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Low-complexity channel estimation for OFDM systems in high-mobility fading channels

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Abstract

Nowadays, multicarrier transmission is very popular because of the high data rate requirement of wireless systems. Orthogonal frequency-division multiplexing (OFDM) is a special case of multicarrier transmission. It is considered an effective technique for frequency-selective channels because of its spectral efficiency, its robustness in different multipath propagation environments, and its ability of combating intersymbol interference. Losing subchannel orthogonality causes interchannel interference, but many conventional channel estimators assume that the channel is time-invariant during one OFDM symbol. This assumption causes error floors along the OFDM symbol for high-mobility cases. In this paper, we propose 2-dimensional pilot symbol-assisted channel estimation for wireless OFDM systems by using frequency and time-domain interpolation. This method has the advantage of minimizing the system complexity and processing delay while closely approximating the actual mobile channel. The performance of our proposed method is compared to coherent modulation with perfect channel estimation as well as other conventional channel estimation methods. Different detection techniques were used to test our proposed channel estimation algorithm.

Key Words: *OFDM systems, time-varying channel estimation, curve fitting, detection techniques*

1. Introduction

Wideband wireless communication is a rapidly growing technology. New techniques such as cognitive radio [1], cooperative diversity [2], and orthogonal frequency-division multiplexing (OFDM) have emerged in the field to meet requirements like high data rates and spectral efficiency that stem from user demands. It was shown in previous studies that multicarrier modulation methods such as OFDM are more robust to intersymbol interference (ISI) and channel noise than single-carrier systems over fast-varying communication channels [3,4].

OFDM is a multicarrier technique that divides the available spectrum into many subcarriers. This allows for individual modulation of each subcarrier and then transmission of entire OFDM blocks at a significantly lower rate than in a single-carrier system. OFDM uses the spectrum much more efficiently by spacing the

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channels much closer together. All subcarriers are orthogonal to one another, allowing the spectrum in each subchannel to overlap another without interfering with it, for preventing interference between the closely spaced carriers. Transmitting over narrow bands in parallel reduces ISI and decreases the need for complex equalization [5].

In low-mobility cases, normal OFDM-based systems have been very successful. In fading channels with very high mobility, however, time variations in one OFDM symbol introduce interchannel interference (ICI), which further degrades the performance for mobile applications because of lost subchannel orthogonality.

The most important part of the receiver structure in OFDM systems is channel estimation. Different methods have been proposed for channel estimation of OFDM systems. They can be grouped into 2 main categories: nonpilot-aided and pilot-aided. Pilot-aided channel estimation methods have better channel estimation performance than others. In the pilot-aided method, known pilot symbols are inserted periodically both in time and frequency dimensions to track the time variations and frequency selectivity of the channel. Pilot-aided methods can categorize block type, which is more favorable for rather slower fading, and comb type, which is more favorable for fast fading [6,7]. The channel complex gain at the pilot symbol positions can be easily obtained from the received signal and the known pilot symbols at the receiver. The channel impulse response (CIR) in the data region can be estimated by interpolating the CIR using the scattered pilot symbols. The channel estimation for resting nonpilot symbols can be found with interpolation.

Interpolation techniques like linear, second-order, low-pass, cubic spline, and time-domain interpolation have been investigated for OFDM systems [3,8-11]. In particular, the method proposed in [8] provides channel estimates based on piecewise constant and piecewise linear interpolations between pilots. It is simple to implement, but it needs a large number of pilots for satisfactory performance. The two-dimensional Wiener filter is accepted as the optimum channel estimation filter that minimizes the mean square error of the channel estimate [12]. It is not practical, however, because of its large implementation complexity. We propose 2 one-dimensional channel estimators that perform independently in the time and frequency domains, instead of the use of a single 2D filter, to reduce the implementation complexity.

The above-mentioned conventional channel estimators make the assumption of a time-invariant model for the channel of one OFDM symbol and do not accommodate the ICI effect caused by high mobility. Therefore, researchers have proposed various detection techniques to reduce the ICI in OFDM systems while the channel is perfectly known at the receiver [4]. In [13], time-domain channel estimation and detection techniques were presented for multicarrier signals in a fast and frequency-selective Rayleigh fading channel. All subcarriers in a given time slot are dedicated to pilot symbols (i.e. dedicated time slots) and estimation is done by using the last OFDM symbol of the frame, which requires a priori information on the channel statistics. Moreover, the piecewise linear model has been used to approximate channel variations for OFDM systems in the case of ICI [14]. However, it will be shown in the simulation section that the performance of the piecewise linear approximation falls short of expectations for high-mobility scenarios [15].

In this paper, we propose 2 one-dimensional channel estimators. We use polynomial curve fitting to approximate time variations of the channel, while a low-pass interpolator is used to find the frequency response of the channel using pilot tones. Simulations comparing the proposed method with previously proposed channel estimators demonstrate the performance of our channel estimation approach.

The paper is organized as follows. We describe our OFDM system model in Section 2. We give the 2D pilot symbol-assisted channel estimation method with frequency-domain and time-domain interpolation methods in Section 3. The computational complexity of the proposed algorithm is also investigated in Section

3. Section 4 explores the detection techniques used for OFDM systems. Simulation results and comparisons are given in Section 5. Finally, the main conclusions of the paper are presented in Section 6.

Notation: Vectors (matrices) are denoted by boldface lower (upper) case letters; all vectors are column vectors; $(.)^*$, $(.)^T$, and $(.)^H$ denote the conjugate, transpose, and conjugate transpose, respectively.

2. System model

Let us consider an OFDM system with N subcarriers and available bandwidth $BW = 1/T_s$, where T_s is the sampling period, and equal frequency spacing $\Delta f = BW/N$. At the transmitter, information bits are mapped into possibly complex-valued transmit symbols according to the modulation scheme employed. The symbols are processed by an N-sample inverse fast Fourier transform operation that transforms the data symbol sequence into the time domain. The time-domain signal is extended by a guard interval containing G sample cyclic extensions of the OFDM symbol, whose length is chosen to be longer than the expected delay spread to avoid ISI. The guard interval includes a cyclically extended part of the OFDM symbol to avoid ICI. Hence, the complete OFDM block duration is $P = N + G$ samples. The resulting signal is converted to an analog signal by a digital-to-analog converter. After shaping with a low-pass filter (e.g., a raised-cosine filter) with bandwidth BW, it is transmitted through the transmit antenna with the overall symbol duration of $T = PT_s$.

Let $h(m, l)$ represent the impulse response of the time-varying channel. The received signal in discrete time can then be expressed as follows:

$$y(m) = \sum_{l=0}^{L-1} h(m, l)d(m - l) + w(m), \tag{1}$$

where the transmitted signal $d(m)$ at discrete sampling time m is given by:

$$d(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k e^{j2\pi mk/N}. \tag{2}$$

L is the total number of paths of the frequency selective fading channel, and $w(m)$ is the additive white Gaussian noise with zero mean and variance $E\{|w(m)|^2\} = \sigma_w^2$. The sequence s_k, $k = 0, 1, ..., N - 1$, in Eq. (2) represents modulated data symbols with $E\{|s_k|^2\} = 1$.

At the receiver, after passing through the analog-to-digital converter and removing the guard interval or cyclic prefix (CP), a fast Fourier transform (FFT) is used to transform the data back into the frequency domain. Lastly, the binary data are obtained after demodulation.

The fading channel coefficients $h(m, l)$ can be modeled as zero-mean complex Gaussian random variables. Based on the wide-sense stationary uncorrelated scattering assumption, the fading channel coefficients in different paths are uncorrelated with each other. However, these coefficients are correlated within each individual path and have a Jakes' Doppler power spectral density as in [14] with an autocorrelation function given by:

$$E\{h(m, l)h^*(n, l)\} = \sigma_{h_l}^2 J_0(2\pi f_d T_s(m - n)), \tag{3}$$

where $\sigma_{h_l}^2$ denotes the power of the channel coefficients of the lth path. f_d is the Doppler frequency (Hz) so that the term $f_d T_s$ represents the normalized Doppler frequency of the channel coefficients. $J_0(.)$ is a zeroth-order Bessel function of the first kind.

By using Eq. (2) in Eq. (1), the received signal can be written as:

$$y(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k \sum_{l=0}^{L-1} h(m,l) e^{j\frac{2\pi k(m-l)}{N}} + w(m), \tag{4}$$

which upon defining the time-varying frequency response,

$$H_k(m) = \sum_{l=0}^{L-1} h(m,l) e^{-j2\pi lk/N}, \tag{5}$$

becomes:

$$y(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k H_k(m) e^{j2\pi mk/N} + w(m), \tag{6}$$

where k and m are the discrete frequency and time indices, respectively. The FFT output at the kth subcarrier can be expressed as:

$$Y_k = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} y(m) e^{-j2\pi mk/N} = s_k H_k + I_k + W_k, \tag{7}$$

where H_k represents the average frequency domain channel response, defined as I_k, which is ICI caused by the time-varying nature of the channel, given as:

$$H_k = \frac{1}{N} \sum_{m=0}^{N-1} H_k(m). \tag{8}$$

I_k is ICI caused by the time-varying nature of the channel, given as:

$$I_k = \frac{1}{N} \sum_{i=0, i\neq k}^{N-1} s(i) \sum_{m=0}^{N-1} H_i(m) e^{j2\pi m(i-k)/N}. \tag{9}$$

W_k denotes the discrete Fourier transform of white Gaussian noise $w(m)$:

$$W_k = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} w(m) e^{-j2\pi mk/N}. \tag{10}$$

The received signal, after excluding the guard interval, can be expressed in vector form as:

$$\mathbf{y} = \mathbf{Hs} + \mathbf{w}, \tag{11}$$

where $\mathbf{y} = [y(0), y(1), ..., y(N-1)]^T$, $\mathbf{s} = [s(0), s(1), ..., s(N-1)]^T$, $\mathbf{w} = [w(0), w(1), ..., w(N-1)]^T$, and the time-varying channel matrix $\mathbf{H}$ is given by Eq. (12), where the first and second indices of $H_k(.)$ represent the discrete frequency and time variables, respectively.

$$\mathbf{H} = \frac{1}{\sqrt{N}} \begin{bmatrix} H_0(0) & H_1(0) & \cdots & H_{N-1}(0) \\ H_0(1) & H_1(1)e^{j2\pi/N} & \cdots & H_{N-1}(1)e^{j2\pi(N-1)/N} \\ \vdots & \vdots & \cdots & \vdots \\ H_0(N-1) & H_1(N-1)e^{j2\pi(N-1)/N} & \cdots & H_{N-1}(N-1)e^{j2\pi(N-1)(N-1)/N} \end{bmatrix} \tag{12}$$

3. Two-dimensional pilot symbol-assisted channel estimation

OFDM demodulation requires an estimation of the fading channel parameters, and this can be provided by inserting pilots into the transmitted signal. Dedicated time slots (DTSs) and dedicated subcarriers (DSs) are well-known pilot patterns that are used in OFDM systems [15]. In the DS pattern, certain subcarriers are dedicated to pilot symbols, whereas in the DTS pattern, all subcarriers in a given time slot are dedicated to pilot symbols. In [13], the minimum mean square error (MMSE) channel estimator technique using DTSs is applied, where the number of subcarriers is taken as N = 32. Therefore, the DTS pilot pattern is sufficient to approximate the channel time variations with a piecewise linear model with a constant slope over the time duration of N = 32. However, increasing the number of subcarriers will increase the time duration of the OFDM symbol and the approximation of the channel time variations over a constant slope is no longer valid. This technique is called the dedicated time slot-based algorithm (DTSBA) and it is subsequently investigated here.

In this paper, we analyze 2D pilot symbol-assisted modulation, where pilot symbols are transmitted on every p subcarriers and in every OFDM symbol. Therefore, pilot symbols are spread throughout the time-frequency grid that is proposed in the DS pattern. The N_{PS} pilot symbols $\{s_k\}$ for $k \in S_{PS}$ in each OFDM symbol are inserted as pilot symbols known by the receiver. Estimation of the average frequency-domain channel response for a subcarrier of the OFDM symbol by using pilot symbols can be written as follows:

$$\hat{H}_k = \frac{Y_k}{s_k} = H_k + \frac{I_k + W_k}{s_k}. \tag{13}$$

To estimate the channel matrix in Eq. (12), the missing channel information in both the frequency and time domains has to be estimated. Optimal techniques such as Wiener filtering need the correlation of the time-varying coefficients. Moreover, they are computationally intensive and incur huge system delays [12]. Therefore, suboptimal techniques such as piecewise constant interpolation, low-pass interpolation, and curve fitting can be applied to find the 2D frequency response of the wireless channel.

3.1. Frequency-domain interpolation

It was shown that low-pass interpolation-based channel estimation performs the best among all interpolation-based channel algorithms [11] for comb-type pilot-based algorithms. Therefore, corresponding to pilot symbols, all channel attenuations in the frequency domain are estimated by a low-pass interpolation filter.

To generate additional data points, the low-pass interpolation insert zeros into the original sequence. It then designs a special symmetric finite impulse response (FIR) filter that allows the original data to pass through unchanged and interpolates in between such that the mean square error between the interpolated points and their actual values is minimized. Finally, the input vector is applied to the FIR filter to yield the low-pass interpolated output vector [11].

3.2. Time-domain interpolation

The time-domain interpolation method is very important for fast OFDM systems because it affects the OFDM performance in the case of ICI. Therefore, firstly, we studied DTSBA, piecewise constant, piecewise linear, low-pass, and curve fitting-based interpolation.

Conventional channel estimators assume a time-invariant model for the channel during one OFDM symbol. Therefore, after finding channel attenuations in the frequency domain, we use these estimated values until the next OFDM symbol. This algorithm is labeled as "conventional" in the Figures. This technique can also be called piecewise constant interpolation. While it is the simplest interpolation method, it is demonstrated in the section on simulation that its performance is not sufficient to detect the OFDM symbols in high mobility.

To find the time variations of each subcarrier for each OFDM sample, low-pass interpolation can also be applied, similar to frequency-domain interpolation. However, the complexity of low-pass interpolation is very high when the number of interpolated data increases, as will be explained in the section on complexity. Therefore, in this paper, we propose curve fitting [16]. Missing information between consecutive OFDM symbols can be found by employing polynomial curve fitting based on the linear model estimator.

The proposed algorithm is labeled as "proposed" in the Figures. In the section on simulation, we show that low-pass interpolation and curve fitting have similar performances for estimating the time variations of each subcarrier, while curve fitting has lower complexity.

3.2.1. Curve-fitting algorithm

We assume that the OFDM frame consists of B number of OFDM symbols. An acceptable model for a subcarrier can be written as follows:

$$\hat{H}_k(t_b) = \theta_{1,k} + \theta_{2,k}t_b + ... + \theta_{p,k}t_b^{p-1} + w_k(t_b), \tag{14}$$

where t_b for $b = 0, N, ..., N(B-1)$ denotes OFDM samples if we want to fit a (p − 1)th-order polynomial to the experimental data. To check availability of the benefits of the linear model, we assume that $w_k(t_b)$ are independent and identically distributed Gaussian variables with zero mean and variance σ^2, or that they are white Gaussian noise samples. We then have the usual linear model form:

$$\hat{\mathbf{H}}_k = \mathbf{T}\theta + \mathbf{w}, \tag{15}$$

where $\widehat{\mathbf{H}}_k = \left[\hat{H}_k(t_0), \hat{H}_k(t_1), ...\hat{H}_k(t_{N(B-1)})\right]^T$, $\theta = [\theta_1 \theta_2 ... \theta_p]^T$, and

$$\mathbf{T} = \begin{bmatrix} 1 & t_o & \cdots & t_o^{p-1} \\ 1 & t_1 & \cdots & t_1^{p-1} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & t_{B-1} & \cdots & t_{B-1}^{p-1} \end{bmatrix}.$$

The minimum variance unbiased estimator for θ is as follows:

$$\hat{\theta} = \left(\mathbf{T}^T\mathbf{T}\right)^{-1}\mathbf{T}^T\widehat{\mathbf{H}}_k. \tag{16}$$

The observation matrix has the special form of a Vandermonde matrix. The resulting curve fit is:

$$\hat{H}_k(m) = \sum_{i=1}^{p}\hat{\theta}_i m^{i-1}, \quad m = 0, 1, ..., NB, \tag{17}$$

where $H_k(m)$ represents the underlying curve or signal. Eq. (17) represents the time variations of a subcarrier for an OFDM frame. Estimation of the time-varying channel matrix, $\hat{\mathbf{H}}$, in Eq. (12) is done by calculating $\hat{H}_k(m)$ for all subcarriers, and then it is used for detection.

The first-order polynomial fitting corresponds to piecewise linear approximation [14]. It will be shown that piecewise linear approximation is not adequate to approximate the variations of the channel for high-mobility cases in the section on simulation.

3.2.2. Computational complexity of the curve-fitting algorithm

While frequency-domain interpolation needs to interpolate data for the number of pilot spacings, time-domain interpolation needs to interpolate data for the total number of OFDM samples in an OFDM frame. The number of the coefficients of the low-pass interpolation filter is proportional to the number of data that will be interpolated. For example, if we resample the sequence in the vector 8 and 256 times, the number of the coefficients of the low-pass interpolation filter will be equal to 67 and 2049.

Therefore, the number of coefficients of the low-pass interpolation filter may be approximated as $8N$. In this case, the low-pass filter has $8N$ complex multiplications for each sample of a subcarrier that requires $8BN^2$ in total. Finally, it requires $8BN^3$ for all subcarriers.

According to Eq. (15), $\mathbf{T}^T\mathbf{T}$ and $(\mathbf{T}^T\mathbf{T})^{-1}$ require p^2B and $O(p^3)$ complex multiplications,

respectively. Multiplication of the $\mathbf{T}^T$ matrix with $\widehat{\mathbf{H}}_k$ has pB complex multiplications, and then multiplication with $(\mathbf{T}^T\mathbf{T})^{-1}$ has p^2 complex multiplications. Eq. (16) requires a total of $p^2+p^2B+pB+O(p^3)$ complex multiplications. According to Eq. (17), for a subcarrier we need $BN(p^2+p)/2$ complex multiplications. Finally, the total complexity of the curve fitting is equal to $\frac{BN^2(p^2+p)}{2}+BN(p^2+p)+N(O(p^3)+p^2)$ for all subcarriers.

4. Simulation results

Modern wireless communication systems require high data rates or a more spectrally efficient system. Higher data rates can be achieved in each subcarrier by using amplitude levels as well as phase states. In this section, the performance of a 16-quadrature amplitude modulation (16QAM) OFDM system operating over fast-frequency selective channels is investigated using computer simulations. The system operates with a bandwidth of 5 MHz and is divided into 256 tones ($N = 256$) with a total symbol period of 57 μs, of which 6 μs constitutes the CP. One OFDM symbol thus consists of 286 samples ($N + G = 286$), 30 of which constitute the CP. The wireless channel between the mobile antenna and the receiver antenna is modeled based on a realistic channel model determined by the COST-207 project in which the rural area (RA) channel model is considered to have the channel length $L = 4$. The normalized Doppler frequencies are $f_{d1}T_s = 0.015$ and $f_{d2}T_s = 0.030$, corresponding to an IEEE 802.16e mobile terminal moving with speeds of 120 km/h and 240 km/h for a carrier frequency of 2.4 GHz, respectively. The pilot insertion rate is chosen as 1:8, for 1 pilot inserted for every 8 data symbols.

We investigated the symbol error rate (SER) performance of least squares (LS) or zero forcing (ZF) and minimum mean square error (MMSE) detection methods. In Figures 1 and 2, the SERs of LS and MMSE detection are shown as functions of the signal-to-noise ratio (SNR) when the mobile terminal is moving with a speed of 120 km/h. Results are shown for 4 different estimation algorithms. It was observed that comb-type estimators significantly outperformed DTSBA for all detection techniques. Moreover, it was shown that 2D low-pass interpolation and the proposed interpolation have similar performances, while the proposed technique has very low computational complexity.

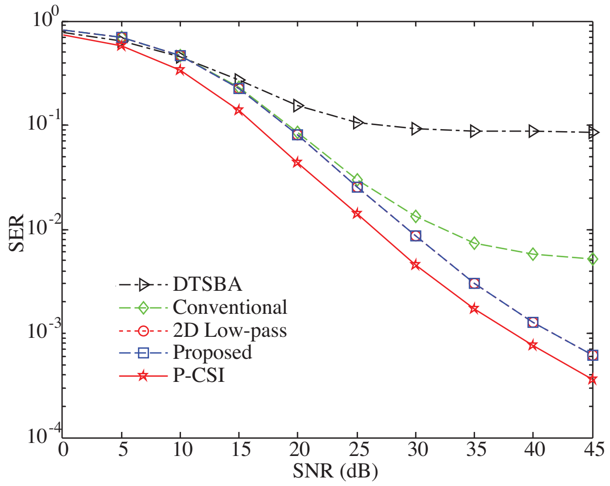

Figure 1. SER comparison of different estimation algorithms for OFDM systems in the case of LS detection (16QAM, 120 km/h).

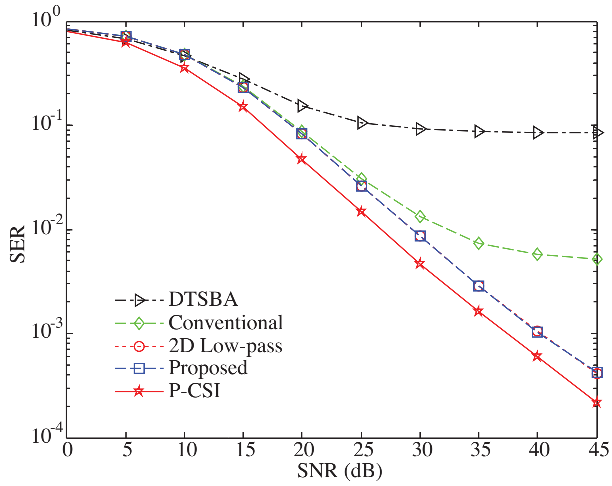

Figure 2. SER comparison of different estimation algorithms for OFDM systems in the case of MMSE detection (16QAM, 120 km/h).

The noise enhancement effect of the LS equalizer can be reduced by using the MMSE equalizer. In particular, it was observed that savings of about 2.5 dB were obtained at $SER = 10^{-2}$ as compared with conventional detection for the RA channel with MMSE detection (Figure 2).

To investigate the performance of channel estimation algorithms for high mobility, we increased the speed of the mobile terminal to 240 km/h. The performance difference is more obvious for high mobility in Figures 3 and 4. In Figure 4, it is also shown that the conventional channel estimator has an error floor. In particular, the conventional estimator has an error floor above 35 dB for 240 km/h, while it is above 45 dB for 120 km/h.

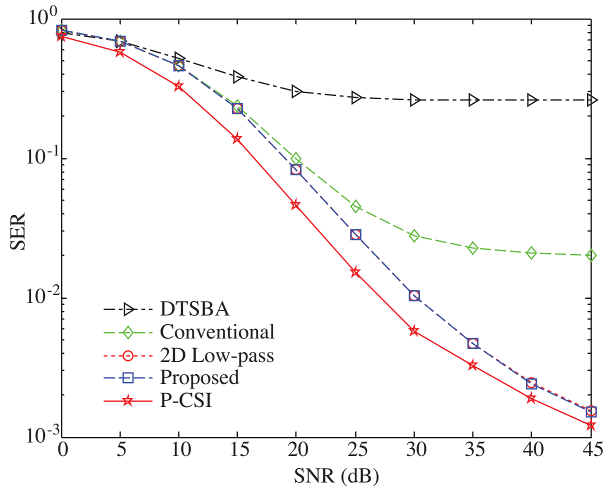

Figure 3. SER comparison of different estimation algorithms for OFDM systems in the case of LS detection (16QAM, 240 km/h).

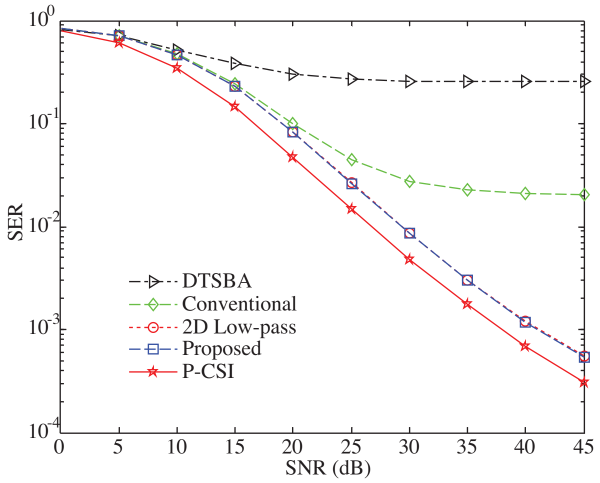

Figure 4. SER comparison of different estimation algorithms for OFDM systems in the case of MMSE detection (16QAM, 240 km/h).

In Figures 5 and 6, the order of the polynomial for MMSE detection for quadrature phase shift keying (QPSK) and 16QAM is shown. It can be concluded from these curves that the selection of the p value is highly dependent on Doppler values and SNR values. In particular, different p values show similar performance for 120 km/h and 240 km/h for SNRs below 35 and 25 dB, respectively, because the effect of ICI is not very obvious relative to noise effects. In Figure 5, it is shown that for the speeds below 120 km/h, a first-order polynomial (also denoted as piecewise linear approximation) is sufficient [14]. In other words, piecewise linear interpolation can be used to approximate variations of the channel for low mobility. For higher mobility values, it is shown in Figure 6 that piecewise linear approximation falls short of expectations for high mobility. It is concluded that we should increase the polynomial order to 2 for the speed of 240 km/h.

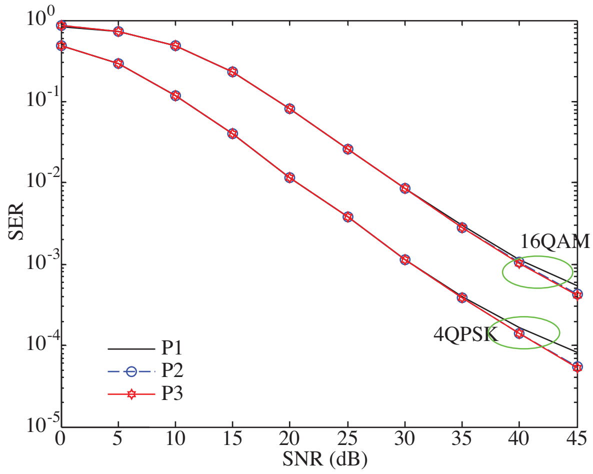

Figure 5. Comparison of polynomial orders of 16QAM and QPSK in the case of MMSE detection (120 km/h).

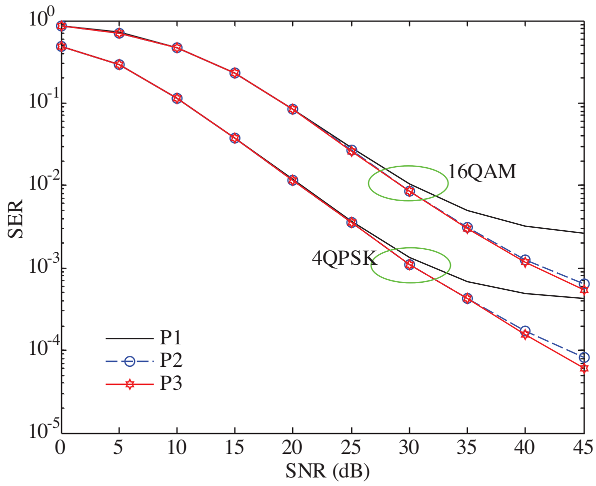

Figure 6. Comparison of polynomial orders of 16QAM and QPSK in the case of MMSE detection (240 km/h).

5. Conclusion

In this paper, we proposed a simple and low-complexity approach for the estimation of fast-varying OFDM channels. It was shown that the proposed techniques outperform conventional methods as well as other estimation methods based on DTSs, while it has performance similar to that of a 2D low-pass interpolator. It was also demonstrated that channel estimation performance is obvious for MMSE detection as compared to LS detection. Therefore, we conclude that the detection method used at the receiver has an important role in improving the system performance. Moreover, the effect of polynomial order was investigated, and it was shown that piecewise linear approximation falls short of expectations for high mobility. Therefore, it was concluded that the polynomial-fit approach yielded the best performance in our channel estimation approach for high-mobility OFDM systems.

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