved either by separating the transmitting antennas spatially or using different polarization directions, it is necessary to measure all physical channels from each transmitting to each receiving antenna, and transmission of more than one data stream in MIMO systems is not possible at all without resolving these channels individually (see, e.g., [7]). Therefore, each transmitting antenna has to be fed with a separate pilot sequence. For HSDPA systems [8], there are defined primary and secondary common pilot channels, which are orthogonal either by using orthogonal symbol patterns or by applying orthogonal spreading codes. In any case, the coded pilot chip level 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. In [9], an iterative approach was proposed to approximately handle this phenomenon. Because equalization of the incoming signal is cancelling the effect of multi-path propagation to a considerable extent this effect does not play a role for data detection itself. However, channel estimation, which is needed for measuring the different propagation paths has to work on un-equalized incoming data and thus is impacted significantly.

As long as only one pilot channel is present, the impact is usually negligible: Although a certain amount of energy is scattered by the scrambling operation into other codes or pilot patterns, the general shape of the signal at the output of the physical channel persists as the energy reduction is more or less proportional to the available energy, i.e., slots with low pilot signal strength also have a low absolute distortion caused by scrambling. If there are however two or more pilot sequences the relative error strongly increases as besides 'scattering out' also 'scattering in' has to be regarded: The intrusion of energy from other codes also happens at time (slot) positions where the respective 'target' pilot signal is weak or even vanishes, causing an error signal, which can be much higher than the 'pure' signal.

If MIMO is configured for more than one data stream these 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 additional effort of cancellation is limited. As interference cancellation 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 user equipment (UE) is served by the base station with all available resources. This means that intra-cell interference (originating from other users) is small, mainly caused by pilot and control channels, which are known to the UE and can be suppressed. 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 systematic 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 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, i.e., the UE does not change the cell, these elements remain stable.

The remainder of this paper is organized as follows: In Section II we give a brief description of MIMO inter-stream interference and its cancellation as this is the main application of the proposed improvement in channel estimation. In Section III we then introduce the signal model with the basic equations. They are required to describe the data transfer through the physical channels to be measured by the UE and will be used in Section IV for formulating the analytical solution for scrambling-exact channel estimation. Section V then demonstrates the gain of the proposed algorithm with some results from link level simulation, followed by a discussion of an efficient implementation of the algorithm in Section VI. After explaining why the deviation of scrambling from orthogonality has no impact on data detection itself in Section VII, we finally draw our conclusion in Section VIII.

II. Cancellation of MIMO Inter-Stream Interference

In W-CDMA HSDPA, the maximum possible data rate has been and will be increased over the years especially by using modulation schemes of higher order such as 16QAM or 64QAM and new modes of operation [10], [11]. The latter can be divided into two classes. The first one is extending the bandwidth by using two or even more carriers. In that case, the procedures on physical layer remain more or less untouched and are hence beyond our focus here. The second one is the introduction of MIMO with transferring several data streams in parallel using both the same physical resources as well as the same spreading codes

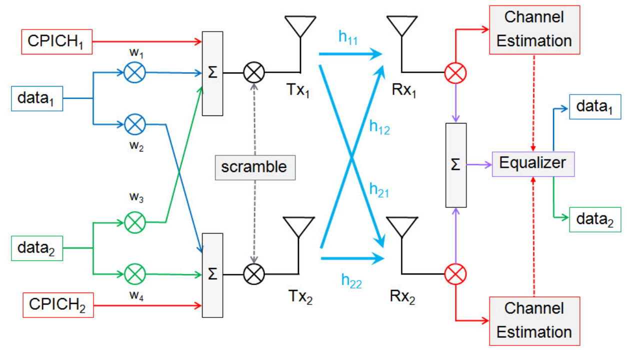


Figure 1. Schematic overview about dual stream transmission in W-CDMA HSDPA MIMO.

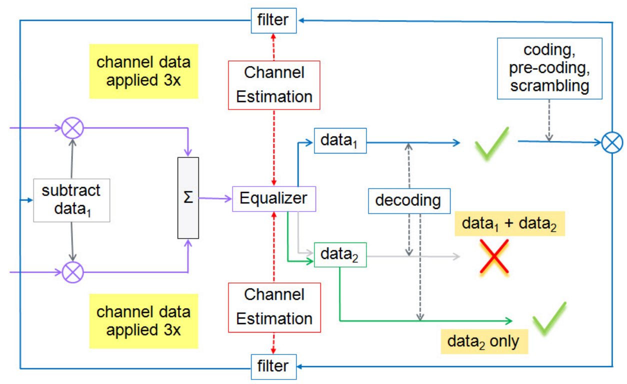


Figure 2. Schematic overview about cancellation of inter-stream interference.

[12]. Fig. 1 gives a schematic overview about MIMO dual stream transmission.

Good channel estimation plays a significant role for effective data transmission, especially if high data rates shall be reached. This is achieved by transmitting pilot patterns with defined power, which are used then by the receiver to estimate the single channel coefficients. Dual stream transmission requires at least two antennas on both transmitter and receiver side. To enable the receiver to measure each physical channel separately, two different pilot signals are sent. In HSDPA each one is fed to one of the transmitting antennas [13], whereas each data stream is distributed to both transmitting antennas with different phase factors, the pre-coding weights. A pre-coding codebook with four entries is defined and the UE signals the index of the pre-coding values, which yield the highest received power. The pre-coding vector for the second data stream is then taken orthogonal to the one of the first stream in order to allow the separation of the streams by the receiver. The received power of the second stream usually is significantly lower due to more destructive interference of the beams from the two transmitting antennas. For that reason, first and second stream often are refered as strong and weak one. Further details are not important for the discussion here but can be found in [12].

In conventional LMMSE equalizers as described, e.g., in [14] both pre-coding weights and channel coefficients are needed to determine proper weights for calculation of soft values of the received symbols, which then are passed to the decoder. Although these equalizer weights are non-linear in the channel impulse response the weight calculation is the only place in the decoding chain where the estimated channel coefficients are required.

The separation of the streams on receiver side is possible due to diversity of the physical channels between each transmitting and receiving antenna either spatially or due to different polarization properties of the affected antennas. If the physical channels are correlated this separation of the streams is feasible only to a certain extend and a substantial inter-stream interference remains after equalizations, which mainly effects the weaker of the two data streams. To keep this stream decodable anyway requires the application of a modulation and coding scheme with rather low transport block size.

If, however, the first data stream is decoded correctly, the complete information for re-constructing its shape as it reaches the receiver antennas is available at the receiver, including the coefficients of the physical channels. It is therefore possible to subtract this contribution from the overall incoming signal, leaving the stripped signal containing only the signal of the second stream, and of course noise and interference beyond inter-stream interference. This cancellation of inter-stream interference thus allows higher data rates also for the weaker data stream.

From the schematical display in Fig. 2 it becomes obvious that now the estimated channel coefficients are not applied only once but three times, namely for the first calculation of the equalizer weights, for the re-construction of the stronger signal as it reaches the antennas, and for the re-calculation of the weights used for equalizing the remaining weaker stream, which now requires single stream weights. As long as the two data streams are not orthogonal at the reciever side, which is usually the case if interference cancellation is executed, application of dual stream weights implicitly assigns within the equalizer equations a part of the stream power to the other data stream, which is, however, no more present. This both reduces the available signal over noise ratio and distorts the signal, i.e., the data detection suffers significantly [15].

The increased impact of channel estimation errors in presence of interference cancellation becomes obvious when looking to the equation for the received soft values $y_{2,\mathrm{IC}}$ of the weaker second data stream at the equalizer output (see, e.g., [16]).

$$y_{2,\mathrm{IC}} = w_{2,\mathrm{single}} \left[r - (1-\delta) \cdot h \cdot s_1 \right]. \tag{1}$$

To the re-spread and re-scrambled signal of the first data stream, s_1, the channel impulse response h is re-applied. Errors in channel estimation are accounted for by the correction δ. As the main part of the overall received signal r originates from the stronger first data stream, both terms in the bracket are of the same order of magnitude and $y_{2,\mathrm{IC}}$ is therefore quite sensitive to errors in the channel estimation with leading order of δ. The additional contribution of δ implicitly contained in $w_{2,\mathrm{single}}$ is present without interference cancellation, too.

Of course it would be much easier to use directly the stronger signal at the equalizer output instead of re-coding the decoded one, especially as this would reduce the detection delay. The benefit of such a procedure however is quite limited because such a signal still contains all noise and interference contributions and not only errors from channel estimation. It can even happen that the manipulated signal obtained this way is worse than without interference cancellation at all.

In Section V we will show that the performance of the weaker data stream experiences a huge gain from improved channel estimation, but also the stronger data stream as well as the weaker one in absence of interference cancellation benefit from this procedure.

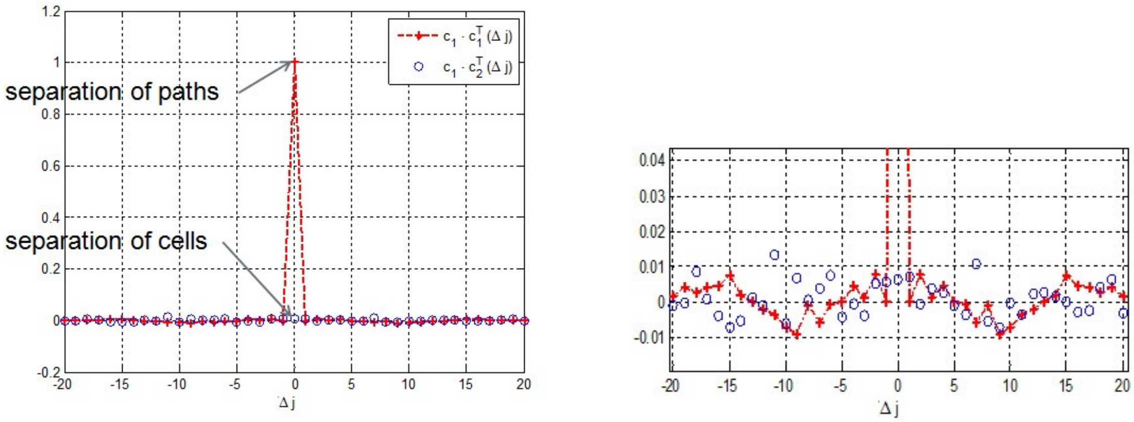
Table I
NAMING CONVENTIONS FOR IMPORTANT VARIABLES

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 of sequence j at integer chip index n	$c_j(n)$
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	$N = \frac{\text{chips per slot}}{N_{\mathrm{ortho}}}$

III. BASIC CHANNEL PILOT EQUATIONS

Throughout the rest of the paper, we use the naming conventions listed in Table I. We assume that two transmitting antennas are present, each fed with a separate pilot sequence, namely s_1 and s_2. On chip level, i.e., after spreading, the signals are then chip-wise multiplied with the elements of a scrambling vector $\mathbf{c}$, which has the length of a frame, i.e., 38400 chips [17]. The scrambling sequence is known also to the receiving UE, which can hence de-scramble the received signal r by multiplication with the hermitian scrambling vector, $\mathbf{c}^{\dagger}$.

The W-CDMA scrambling sequences are constructed from the pseudorandom complex-valued Gold sequences [17] and hence their autocorrelation is quasi-orthogonal (see,



(a) (b)

Figure 3. Scalar product of scrambling vector with itself (red) and with another (blue) scrambling vector as function of shifted index.

e.g., [18]. This operation preserves both amplitude and phase to a high extend if the applied scrambling vector matches the one used on the transmitter side without shift in the chip index as can be seen from the red curve at $j = 0$ in Fig. 3a. On the other hand, de-scrambling with another scrambling sequence (blue curve in Fig. 3a) or the same one but shifted in the chip index yields an correlation result close to zero. If really orthogonal instead of quasi-orthogonal codes would be used non-matching de-scrambling would provide exactly vanishing results. Scrambling is therefore a powerful means for both separation of signals from different cells, which use different scrambling codes, as well as for separation of contributions of the same base station but propagating via differently delayed paths, i.e., application of the same but time-shifted scrambling sequence.

Fig. 3b shows the error caused by the deviation of scrambling vectors from orthogonality in more detail. Although this error is in the range of a few percents only it can cause a degradation of system performance, especially if advanced receivers are used.

As we want to include oversampling and raised cosine filtering in our description, s and r are defined in sub-chip time steps l. For a single receiving antenna, the received signal is then given by (see, e.g., [18], Section 1.3):

$$r(l) = \sum_{k=0}^{N_{\text{OSF}}\cdot M} c\left(\left\lfloor \frac{l-k}{N_{\text{OSF}}} \right\rfloor\right) \left[h_1(k) s_1\left(\frac{l-k}{N_{\text{OSF}}}\right) + h_2(k) s_2\left(\frac{l-k}{N_{\text{OSF}}}\right)\right], \tag{2}$$

where $\lfloor \cdots \rfloor$ denotes the interger part of the argument.

Usually, the raised cosine filtering is counted for in the channel impulse response h. The signal $s_i, i \in \{1, 2\}$ up-sampled with oversampling factor N_{OSF} is then non-zero only for integer arguments and hence only M terms contribute to the summation in (2). For further receiving antennas the same equation holds with the corresponding channel impulse responses.

Although the conventional correlation approach of channel estimation, which de-scrambles the received signal ignoring the error made by this quasi-orthogonal operation (see, e.g., [2]), allows to separate both signals from different cells with different scrambling sequences c_1 and c_2 as well as to separate different delays Δj of the signals from the same cell, it contains a systematic error. This error is visible in Fig. 3 where the scalar product of a scrambling vector with itself (red) and with another scrambling vector (blue) is displayed. In both cases, the second vector is circularly shifted in its index Δj. In case of real orthogonality, only the product of a scrambling vector with itself and matching indices (red cross at $\Delta j = 0$) would yield a non-vanishing result.

In order to find an exact solution for de-scrambling, we build the projection of the received signal to the scrambling vector but keep all terms resulting from deviation of scrambling from orthogonality:

$$\begin{aligned}
&\sum_{l=l_0+1}^{l_0+N_{\text{SF}}} c^{\dagger}(l)\, r(l+d) = \\
&= \sum_{l=l_0+1}^{l_0+N_{\text{SF}}} \sum_{k=0}^{N_{\text{OSF}}\cdot M} c^{\dagger}(l) c\left(\left\lfloor \frac{l-k+d}{N_{\text{OSF}}} \right\rfloor\right) \cdot \\
&\quad \left[h_1(k) s_1\left(\frac{l-k+d}{N_{\text{OSF}}}\right) + h_2(k) s_2\left(\frac{l-k+d}{N_{\text{OSF}}}\right)\right] \approx \\
&\approx h_1(d) s_1(l_0) + h_2(d) s_2(l_0).
\end{aligned} \tag{3}$$

The approximation in the last step applies the simplification of the correlation approach.

In (3), 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 $l_0 = N_{SF}$, the spreading sequence length, and higher symbol indices are related to the corresponding integer multiples of N_{SF}.

Although the scrambling vector $\mathbf{c}$ and its index l operate on chips, both the received data $\mathbf{r}$ and the channel impulse response function $\mathbf{h}$ are defined on sub-chip level and thus d and k take into account oversampling. During the upsampling operation, the data vectors $\mathbf{s}_i, i \in \{1,2\}$, are extended according to the oversampling factor by filling the new elements with zeros. This means that all terms on the right hand side of (3) with $(l-k+d)/N_{\text{OSF}} \notin \mathbb{Z}$ vanish because the corresponding $s_i = 0$. If we now define

$$\begin{aligned}
\kappa_{l_0,j}(d,k) &\equiv \kappa_{l_0,j}(d-k) = \\
&= \sum_{l=l_0+1}^{N_{\text{OSF}}\cdot N_{\text{SF}}} c^\dagger(l) c\left(\left\lfloor \frac{l-k+d}{N_{\text{OSF}}} \right\rfloor\right) s_j\left(\frac{l-k+d}{N_{\text{OSF}}}\right) \\
\kappa_{l_0,j}(d-k) &= 0 \quad \forall d-k \notin \mathbb{Z} \\
\rho_{l_0}(d) &= \sum_{l=l_0+1}^{N_{\text{SF}}} c^\dagger(l) r(l+d) \qquad (4)
\end{aligned}$$

for a channel of length $M+1$ measured in chip duration equation (3) can be written as

$$\rho_{l_0}(d) = \sum_{k=0}^{N_{\text{OSF}}\cdot M} \left[h_1(k)\kappa_{l_0,1}(d-k) + h_2(k)\kappa_{l_0,2}(d-k)\right]. \qquad (5)$$

The correlative de-scrambling is contained in this formulation as special case, namely setting $\kappa(d-k) = \delta_{d,k}$. The channel coefficients are then directly given by ρ.

We are now 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

$$\begin{aligned}
d &= N_{\text{OSF}} \left\lfloor \frac{d}{N_{\text{OSF}}} \right\rfloor + d \mod N_{\text{OSF}} = \hat{d} + \check{d} \\
\hat{d} &= \left\lfloor \frac{d}{N_{\text{OSF}}} \right\rfloor \qquad (6) \\
\check{d} &= d \mod N_{\text{OSF}}
\end{aligned}$$

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

$$\begin{aligned}
&\rho_{l_0}(\hat{d}+\check{d}) = \\
&= \sum_{k=0}^{M} \sum_{j=0}^{N_{\text{OSF}}-1} \Big[h_1(N_{\text{OSF}}k+j)\kappa_{l_0,1}(\hat{d}+\check{d}-N_{\text{OSF}}k-j) + \\
&\qquad h_2(N_{\text{OSF}}k+j)\kappa_{l_0,2}(\hat{d}+\check{d}-N_{\text{OSF}}k-j)\Big] = \\
&= \sum_{k=0}^{M} \Big[h_1(N_{\text{OSF}}k+\check{d})\kappa_{l_0,1}(\hat{d}-N_{\text{OSF}}k) + \\
&\qquad h_2(N_{\text{OSF}}k+\check{d})\kappa_{l_0,2}(\hat{d}-N_{\text{OSF}}k)\Big]. \qquad (7)
\end{aligned}$$

In (7), we have separated integer and oversampled parts in the sum of the right hand side of (5) by splitting

$$k = N_{\text{OSF}} \left\lfloor \frac{k}{N_{\text{OSF}}} \right\rfloor + j \overset{\left\lfloor \frac{k}{N_{\text{OSF}}} \right\rfloor \to k}{\longrightarrow} N_{\text{OSF}}k + j, \qquad (8)$$

i.e., the summation over k is restricted to run from 0 to M whereas the oversampled indices are collected in j, which runs from 0 to N_{OSF} and finally is executed according to the definition given in (4).

IV. Scrambling-Exact Channel Estimation

In [8], 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. The same scrambling code for both cases is used in any case.

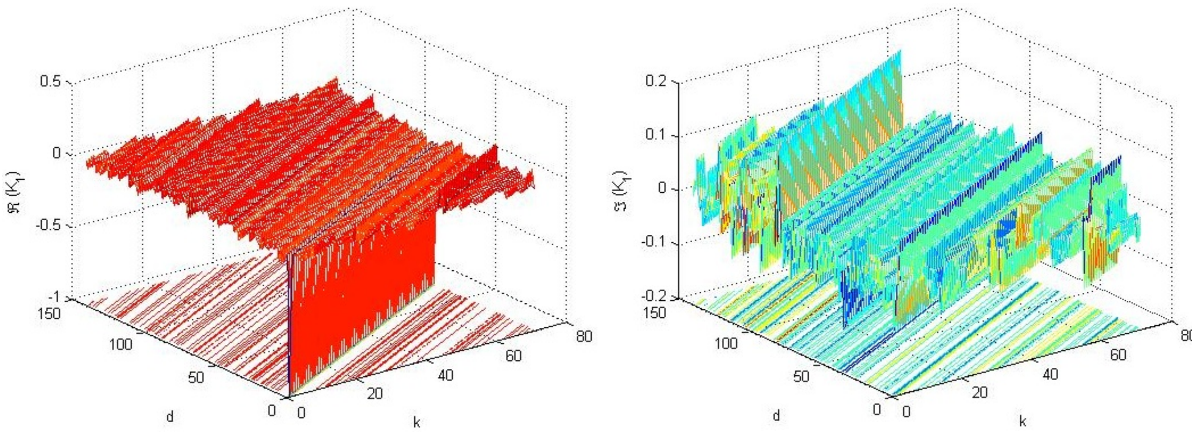
If two adjacent symbols are merged both options can formally be described by a virtual spreading with length 512 where the effective spreading codes can be defined in the first case as $[c_{\text{SF1}}\, c_{\text{SF1}}]$ and $[c_{\text{SF2}}\, \text{inv}(c_{\text{SF1}})]$, and in the second case as $[c_{\text{SF1}}\, c_{\text{SF1}}]$ and $[c_{\text{SF2}}\, c_{\text{SF2}}]$. 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.,

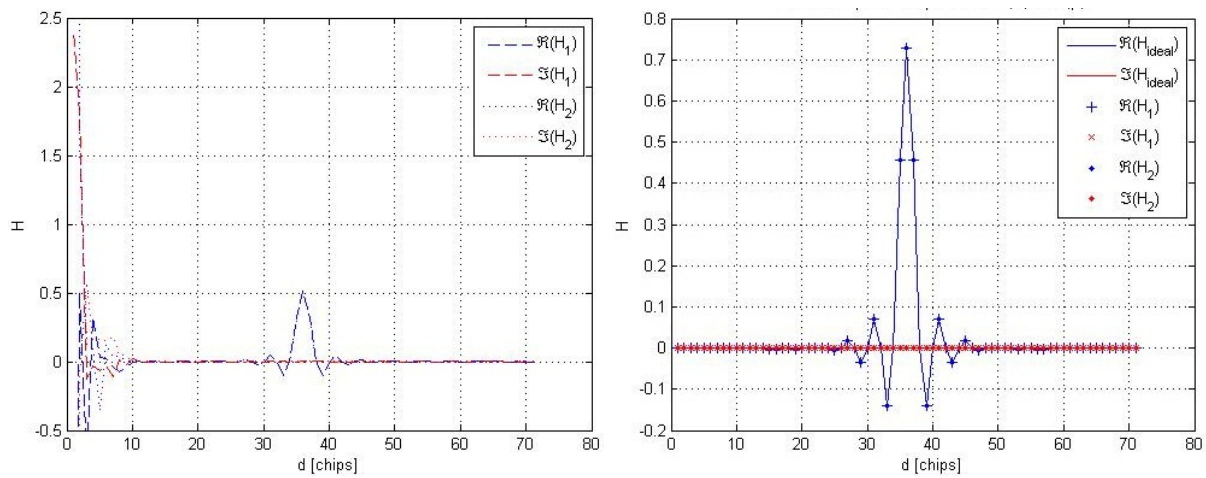
$$\begin{pmatrix} \rho_{l_0}(N_{\text{OSF}}\cdot 0+\check{d}) \\ \vdots \\ \rho_{l_0}(N_{\text{OSF}}\cdot M+\check{d}) \\ \vdots \\ \rho_{l_0}(N_{\text{OSF}}\cdot(2M+1)+\check{d}) \end{pmatrix} = \mathbf{K}_{l_0} \begin{pmatrix} h_1(N_{\text{OSF}}\cdot 0+\check{d}) \\ h_2(N_{\text{OSF}}\cdot 0+\check{d}) \\ \vdots \\ h_1(N_{\text{OSF}}\cdot M+\check{d}) \\ h_2(N_{\text{OSF}}\cdot M+\check{d}) \end{pmatrix}, \qquad (9)$$

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



(a) (b)

Figure 4. Real (a) and imaginary (b) part of odd columns of pilot scrambling matrix K for a single tap channel including raised cosine filtering.



(a) (b)

Figure 5. Channel Impulse Response H averaged over all symbols of a frame (a) and calculated with averaged K (b).

$$\mathbf{K}_{l_0} = \left(\begin{array}{ccccc} \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) & \ldots & \kappa_{l_0,1}(-M) & \kappa_{l_0,2}(-M) \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ \kappa_{l_0,1}(M) & \kappa_{l_0,2}(M) & \ldots & \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ \kappa_{l_0,1}(2M{+}1) & \kappa_{l_0,2}(2M{+}1) & \ldots & \kappa_{l_0,1}(M{+}1) & \kappa_{l_0,2}(M{+}1) \end{array} \right). \quad (10)$$

If the scrambling sequences would be completely orthogonal, the pilot scrambling matrix (10) would contain non-zero elements ($\kappa_j(0)$) in the first $M+1$ rows only, i.e., rows $M+1 \ldots 2M+1$, would not contribute at all to the solution thus making $\mathbf{K}$ un-invertible. The accuracy of these equations is therefore small even if the non-orthogonality of scrambling 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 IV-B), or the approach has to be extended such as to obtain equations for each pilot symbol (see Section IV-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. 4.

The upper half of $\mathbf{K}$ shows large real entries on the main diagonal related to κ with argument 0, whereas the other elements are dominated by the small scrambling correlation coefficients with arguments different from 0. 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. 5a for an AWGN channel with raised cosine filter.

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

$$\begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix} = \mathbf{K}^{-1}_{\text{single},l_0} \begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \vdots \\ \rho_{l_0}(2M+1) \end{pmatrix}. \qquad (11)$$

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 $\mathbf{K} \to \langle \mathbf{K} \rangle$ and $\rho \to \langle \rho \rangle$. 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 $\langle \mathbf{K} \rangle_{\text{frame}}$ becomes invertible. For an AWGN channel with raised cosine filter it provides a perfectly estimated channel impulse response, see Fig. 5b.

C. Twin Symbol Channel Estimation

The averaging procedure discussed in Section IV-B assumes that $\mathbf{K}$ 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 does not need the above restriction: Both pilot scrambling matrix and de-spread received signal vector are constructed using two symbols with different pilot sequences,

$$\begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \rho_{l_1}(0) \\ \vdots \\ \rho_{l_1}(M) \end{pmatrix} = \mathbf{K}_{\text{twin},l_{0,1}} \begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix}, \qquad (12)$$

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

$$\mathbf{K}_{\text{twin},l_{0,1}} = \qquad (13)$$
$$\begin{pmatrix} \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) & \ldots & \kappa_{l_0,1}(-M) & \kappa_{l_0,2}(-M) \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ \kappa_{l_0,1}(M) & \kappa_{l_0,2}(M) & \ldots & \kappa_{l_0,1}(0) & \kappa_{l_0,2}(0) \\ \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) & \ldots & \kappa_{l_1,1}(-M) & \kappa_{l_1,2}(-M) \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ \kappa_{l_1,1}(M) & \kappa_{l_1,2}(M) & \ldots & \kappa_{l_1,1}(0) & \kappa_{l_1,2}(0) \end{pmatrix}.$$

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

$$\begin{pmatrix} h_1(0) \\ h_2(0) \\ \vdots \\ \vdots \\ h_1(M) \\ h_2(M) \end{pmatrix} = \mathbf{K}^{-1}_{\text{twin},l_{0,1}} \begin{pmatrix} \rho_{l_0}(0) \\ \vdots \\ \rho_{l_0}(M) \\ \rho_{l_1}(0) \\ \vdots \\ \rho_{l_1}(M) \end{pmatrix}. \qquad (14)$$

Now, all rows and columns of the pilot scrambling matrix contain (at least) one main element as displayed in Fig. 6. 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 symbols of the second pilot being inverse to the ones of the first pilot.

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. 7), but also for each twin symbol. This means that no averaging is required in the absence of noise.

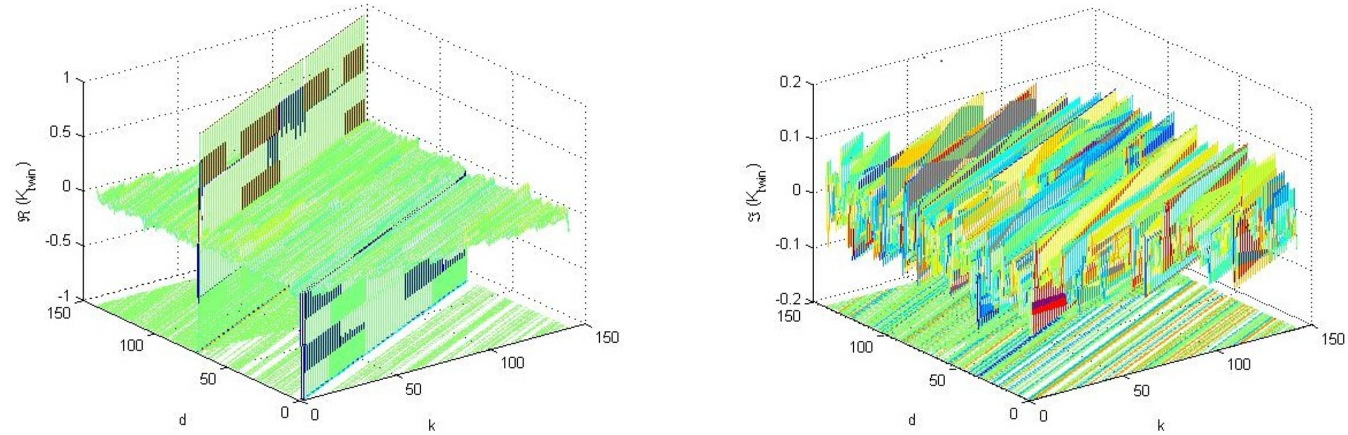
V. Link Level Simulations

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 as it is defined, e.g., in [19], table 1.2.2.2, Channel 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 [12] was used. The transmitting antennas were assumed to be mounted crosswise diagonal (X), the receiving antennas crosswise perpendicular vertical and horizontal (+) and crosstalk between the polarization directions was allowed. The basic model is displayed in Fig. 8.

The general relation between signals $\mathbf{s}_i, i \in \{1,2\}$ sent from antenna i and the signal $\mathbf{r}_k$ received at antenna k is then given by

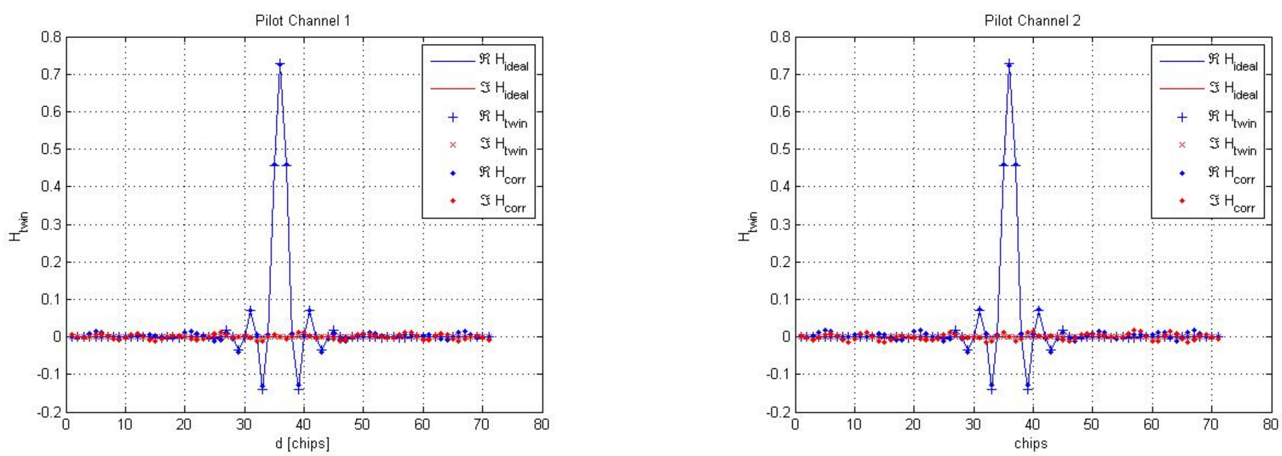
$$\mathbf{r}_k = \begin{pmatrix} \cos\theta_k & \sin\theta_k \end{pmatrix} \begin{pmatrix} h_{Hk}\sqrt{1-\kappa_{HV}^2} & h_{Vk}\kappa_{VH} \\ h_{Hk}\kappa_{HV} & h_{Vk}\sqrt{1-\kappa_{VH}^2} \end{pmatrix} \begin{pmatrix} \cos\phi & -\sin\phi \\ \sin\phi & \cos\phi \end{pmatrix} \begin{pmatrix} \mathbf{s}_1 \\ \mathbf{s}_2 \end{pmatrix}, \qquad (15)$$

with $\phi = \pi/4$ for X-mounted transmitting antennas and $\theta = \{\pi/2, 0\}$ for receiving antennas with + configuration. The parameters κ_{HV} and κ_{VH} take into account crosstalk from horizontal to vertical and from vertical to horizontal polarization direction [20], respectively. Besides crosstalk the channels were assumed to be uncorrelated.



(a) (b)

Figure 6. Real (a) and imaginary (b) part of Pilot scrambling matrix K_{twin} for one twin symbol.



(a) (b)

Figure 7. Twin Channel Impulse Response averaged over all symbols, compared with conventional correlation result and ideal case.

The block error rate (BLER) for each data stream was determined as a function of SNR. As measure for performance serve the SNRs for a working point (WP) at 10% BLER. The simulator operated on sub-chip level with oversampling factor $N_{\text{OSF}} = 2$, and also raised cosine filtering was included.

To demonstrate the enhanced impact of channel estimation errors in case of cancelling inter-stream interference, Fig. 9 shows simulation results with channel estimation via the correlation approach (denoted with *corr*) and with scrambling-exact (*twin*) channel estimation for both interference cancellation (*IC*) switched on and off. The respective SNR of the working points measured in dB is given listed in the legend.

Regarding the results for the first, stronger data stream displayed with solid lines it can be seen that interference cancellation has practically no impact on the performance, the SNR difference of the working point is 0.1dB, whereas scrambling exact channel estimation gains about 1.5dB. Without interference cancellation, the situation is similar also for the second, weaker data stream: The benefit of scrambling-exact channel estimation for the working point is about 1.5dB, but a really huge gain is achieved if interference cancellation and scrambling-exact channel estimation are combined: In this case, the working point is shifted overall by more than 5dB. This result confirms the expectation of Section II where we explained the increased impact of errors in channel estimation when the contribution of the correctly decoded data stream is removed from the incoming signal.

VI. Runtime-Efficient Implementation

From the explanation in Section III it is clear that a scrambling-exact channel estimation has to be paid with increased implementation and runtime effort. Instead of

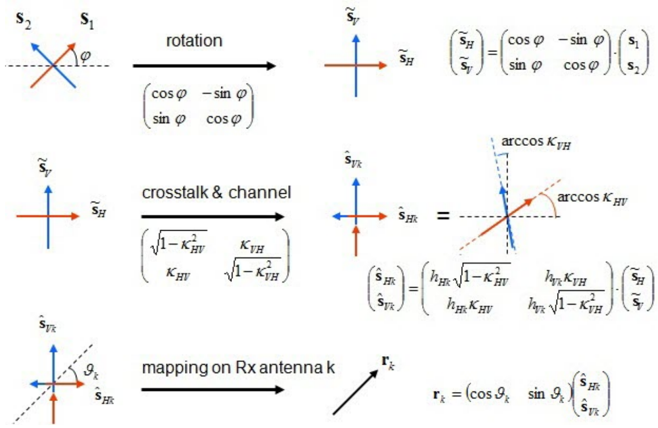


Figure 8. Channel model including crosstalk between horizontal and vertical polarized beams.

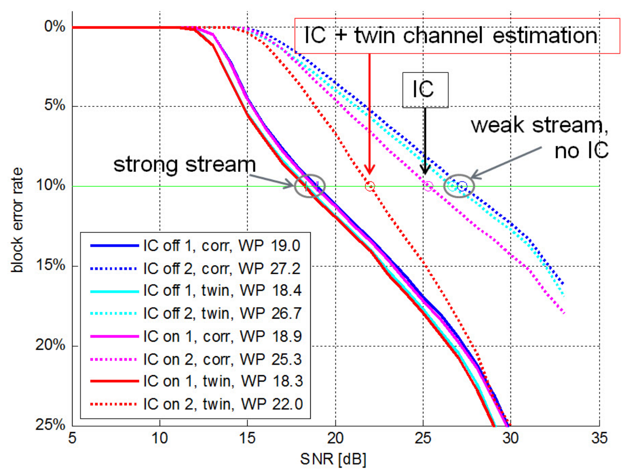


Figure 9. Link level simulation results with simple correlation and scrambling-exact channel estimation with and without interference cancellation.

solving (5) in the correlative limit,

$$\rho_{l_0}(d) = h_1(d) + h_2(d), \tag{16}$$

the elements κ of the pilot scrambling matrix $\mathbf{K}$ defined in (4) have to be calculated and the received signal convoluted with the scrambling vector to be divided by $\mathbf{K}$.

Fortunately, the pilot symbols s are fix and the scrambling code elements c are constant at least within a cell. This allows executing the more expensive part, the calculation of the matrix elements, in an initialisation phase during the handover. Please note that the pilot scrambling matrix is required only for decoding the HSDPA payload, not for the control channels as the latter are not transmitted in MIMO but in legacy SISO mode even if MIMO is active for the respective UE.

The cheaper part is the division by $\mathbf{K}$, a matrix of size $2(M+1)\times 2(M+1)$, which has to be executed permanently,

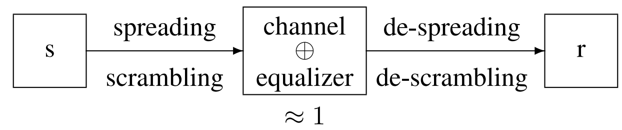


Figure 10. Spreading and de-spreading chain for transmitted data.

i.e., during each update of the channel estimation as it is applied to the varying received pilot signal. As the dimensions of $\mathbf{K}$ are comparable to the size of the LMMSE equalizer matrix, which also contains matrix division and several matrix multiplications, the equalization procedure can be regarded as upper boundary for estimating the runtime effort of this operation.

In principle it is possible to calculate the inverse of $\mathbf{K}$ in advance, too, but this brings only minor benefit as matrix division and matrix multiplication are of similar complexity for good conditioned matrix inversion. Even more important is, however, the fact that the scrambling sequence c has a length of 38400 [17] (number of chips in one frame) and the inverse of all potential excerpts of it have to be prepared if they shall be pre-calculated. Even if one restricts to solving the channel estimation equation always within the twin symbol boundaries, which have a length of 512 chips, 75 different matrices would have to be stored.

In the calculation of the matrix elements κ of (4) two types of arguments occur, namely the summation index l and the shifted index $(l+d-k)/N_{\text{OSF}}$. Including the raised cosine filter in the channel impulse response leaves the upsampled but not yet filtered signal for s, which is non-zero only for integer arguments, reducing thus the number of elements contributing to the sum to $M+1$. The elements of κ hence are non-zero only on chip-level,

$$\begin{aligned}\kappa_{l_0,j}(k) &= \sum_{l=l_0+1}^{N_{\text{SF}}} c^\dagger(l)c(l+k)s_j(l+k)\\ &= \sum_{l=l_0+1}^{N_{\text{SF}}} c^\dagger(l)\sigma_j(l+k),\\ \sigma_j(i) &\equiv c(i)s_j(i).\end{aligned} \tag{17}$$

Now, the elements for the required sequence of the channel impulse response including the raised cosine filter, $k \in -M \ldots M$, can be calculated within one loop by just circular shifting the index of σ by one step to the right with increasing k. The matrix $\mathbf{K}$ is then constructed by cutting out line-wise the required part of κ.

VII. Scrambling Impact on Data Detection

The elaboration in the previous sections has shown the significant impact of scrambling on channel estimation and hence rises the question how quasi-orthogonality of scrambling affects data detection. To discuss this in more detail we will start with the data at the equalizer output and assume that the data streams are well separated and the transmitted signal before upsampling is re-constructed. We can then restrict the discussion to the process chain as displayed in Fig. 10.

This operation chain can formally be written as

$$r(l) = \sum_{k=0}^{N_{\mathrm{SF}}-1} c^{\dagger}(l+k)c(l+k)s(l+k). \tag{18}$$

From (18) we see that de-scrambling the data only makes use of the normality but not of the quasi-orthogonal property of scrambling as only coefficients c at the same time index $l+k$ occur. De-scrambling the data is therefore not affected by the deviation of scrambling sequences from orthogonality. Of course quasi-orthogonality is required also for data detection in order to remove interference from other cells but this cannot be included correctly here as otherwise both scrambling sequences and signal strengths of all potential interferers would have to be known. In contrast to the two pilot signals, interfering signals propagate along different physical channels, which are especially different from the channel relevant for both data and pilot signals of the UE of interest. Therefore, it can be expected that superposition of different interferers and equalisation with channel coefficients not matching these physical channels the interfering signals propagate through will lead to a further whitening of these contributions so that no systematic error persists the interfering signals from other cells.

VIII. Conclusion

The usual procedure for estimating the radio channels discards the small non-orthogonality of the autocorrelation function of the pseudorandom scrambling sequence. As long as only one pilot signal is transmitted this effect is negligible as the error is more or less proportional to the strength of the received pilot signal at the respective chip position. Delayed multi-path taps bearing no pilot signal for the respective chip are not affected at all.

The usage of multiple transmitting antennas, e.g., in STTD or MIMO systems, allows to benefit from spatial and polarization diversity, which however requires to measure all physical channels from each transmitting to each receiving antenna. This is possible only if different pilot patterns (or different linear combinations of them generated by different pre-coding) are sent from the different transmitting antennas. Ignoring deviation from orthogonality then causes that energy scattered by scrambling out of a certain pilot pattern will appear in another pilot pattern. This contribution is no more proportional to the signal strength of the respective pilot signal at the regarded chip. Also taps in fact containing no pilot signal from the investigated chip position will experience a signal contribution causing a huge relative error.

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.

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 [4], cancelling inter-antenna interference [5] or extending the RAKE correlation approach to a LMMSE one [3] and thus taking into account all kinds of interference. All of these suggestions can basically be extended by scrambling-exact dealing the pilot patterns as proposed in [1] even if we have investigated here only the simple RAKE correlation approach, including appropriate filtering as described in [4], in the simulator.

Although the scrambling induced error does not occur in data detection itself and is negligible in channel estimation in many cases, high data rates in combination with MIMO and transmission of two data streams require more accurate channel estimation. This can be reached by taking into account the scrambling 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 applied to the received signal providing the channel impulse response.

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. The complexity of an efficient implementation is then comparable to the one of an LMMSE equalizer. 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 more than 5dB for 10% block error rate.

Currently it is discussed in 3GPP to extend MIMO to 4 data streams to be sent in parallel (see, e.g., [11]). As this requires 4 pilot signals to be transmitted an even stronger mutual distortion impact of pilot signals due to scrambling can be expected. Although it is far from being clear how the pilot scheme will be extended, scrambling induced deviation from orthogonality has to be included in the course of its definition.

ACKNOWLEDGMENT

The Author would like to thank Elke Hallmann and Klaus Knoche for many inspiring discussions during this study.

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