Inter-Carrier Interference Estimation in MIMO OFDM Systems with Arbitrary Pilot Structure

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***Abstract*—In scenarios with time-varying channels such as intelligent traffic systems or high speed trains, the orthogonality between subcarriers in orthogonal frequency division multiplexing (OFDM) is destroyed leading to inter-carrier interference (ICI). In the literature, ICI equalization algorithms have been proposed; however, they assume perfect channel knowledge at sample level. Unfortunately, existing channel estimation algorithms do not provide accurate channel estimates at high Doppler spreads. Therefore, it is not possible to transmit data with high spectral efficiency. In this paper, we propose an algorithm for ICI estimation that can be applied to OFDM systems with an arbitrary pilot structure. Thus, our algorithm can be applied to any already standardized OFDM system. Our ICI estimator models the channel variation by means of a basis expansion model (BEM). The performance of the estimator and of the subsequent equalization is evaluated in a UMTS long term evolution (LTE) link level simulator. In a Rayleigh fading scenario, the proposed algorithm allows a velocity increase of 150 km/h without throughput degradation. The gain in terms of the post-equalization signal to interference and noise ratio (SINR) is about 3.7 dB at a user speed of 300 km/h.**

***Index Terms*—LTE, ICI, Channel Estimation, Fast Fading, OFDM, MIMO.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) is used in most current and upcoming mobile communication systems. Such systems perform well when the channel is not varying during the duration of one OFDM symbol. However, mobile scenarios in which the channel is varying rapidly are becoming more and more important for intelligent traffic systems or high speed trains. If the channel is not constant during the transmission of one OFDM symbol, inter-carrier interference (ICI) occurs and the performance of the system is degraded. Therefore, there is a need to introduce receivers that combat ICI.

Related Work: ICI estimation and equalization have been addressed in previous work, for example in [1–11]. In [1], the ICI for single input single output (SISO) and multiple input multiple output (MIMO) transmissions is analyzed. The authors propose to employ pilot symbols located at adjacent subcarriers in order to estimate the ICI. This approach is in contradiction to the common agreement that scattered pilot symbols are optimal [2, 3]. Nevertheless, such an approach would be suitable for ICI estimation in the case of a SISO system. In the case of a MIMO system, such a pilot symbol pattern results in a huge overhead. In [4–6] ICI estimation and mitigation assume that the channel is varying linearly in the time domain. Hijazi and Ros propose to use polynomials for channel estimation in [7]. However, their estimator works only with a limited order of the polynomials. Numerous different equalization algorithms are proposed in [8–10], in which the authors assume perfect channel knowledge for each signal sample. However, this information is not available at the receiver and algorithms proposed so far [4–7] cannot estimate the time-variant channel impulse response at sample level precisely enough at high Doppler spreads.

Scientific Contribution:

1) Building on [7] we model the channel variation in the time domain by means of a general set of orthogonal basis functions to optimize the square-bias variance tradeoff for time-variant channel estimation.
2) In Section IV, we address the problem of inverting an ill conditioned matrix in [7] by introducing an additional regularization step, such that the basis function set is orthogonal on the given pilot grid.
3) Our algorithm can be utilized as a simple extension to any existing channel estimator. Therefore, our method does not require any special pilot structure and can be applied in existing standardized systems such as UMTS LTE.
4) In Section V, we evaluate the performance of linear, polynomial and discrete prolate spheroidal basis functions for ICI estimation in a fully standard compliant LTE link level simulator [12].

II. System Model

long term evolution (LTE) is the current standard of the cellular communication 3rd Generation Partnership Project (3GPP). It supports technologies such as different MIMO schemes, adaptive coding and modulation (ACM) and hybrid automated repeat request (H-ARQ) that allow to transmit data with high spectral efficiency. LTE supports bandwidth from 1.4 MHz up to 20 MHz, corresponding to a number of data subcarriers ranging from 72 to 1200. The subcarrier spacing is fixed to 15 kHz. Depending on the cyclic prefix length, being either extended or normal, each LTE subrame consists of 12 or 14 OFDM symbols, respectively. The duration of an LTE subframe is 1 ms.

The structure of the pilot symbols is described in [13]. This pilot symbol pattern allows to estimate a MIMO channel as independent SISO channels, neglecting spatial correlation.

Therefore it is sufficient to consider a SISO system model. The n-th received OFDM symbol $\mathbf{y}_n$ at one receive antenna port can be written as

$$\mathbf{y}_n = \mathbf{G}_n \mathbf{x}_n + \mathbf{w}_n, \tag{1}$$

where the matrix $\mathbf{G}_n$ is the channel matrix in the frequency domain of the n-th OFDM symbol and $\mathbf{w}_n$ denotes additive complex white Gaussian noise with zero mean and variance σ_w^2. The vector $\mathbf{x}_n$ is comprised of data symbols $\mathbf{x}_{\mathrm{d},n}$ and pilot symbols $\mathbf{x}_{\mathrm{p},n}$

$$\mathbf{x}_n = \mathbf{P} \left[\mathbf{x}_{\mathrm{d},n}^{\mathrm{T}} \mathbf{x}_{\mathrm{p},n}^{\mathrm{T}}\right]^{\mathrm{T}}, \tag{2}$$

permuted with a permutation matrix $\mathbf{P}$. The length of the vector $\mathbf{x}_n$ is K, corresponding to the number of subcarriers. Note that according to Equation (2), the vectors $\mathbf{y}_n$ and $\mathbf{w}_n$ can also be divided into two parts corresponding to the pilot symbol positions and to the data symbol positions.

The channel matrix of the n-th OFDM symbol in the time domain is given by

$$\mathbf{H}_n = \tag{3}$$
$$\begin{bmatrix} h_{n,1,0} & 0 & \cdots & & \cdots & & \cdots & 0 \\ \vdots & \ddots & \ddots & & & & & \vdots \\ h_{n,1,N_\mathrm{h}-1} & & \ddots & & \ddots & & & \vdots \\ 0 & \ddots & & & \ddots & & \ddots & \vdots \\ \vdots & \ddots & \ddots & & & & \ddots & 0 \\ 0 & \cdots & 0 & & h_{n,N_\mathrm{s}-N_\mathrm{h}+1,N_\mathrm{h}-1} & & \cdots & h_{n,N_\mathrm{s},0} \end{bmatrix},$$

where $h_{n,m,\tau}$ is the channel coefficient of the m-th sample within the n-th OFDM symbol at delay τ. The length of the channel impulse response is N_h. The time domain channel matrix $\mathbf{H}_n$ is transformed to the frequency domain by considering the OFDM signal structure:

$$\mathbf{G}_n = \underbrace{\mathbf{F}_{\mathrm{guardrem}} \mathbf{D} \mathbf{F}_{\mathrm{CPrem}}}_{\mathbf{D}_1} \mathbf{H}_n \underbrace{\mathbf{F}_{\mathrm{CP}} \mathbf{D}^{\mathrm{H}} \mathbf{F}_{\mathrm{guard}}}_{\mathbf{D}_2} \tag{4}$$

The matrices $\mathbf{F}_{\mathrm{guardrem}}$ and $\mathbf{F}_{\mathrm{CPrem}}$ correspond to the removal of the guard subcarriers and the cyclic prefix, respectively [14]. The matrices $\mathbf{F}_{\mathrm{guard}}$ and $\mathbf{F}_{\mathrm{CP}}$ add guard subcarriers and the cyclic prefix, respectively. The matrix $\mathbf{D}$ is the DFT matrix. If the channel is not varying during the transmission of one OFDM symbol, $\mathbf{G}_n$ is a diagonal matrix. If the channel is varying within one OFDM symbol, $\mathbf{G}_n$ is not diagonal and ICI occurs.

We define the Toeplitz operator $\mathrm{Toep}(\cdot)$ as

$$\mathrm{Toep}(\mathbf{b}) \triangleq \begin{bmatrix} b(1) & 0 & \cdots & \cdots & \cdots & 0 \\ \vdots & \ddots & \ddots & & & \vdots \\ b(N) & & \ddots & \ddots & & \vdots \\ 0 & \ddots & & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & & \ddots & 0 \\ 0 & \cdots & 0 & b(N) & \cdots & b(1) \end{bmatrix}, \tag{5}$$

with the vector $\mathbf{b}$ being of length N.

The channel in the time domain can be split into two parts, one corresponding to the mean channel and one to the time variation of the channel:

$$\mathbf{H}_n = \mathrm{Toep}\left(\bar{\mathbf{h}}_n\right) + \Delta\mathbf{H}_n \tag{6}$$

Here the vector $\bar{\mathbf{h}}_n$ comprises the mean channel impulse response as experienced by the n-th OFDM symbol. The channel in the frequency domain using this structure is given by

$$\mathbf{G}_n = \mathbf{D}_1 \left(\mathrm{Toep}\left(\bar{\mathbf{h}}_n\right) + \Delta\mathbf{H}_n\right) \mathbf{D}_2 \tag{7}$$
$$= \mathrm{diag}\left(\mathbf{g}_n\right) + \mathbf{D}_1 \Delta\mathbf{H}_n \mathbf{D}_2, \tag{8}$$

where the vector $\mathbf{g}_n$ contains the diagonal elements of the channel matrix in the frequency domain. The operator $\mathrm{diag}(\mathbf{b})$ creates a diagonal matrix with the vector $\mathbf{b}$ on the main diagonal.

III. Channel Estimation

In this section, we present state-of-the-art channel estimators for estimating the main diagonal elements of the channel matrix in the frequency domain. Note that, although we restrict ourselves to least squares (LS) and linear minimum mean square error (LMMSE) channel estimation here, our ICI estimation method presented in the next section can be combined with any OFDM channel estimator. A typical OFDM channel estimator estimates only the main diagonal elements of the frequency domain channel matrix $\mathbf{G}_n$. Our ICI estimator, on the other hand, estimates the remaining off-diagonal elements of the channel matrix by using the main diagonal elements of the channel matrices of the remaining OFDM symbols through interpolation.

A. LS Channel Estimation

The LS channel estimator [15] for the pilot symbol positions is given by

$$\hat{\mathbf{g}}_{\mathrm{p}}^{\mathrm{LS}} = \arg\min_{\hat{\mathbf{g}}_{\mathrm{p}}} \left\|\mathbf{y}_{\mathrm{p}} - \mathbf{X}_{\mathrm{p}}\, \hat{\mathbf{g}}_{\mathrm{p}}\right\|_2^2 = \mathbf{X}_{\mathrm{p}}^{-1}\, \mathbf{y}_{\mathrm{p}}, \tag{9}$$

where the matrix $\mathbf{X}_{\mathrm{p}}$ is a diagonal matrix comprising pilot symbols on the main diagonal.

B. LMMSE Channel Estimation

The LMMSE channel estimator requires the knowledge of the second order statistics of the channel and the noise. It can be shown that the LMMSE channel estimate is obtained by multiplying the LS estimate with a filtering matrix $\mathbf{A}_{\mathrm{LMMSE}}$ [16, 17]

$$\hat{\mathbf{g}}_{\mathrm{LMMSE}} = \mathbf{A}_{\mathrm{LMMSE}} \hat{\mathbf{g}}_{\mathrm{p}}^{\mathrm{LS}}. \tag{10}$$

In order to find the LMMSE filtering matrix, the mean square error (MSE) is minimized, leading to

$$\mathbf{A}_{\mathrm{LMMSE}} = \mathbf{R}_{\mathbf{g},\mathbf{g}_{\mathrm{p}}} \left(\mathbf{R}_{\mathbf{g}_{\mathrm{p}},\mathbf{g}_{\mathrm{p}}} + \sigma_w^2 \mathbf{I}\right)^{-1}, \tag{11}$$

where the matrix $\mathbf{R}_{\mathbf{g}_{\mathrm{p}},\mathbf{g}_{\mathrm{p}}} = \mathbb{E}\left\{\mathbf{g}_{\mathrm{p}} \mathbf{g}_{\mathrm{p}}^{\mathrm{H}}\right\}$ is the channel covariance matrix at the pilot symbol positions, and the matrix $\mathbf{R}_{\mathbf{g},\mathbf{g}_{\mathrm{p}}} = \mathbb{E}\left\{\mathbf{g} \mathbf{g}_{\mathrm{p}}^{\mathrm{H}}\right\}$ is the channel cross-correlation matrix.

IV. ICI Estimation

In this section, we generalize a method for estimating the ICI in the frequency domain introduced in [7] to arbitrary basis functions. Furthermore, we propose improvements that allows to increase the polynomial order and as a consequence to obtain better ICI estimates at high Doppler spreads. Furthermore, the overall complexity of the estimator is reduced. The estimated channel coefficients of one subframe are the only input to the ICI estimator. We assume that $\overline{\mathbf{h}}_n$ is the same as the channel observed in the middle of the particular OFDM symbol. The ICI estimation algorithm can be used as add-on to any channel estimator. Here, as "channel estimator" we understand a signal processing block that estimates only the diagonal elements of the channel matrix in the frequency domain. The off-diagonal elements are calculated by the ICI estimator as explained below.

A. General ICI Estimator

The frequency domain channel matrix can be decomposed using a set of basis functions

$$\mathbf{G}_n = \sum_{i=0}^{N_{\text{order}}} \text{diag}\left(\gamma_n^{(i)}\right) \mathbf{D}_1 \mathbf{T}^{(i)} \mathbf{D}_2, \tag{12}$$

where the matrices $\mathbf{T}^{(i)}$ are diagonal matrices comprised of the corresponding basis vectors on their main diagonals. The channel estimator delivers an estimate of the diagonal elements of the channel matrix in the frequency domain, which corresponds to the mean of the channel during the transmission of one OFDM symbol. In [7] it is shown, that if the mean channel of several consecutive OFDM symbols is known, the optimal coefficient can be obtained by means of a linear regression. Using polynomials as the basis functions in (12), the coefficients of the basis expansion model (BEM) are obtained as follows

$$\left[\hat{\gamma}_n^{(0)}\, \hat{\gamma}_n^{(1)}\, \hat{\gamma}_n^{(2)} \cdots \hat{\gamma}_n^{(N_{\text{order}})}\right] = \left(\mathbf{M}^{\mathrm{H}}\mathbf{M}\right)^{-1} \mathbf{M}^{\mathrm{H}} \left[\hat{\mathbf{g}}_1 \hat{\mathbf{g}}_2 \cdots \hat{\mathbf{g}}_{N_{\text{symbol}}}\right]^{\mathrm{T}}, \tag{13}$$

where the matrix $\mathbf{M}$ contains the sampled basis function column wise. For example, for a polynomial basis the matrix $\mathbf{M}$ is given as

$$\mathbf{M} = \left[\mathbf{1}\, \mathbf{m}\, \mathbf{m}^2 \cdots \mathbf{m}^{N_{\text{order}}}\right], \tag{14}$$

where the operator $\cdot^i$ denotes the element-wise raise to the power of i and the vector $\mathbf{m}$ has the following structure

$$\mathbf{m} = \left[\left\lfloor \frac{N_{\text{s}}}{2} \right\rfloor, N_{\text{s}} + \left\lfloor \frac{N_{\text{s}}}{2} \right\rfloor, \cdots, N_{\text{s}}\left(N_{\text{symbol}} - 1\right) + \left\lfloor \frac{N_{\text{s}}}{2} \right\rfloor\right]^{\mathrm{T}}, \tag{15}$$

with N_{symbol} being the number of OFDM symbols within one subframe.

B. Discussion

In this subsection, we will discuss the concept of the ICI estimator introduced in Section IV-A and furthermore the choice of the basis functions.

1) Linear Case: If we use polynomials as the basis spanning the channel space and set the variable $N_{\text{order}} = 1$, we assume that the channel is varying linearly in time. Higher order channel variations are not taken into account. The same assumption has been made in [4–6]. It was shown, that such an assumption is valid at low Doppler spreads.

2) Discrete prolate spheroidal (DPS) sequences: In [18] a low-dimensional subspace spanned by discrete prolate spheroidal sequences is used for time-variant channel estimation. The subspace is designed according to the maximum velocity $v_{\max}$ of the user. It is shown in [18] that the channel estimation bias obtained with the Slepian basis expansion is more than a order of magnitude smaller compared to the Fourier basis expansion (i.e. a truncated discrete Fourier transform) [19] or a polynomial. The concept introduced in [18], can be directly extended to the ICI estimation. The polynomials in (12) are replaced by DPS sequences. This approach allows to estimate the ICI more accurately as we will show by numeric simulations in Section V.

3) Orthogonalized sequences: The authors of [7] limit the polynomial order to four due to the ill-conditioned matrix in (13). With increasing number of basis functions N_{order}, the condition number of the matrix $\mathbf{M}^{\mathrm{H}}\mathbf{M}$ in (13) is also increasing. Therefore, the result of the inversion is not reliable. The maximum modeling order N_{order} depends on the choice of the basis vectors $\mathbf{m}$. The main requirement on the basis vectors is the orthogonality between their sampled version. Orthogonal sampled sequences that span the same space as the sequences $\mathbf{m}^0, \mathbf{m}^1, \mathbf{m}^2, \cdots, \mathbf{m}^{N_{\text{order}}}$ have therefore to be found. During the search for the new orthogonal sequences it has to be considered that we have to be able to construct corresponding sequences $\mathbf{t}, \mathbf{t}^2, \cdots, \mathbf{t}^{N_{\text{order}}}$ at the sample level. In order to solve the given problem with defined requirements, one can apply the Gram Schmidt orthonormalization algorithm [20] on the vectors $\mathbf{m}^0, \mathbf{m}^1, \mathbf{m}^2, \cdots, \mathbf{m}^{N_{\text{order}}}$. During the orthogonalization process, also the vectors $\mathbf{t}^0, \mathbf{t}^1, \cdots, \mathbf{t}^{N_{\text{order}}}$ have to be transformed in the same manner. The new sampled basis vectors $\mathbf{l}^i$ and the new basis vectors at the sample level $\mathbf{k}^i$ can be constructed as:

$$\mathbf{l}^i = \tilde{\mathbf{m}}^i - \sum_{j=1}^{i-1} \frac{\mathbf{l}^{j\,\mathrm{T}} \tilde{\mathbf{m}}^i}{\mathbf{l}^{j\,\mathrm{T}} \mathbf{l}^j} \mathbf{l}^j \quad 0 \le i \le N_{\text{order}} \tag{16}$$

$$\mathbf{k}^i = \mathbf{t}^i - \sum_{j=1}^{i-1} \frac{\mathbf{l}^{j\,\mathrm{T}} \mathbf{t}^i}{\mathbf{l}^{j\,\mathrm{T}} \mathbf{l}^j} \mathbf{k}^j \quad 0 \le i \le N_{\text{order}} \tag{17}$$

By applying this procedure, we manage to find basis vectors $\mathbf{k}^i$, such that their sampled versions $\mathbf{l}^i$ are orthonormal. Therefore, the linear regression from (13) simplifies to

$$\left[\hat{\gamma}_n^{(0)}\, \hat{\gamma}_n^{(1)} \cdots \hat{\gamma}_n^{(N_{\text{order}})}\right] = \mathbf{L}^{\mathrm{H}} \left[\hat{\mathbf{g}}_1 \hat{\mathbf{g}}_2 \cdots \hat{\mathbf{g}}_{N_{\text{symbol}}}\right]^{\mathrm{T}}, \tag{18}$$

TABLE I
SIMULATOR SETTINGS FOR FAST FADING SIMULATIONS

Parameter	Value
Bandwidth	1.4 MHz
Number of transmit antennas	2
Number of receive antennas	2
Receiver type	ZF
Transmission mode	Open-loop spatial multiplexing
Channel type	Rayleigh fading, Jakes Doppler spectrum
CQI	14
coding rate	873/1024 = 0.852
symbol alphabet	64 QAM
carrier frequency	2.5 GHz
number of subframes	1000

where the matrix $\mathbf{L}$ contains the vectors $\mathbf{l}^i$, $\mathbf{L} = [\mathbf{l}^0 \mathbf{l}^1 \cdots \mathbf{l}^{N_\text{order}}]$. The ICI estimate is obtained with the help of (12), where the matrices $\mathbf{K}^{(i)}$ are used instead of the matrices $\mathbf{T}^{(i)}$. The matrices $\mathbf{K}^{(i)}$ contain the vectors $\mathbf{k}^i$ on the main diagonals.

This ortogonalization process can be applied not only to polynomials but also to DPS sequences. Furthermore, the basis sequences orthogonal at OFDM symbol level, not only solve the problem of the ill-conditioned matrix, but also decrease the overall complexity of the estimator, since instead of matrix inversion in (13) only a matrix multiplication in (18) is necessary.

V. SIMULATION RESULTS

In this section, we present simulation results and discuss the performance of the presented ICI estimation algorithm. All results are reproduced with the LTE Link Level Simulator version "1.2r715" [12], which can be downloaded from www.nt.tuwien.ac.at/ltesimulator. All figures presented in this paper, can be obtained by running a script file with name LTE_sim_batch_michal_vtc_2011.m. For the simulated curves we calculated the 95% confidence intervals, which turned out to be smaller than the size of the markers plotted in the figures. Table I presents the most important simulator settings.

The time correlated channel was generated by an implementation of the Rosa Zheng model with modifications according to [18]. We generate a time correlated channel impulse response for each sample of the baseband transmit signal. Using a time-variant convolution, the output signal of the channel is calculated.

In this paragraph, we briefly describe the complete ICI aware receiver. Firstly, the ICI aware receiver carries out LMMSE channel channel estimation. Afterwards, ICI estimation for each OFDM symbol of the subframe is performed by utilizing the channel estimates of all OFDM symbol within the subframe. The ICI in the n-th OFDM symbol is given by (12). Afterwards, the ICI part caused by the pilot symbols located in a particular OFDM symbol is subtracted from the received OFDM symbol. The data symbols are then subcarrier-wise equalized using the zero forcing (ZF) equalizer of [21] extended for MIMO. We assumed that the number of non-zero off-diagonal elements of the channel matrix $\mathbf{G}_n$ is $N_\text{off} = 5$.

The signal to noise ratio (SNR) is fixed to 30 dB in all simulations. We decided to choose such a high SNR value to emphasize the effect of ICI. Furthermore such an SNR value is practical in some applications, like high speed trains or intelligent traffic systems. Under the term "typical OFDM receiver" we understand an OFDM receiver that neglects ICI.

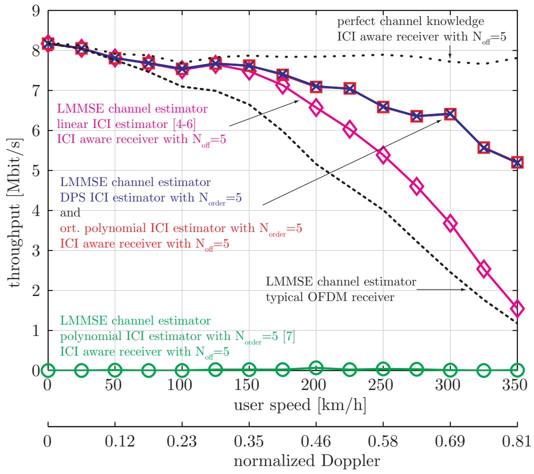

Fig. 1. Throughput for LTE with Rayleigh fading versus user velocity

In Figure 1, the throughput of a 2×2 LTE system in a flat Rayleigh fading channel is plotted versus user velocity. In this scenario, a user employing the proposed ICI estimation algorithm with $N_\text{order} = 5$ can move approximately 150 km/h faster than a user employing a typical OFDM receiver, that is, a receiver neglecting the ICI due to high user velocity. At lower user velocities, the performance of the linear ICI estimator is close to our proposed algorithm. In this region, the assumption of linear channel variation is fulfilled. At velocities higher than 100 km/h , our proposed algorithm outperforms the linear ICI estimator [4–6] and results in a speed gain of about 90 km/h. Furthermore, performance of the algorithm proposed in [7] with $N_\text{order} = 5$, is strongly limited due to the ill conditioned matrix $\mathbf{M}^\text{H}\mathbf{M}$ in (13). Note, that ICI estimator proposed in [7] with $N_\text{order} = 4$ works well up to certain Doppler spread. In Figure 1, we use $N_\text{order} = 5$ to demonstrate the ill conditioned matrix $\mathbf{M}^\text{H}\mathbf{M}$ in (13) using method proposed in [7].

In Figure 2, the post-equalization signal to interference and noise ratio (SINR) for different receivers over user velocity is shown. It can be seen that with our proposed method an SINR gain of approximately 3.7 dB can be achieved at a user speed of 300 km/h. The SINR gain of our proposed ICI estimator compared to the linear ICI estimator [4–6] is about 2.5 dB at a user speed of 300 km/h. The SINR gain at 300 km/h to ICI estimator [7] with $N_\text{order} = 5$ is about 6 dB.

Figure 3 compares the ICI estimator with different basis sequences. Utilizing DPS sequences as the channel basis instead of orthogonalized polynomials is beneficial in terms of throughput and receiver complexity. To achieve the same throughput at high Doppler spreads, less basis sequences are necessary if DPS is employed.

VI. CONCLUSION

We have proposed an ICI estimation algorithm that can be applied to OFDM systems with arbitrary pilot structure. Using

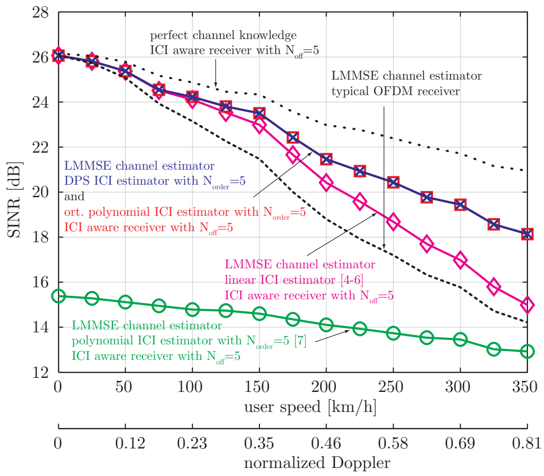

Fig. 2. Post-equalization SINR for LTE for Rayleigh fading versus user velocity

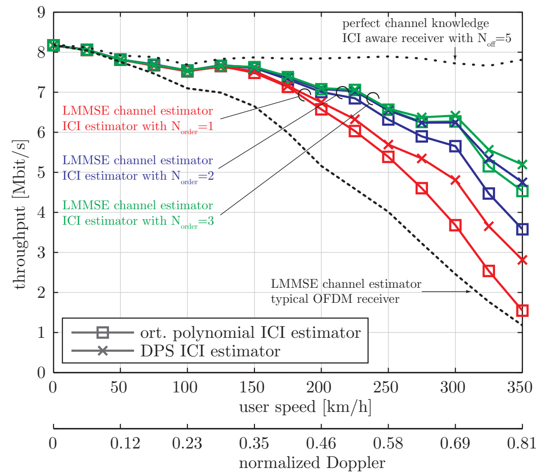

Fig. 3. Throughput for LTE for Rayleigh fading versus user velocity

the ICI estimate of our proposed estimator in combination with the assumption that the channel matrix in the frequency domain is a band matrix, we can achieve a large performance gain (SINR gain up to 3.7 dB) by means of a ZF receiver. In a flat Rayleigh fading scenario, a user is able to move approximately 150 km/h faster while obtaining the same throughput as if he would utilize an OFDM receiver that does not consider ICI. The gain while utilizing our proposed ICI estimator in terms of the post-equalization SINR is around 3.3 dB at a user speed of 300 km/h compared to the case when employing a normal receiver.

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