Challenge of Channel Estimations and Its Way Out in MIMO OFDM Systems for Mobile Wireless Channels

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Abstract: - In current common channel estimation schemes for MIMO OFDM systems, channel state information is usually achieved by estimating channels according to frequency domain pilot sequences. However, when the length of MIMO OFDM symbols is larger than that of wireless channel delay, there are two intractable issues in the case of cellular fast fading channel scenarios with large numbers of users, i.e., the bandwidth overhead of channel estimation and the difficulty to construct large numbers of orthogonal training sequences. Inspired from the Steiner channel estimation method in multi-user CDMA uplink wireless channels, we proposed a new design scheme of training sequence in time domain to conduct channel estimation in MIMO OFDM systems. According to the proposed schemes, training sequences of different transmit antennas can be simply obtained by truncating the circular extension of one basic training sequence, and the pilot matrix assembled by these training sequences is one circular matrix with good reversibility. Furthermore, when the length of channel profiles is less than that of MIMO OFDM symbols, more bandwidth resources can be saved, as the training sequence only occupies a part of MIMO OFDM symbols. At last, we analyze the classical common time and frequency domain schemes in terms of channel overheads and computation complexity, and the results disclose the proposed scheme can be one way out for current common frequency domain channel estimation issues. For typical ITU indoor, pedestrian and vehicular channel profiles, the corresponding numerical results indicate the proposed method can save abundant bandwidth and achieve good channel estimation accuracy when compared with classical frequency and time domain approaches, respectively.

Key-Words: - channel estimation, wireless communications, MIMO, OFDM, pilot sequences

1 Introduction

The wireless communication system [1-5] coupled with multiple transmit/receive antennas and orthogonal frequency-division multiplexing (OFDM), is regarded as a promising solution for enhancing the data rates of next-generation wireless communication systems operating in frequency-selective fading environments. Channel parameters provide key information for the operation of wireless systems and need to be estimated accurately. So, many training-based MIMO OFDM channel estimation methods have been widely studied [6-9], which could be put into two categories such as frequency domain [8-9] and time domain [6-7] approaches. Nevertheless, for the scenarios with large numbers of users in cellular fast fading channels, there are two difficult problems [10-12] that have to be figured out, i.e., the challenge to construct large numbers of orthogonal training sequences and the bandwidth overhead of channel estimation when the length of MIMO OFDM symbols is larger than that of wireless channel delay. So, it's necessary to find a new time approach to overcome these drawbacks.

In fact, spatial channels for every receive antenna in MIMO links can be considered as multiple-input single-out (MISO) channels, i.e., equivalents of links in multi-user CDMA uplinks. So, inspired from the Steiner channel estimation in uplink CDMA wireless links [13-14], we propose one new scheme for MIMO OFDM systems with mobile wireless channels. According to the scheme, training sequences of different transmit antennas can be simply obtained by truncating the circular extension of one basic training sequence, and the pilot matrix assembled by these training sequences is one circular matrix with good reversibility. Furthermore, the high dimensional matrix inversion in typical time domain schemes can be avoided via the diagonalization of pilot matrix with unitary DFT matrices [15]. Furthermore, when the length of channel profiles is less than that of MIMO OFDM symbols, more bandwidth resources can be saved, as the training sequence only occupies a part of MIMO OFDM symbols. Channel information is firstly estimated in time domain, and then its frequency versions at different sub-carriers are obtained by its Fourier transformation.

The remainder of this paper is organized as following. In Section 2, we briefly introduce system model of MIMO OFDM systems, and the proposed design method is detailed in Sections 3. Then, we formulate the proposed channel estimation scheme and its performance analysis. Finally, numerical results and some conclusions are presented in Section 5 and Section 6, respectively.

2 System Model

For one MIMO-OFDM system with M_t transmits and M_r receives antennas, one OFDM modulator is employed on each transmit antenna. Typically for any multi-carrier modulation scheme, one data stream is first divided into N parallel sub-streams. Then, the k-th sub-stream of the n-th symbol block transmitted from the v-th antenna is denoted by $X^{v}_{n,k}$. An inverse DFT with N points is performed on each block, and a guard interval (GI) with N_{GI} samples is inserted in the form of a cyclic prefix. Subsequently, these data are transmitted over spatial multi-path fading channels. At the u-th receiver, the guard interval is removed from the received data symbols and then DFT is followed to obtain the estimation of transmitted data symbols. The received data consists of superimposed data from M_t transmit antennas. We assume carriers to be kept orthogonal by cyclic prefix and channels to be constant over one OFDM symbol. Then, the received signal at the k-th sub-channel of the v-th receive antenna for the n-th OFDM symbol is given by

$$Y^{v}_{n,k} = \sum_{u=1}^{M} X^{u}_{n,k} H^{(v,u)}_{n,k} + W^{v}_{n,k} \tag{1}$$

where $X^{u}_{n,k}$, $H^{(v,u)}_{n,k}$ and $W^{v}_{n,k}$ respectively denote the transmitted symbols at the k-th carrier of the u-th transmit antenna for the n-th OFDM symbol, the channel fading coefficient at the k-th carrier of the spatial channel between the v-th receive antenna and the u-th transmit antenna, and the additive white Gaussian noise (AWGN) with zero mean and variance σ^2.

We consider a time-variant Rayleigh fading channel with frequency selectivity, which is modeled by one tapped delay line with L_h nonzero taps [15]. The frequency fading coefficients for different carriers could be obtained via the Fourier transform of channel impulse response (CIR) for different transmit / receive antenna pairs. So, the coefficient $H^{(v,u)}_{n,k}$ is described by

$$H^{(v,u)}_{n,k} = \sum_{l=0}^{L_h-1} h^{(v,u)}_{n,l} e^{-j2\pi\tau^{(v,u)}_{l} l/T} \tag{2}$$

where *1/T* is sub-carrier spacing, channel gain $h^{(v,u)}_{n,l}$ of the l-th tap at time delay $\tau^{(v,u)}_{l}$ is a wide sense stationary (WSS), complex Gaussian random variable with zero mean. The channel taps of the pairs between transmit and receive antennas, are assumed to be mutually uncorrelated. Due to motion of the vehicle, $h^{(v,u)}_{n,l}$ is time-variant and band-limited by the maximum Doppler frequency v_{max}.

3 Design Training Sequences

Denote L_m as length of training sequence, W as length of spatial CIRs, and L = W*M_t as length of a basic sequence, which can be delineated as

$$\mathbf{m} = \left(m_1, m_2, ..., m_L\right) \tag{3}$$

Its circular extension version is given by

$$\bar{\mathbf{m}} = \left(\bar{m}_1, \bar{m}_2, ..., \bar{m}_{L_m+(M_t-1)W}\right) \tag{4}$$

where the first L elements are consistent with corresponding elements of the basic sequence and other elements are determined by

$$\bar{m}_i = m_{i-L},\ i = \left(L+1\right), \cdots, \left[L_m + \left(M_t - 1\right)W\right] \tag{5}$$

Then, training sequences for different transmit antennas are obtained by truncating the circular extended version as showed in Fig.1. Furthermore, the pilot for the u-th transmit antenna is presented as

$$\mathbf{m}^{(u)} = \left(m^{(u)}_1, m^{(u)}_2, ..., m^{(u)}_{L_m}\right) \tag{6}$$

where

$$m^{(u)}_i = m_{i+(M_t-1)W},\ i = 1, ..., L_m,\ u = 1, ..., M_t \tag{7}$$

If the design parameters showed above can be denoted as a quaternion (L_m,L,M_t,W), there exists the following relationship among these parameters, i.e.,

$$W = \left\lfloor \frac{L_m}{M_t + 1} \right\rfloor,\ L = WM_t \tag{8}$$

where operator $\lfloor \cdot \rfloor$ denotes the largest integer not more than a given real number in the operator.

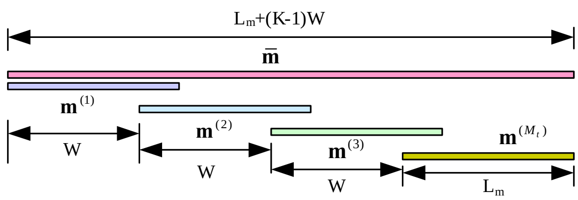

Fig.1. Construction of training sequences from a circular extended sequence, where the training sequence for every transmit antenna is obtained via truncating a circular extended sequence according to its position in the circular extended sequence.

However, the training sequences given by (6) are generally described as binary sequences, and should be converted into complex number. Firstly, they are re-presented as bi-polar sequences and then further converted into complex numbers.

Denote $\vec{\mathbf{m}}^{(u)}$ as one bi-polar sequence, $\mathbf{m}_c^{(u)}$ as its correspondent complex form, which can be determined by

$$m_c^{(u)}(i) = (j)^i \cdot \vec{m}^{(u)}(i),\ i = 1,\dots,L \tag{9}$$

Where j is unit of imaginary number.

3 The Proposed Scheme

For the considering MIMO OFDM systems, one receive antenna must estimate all the spatial channels between the receive antenna and all the transmit antennas, simultaneously. Then, the channel information estimated by all receive antennas will be assembled together to coherently decode data symbols carried by one MIMO OFDM symbol. Therefore, the following algorithm will be described for one receive antenna, the same counterparts can be easily applied to all the receive antennas.

Provided that all spatial channels for MIMO radio links have the same delay profiles, the channel impulse response between the *u*-th transmit antenna and the receive antenna, is described as

$$\mathbf{h}^{(u)} = \left(h_1^{(u)}, h_2^{(u)}, \cdots, h_W^{(u)}\right)^T,\ u = 1, 2, \cdots, M_t \tag{10}$$

Furthermore, training sequence transmitted by the *u*-th transmit antenna is given by

$$\mathbf{m}^{(u)} = \left(m_1^{(u)}, m_2^{(u)}, \cdots, m_{L+W-1}^{(u)}\right)^T \tag{11}$$

Which is obtained by truncating the first *L+W-1* elements from designed training sequence as showed above.

After spatial channel filtering, its received version is given by

$$\tilde{\mathbf{e}}^{(u)} = \tilde{\mathbf{G}}^{(u)}\mathbf{h}^{(u)} + \tilde{\mathbf{n}}^{(u)},\ u = 1, 2, \cdots, M_t \tag{12}$$

where $\tilde{\mathbf{e}}^{(u)}$ and $\tilde{\mathbf{n}}^{(u)}$ denote received training symbols and correspondent noise vector, respectively, and $\tilde{\mathbf{G}}^{(u)}$ is the pilot symbol matrix assembled by training sequence from the *u*-th transmit antenna, which can be delineated as (13).

According to linear convolution theory [15], the received training symbols are the polluted version of training symbols transmitted by the *u*-th transmit antenna. Hence, their first *W-1* elements and last *W-1* elements will be discarded away as they are interfered by the transmit signals before and after pilot sequence, respectively. So, the usable received data is given by (12).

$$\tilde{\mathbf{G}}^{(u)} = \begin{bmatrix} m_1^{(u)} & & & \\ m_2^{(u)} & m_1^{(u)} & & \\ \vdots & \vdots & \ddots & \\ m_W^{(u)} & m_{W-1}^{(u)} & \cdots & m_1^{(u)} \\ \vdots & m_W^{(u)} & & m_2^{(u)} \\ \vdots & \vdots & & \vdots \\ m_{W+L-1}^{(u)} & m_{W+L-2}^{(u)} & \cdots & m_L^{(u)} \\ & m_{W+L-1}^{(u)} & & \vdots \\ & & \ddots & \vdots \\ & & & m_{W+L-1}^{(u)} \end{bmatrix} \tag{13}$$

$$\mathbf{e}^{(u)} = \left(\tilde{e}_W^{(u)}, \tilde{e}_W^{(u)}, \cdots, \tilde{e}_{W+L-1}^{(u)}\right)^T \tag{14}$$

At the same time, let $\mathbf{n}^{(u)} = \left(\tilde{n}_W^{(u)}, \tilde{n}_{W+1}^{(u)}, \cdots, \tilde{n}_{W+L-1}^{(u)}\right)^T$, we could obtain following relationship based on formula (12)

$$\mathbf{e}^{(u)} = \mathbf{G}^{(u)}\mathbf{h}^{(u)} + \mathbf{n}^{(u)} \tag{15}$$

Where $\mathbf{G}^{(u)}$ is one sub-matrix consisting of all the rows between the *W*-th row and (*W+L-1)*-th row of pilot matrix $\tilde{\mathbf{G}}^{(u)}$, i.e.,

$$\mathbf{G}^{(u)} = \begin{bmatrix} m_W^{(u)} & m_{W-1}^{(u)} & \cdots & m_1^{(u)} \\ \vdots & m_W^{(u)} & & m_2^{(u)} \\ \vdots & \vdots & & \vdots \\ m_{W+L-1}^{(u)} & m_{W+L-2}^{(u)} & \cdots & m_L^{(u)} \end{bmatrix} \tag{16}$$

In fact, the received data at one receive antennas is the superimposed data from all transmit antennas, which can be described as

$$\mathbf{e} = \sum_{u=1}^{M_t} \mathbf{e}^{(u)} \tag{17}$$

Which could be further expressed in matrix form according to (15), i.e.,

$$\mathbf{e} = \mathbf{G}\mathbf{h} + \mathbf{n} \tag{18}$$

Simultaneously, other terms in (18) expression are given by

$$\mathbf{G} = \left(G^{(M_t)}, G^{(M_t-1)}, \cdots, G^{(1)}\right)\ ;\ \mathbf{n} = \sum_{u=1}^{M_t}\mathbf{n}^{(u)}\ ;$$

$$\mathbf{h} = \left(\mathbf{h}^{(M_t)T}, \mathbf{h}^{(M_t-1)T}, \cdots, \mathbf{h}^{(1)T}\right)^T \tag{19}$$

Assume the noise vector in (18) is a wide-sense stationary (WSS) complex Gaussian vector with zero means and covariance matrix given by $\mathbf{R}_n = \sigma^2\mathbf{I}$, where σ^2 is noise variance and $\mathbf{I}$ is n-order unit matrix. Then, according to (18), the spatial channels between all transmit antennas and the receive antenna, could be estimated by

$$\hat{\mathbf{h}} = \left(\mathbf{G}^H\mathbf{G}\right)^{-1}\mathbf{G}^H\mathbf{e} \quad (20)$$

If **G** is invertible, the above expression is further rewritten as

$$\hat{\mathbf{h}} = \mathbf{h} + \mathbf{G}^{-1}\mathbf{n} \quad (21)$$

Moreover, according to (16) and (19), **G** is actually one L-order circular matrix, which could be diagonalized by one unitary DFT matrix [15], that is

$$\mathbf{G} = \mathbf{F}\boldsymbol{\Delta}\mathbf{F}^H \quad (22)$$

where **F** is a L-order DFT matrix, **Δ** is a L-order diagonal matrix whose elements are given by the DFT of the basic training sequence.

Subsequently, substitute (22) into (20), then we will get

$$\hat{\mathbf{h}} = \mathbf{F}\boldsymbol{\Delta}^{-1}\mathbf{F}^H\mathbf{e} \quad (23)$$

where operator $\mathbf{F}(\cdot)$ and $\mathbf{F}^H(\cdot)$ could be explained as performing DFT and inverse discrete fourier transformation (IDFT) to one vector, respectively. So, we can rewrite (23) into an alternative form as

$$\hat{\mathbf{h}} = dft\left[dft\left(\mathbf{e}\right)./\,idft\left(\tilde{\mathbf{m}}\right)\right] \quad (24)$$

Where operator $dft(\cdot)$ and $idft(\cdot)$ denote to perform DFT and IDFT on a vector, respectively, while $(\cdot)./(\cdot)$ denotes array right division operator in element-wise. Note that $\tilde{\mathbf{m}}$ is the reverse version of basic sequence **m**, the result spatial channel are also estimated in the reverse order.

4 Overhead and Complexity

Here, we use the appropriator intervals as bandwidth overhead of channel estimation, which are quantified into sample periods in corresponding MIMO OFDM system configurations, and the number of multiplication as metric of complexity for channel estimation. The bandwidth overhead and complexity for different channel estimation approaches are analyzed and compared with each other in the followings.

For the frequency approaches [8-9], pilot sequences transmitted by every transmit antenna are designed to be orthogonal to each other and carried by M_t OFDM symbols at least, and the corresponding pilot patterns can be showed in Fig.2. Following these schemes, M_t times N-IFFT for OFDM modulation, M_t times N-FFT for OFDM demodulation and M_t*N times division are needed to estimate all the spatial channels between all the transmit antennas and the receive antenna.

However, M_t*L samples must be observed at one receive antenna to estimate the M_t spatial channel impulse responses in the time approaches [6-7]. The training sequences transmitted at different antennas only occupy M_t*L samples. Firstly, one M_t *by* L matrix inversion is involved to estimate all the CIRs of M_t spatial channels, and then these spatial CIRs must be further converted into frequency domain via M_t times N-FFT transforms to obtain corresponding fading coefficients at different carriers. As a result, M_t times N-FFT and a $Mt*L$ dimension matrix inversion calculation are used to obtain these MIMO OFDM channel coefficients. Without matrix inversion, it will take a M_t*L point IFFT, two M_t*L point FFT and M_t times N-FFT for the proposed scheme to obtain all MIMO OFDM sub-channel coefficients with the same bandwidth overhead as the time approaches.

Their bandwidth overhead and complexity are delineated in Table 1, where the proposed approach is indicated to have smaller bandwidth overhead and complexity when compared with frequency and time approaches, respectively. Thus, the scheme can make good tradeoff between classical frequency approaches and classical time approaches with respect to complexity and bandwidth overhead.

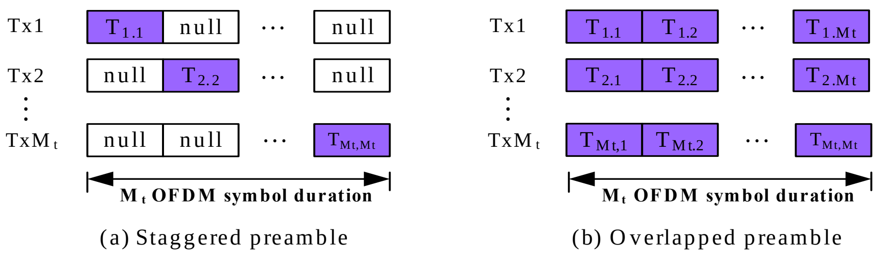

Fig.2 Two typical pilot patterns for channel estimation in MIMO OFDM systems

Table.1 Overhead and complexity for proposed scheme, classical frequency and time approaches

Scheme candidate	Bandwidth Overhead (sample period)	Complexity (multiplication operations)
Frequency Scheme	$5M_tN/4$	$2M_t\, N(\log_2N+1)$
Time Scheme	M_tL	$M_tN\log_2N+(M_t*L)^{2.37}$
Proposed Scheme	M_tL	$3M_tL\log_2L+M_tN\log_2N$

5 Numerical Results

In order to evaluate the performance of the proposed scheme, we consider one MIMO OFDM system with 2048 carriers at carrier frequency of 2.4 GHz. It has 20MHz bandwidth with 1/4 OFDM word symbols as guard intervals, which can eliminate inter-symbol interference (ISI) caused by frequency selective channels. When the transceivers are equipped with 8x2 antennas as base stations (BS) and mobile stations (MS) respectively and system sample period is given as 48.828125ns, we figure out the corresponding bandwidth occupancy for the classical time and frequency channel estimation approaches under the case of the typical cellular wireless channel profiles with different maximal delay.

Following the design of TD-SCDMA systems [16], the channel estimation interval for considered MIMO OFDM systems is given as an fixed value of 675us. In case of down links with 8 transmit antennas, the bandwidth occupancy of time approaches is calculated to 0.3674%, 0.4859% and 2.9748% for typical ITU indoor, pedestrian and vehicular channel profiles with maximal delay of 310ns, 410ns and 2510ns respectivley. However, the result for frequency approaches would be 145.64%, and that means the interval is not sufficient to estimate channel information in down link channels. The bandwidth occupancy for classical frequency approaches is 36.41% in up links with two transmit antennas. For the time domain approaches in up links with two transmit antennas, it is only one quarter of that in down links, while it is unchanged for frequency methods. Accoridng to the Table 1, the bandwidth occupancy in time domains is increased linearly with the r.m.s of channel profiles, and will not exceed its equivalent in frequency domains when the length of channel profiles is less than that of one MIMO OFDM word symbol. This results indicate that the channel estimation schemes in time domain can save a lot of bandwidth resouces when compared with its versions in frequency domains.

Furthermore, as showed in the analysis of computational complexity for the proposed method and the typical time and frequency approaches, their computational complexity can be numercial as followings. For down links of current MIMO OFDM systems, 1.9126e+005, 2.0163e+005 and 1.7484e+006 multiplication operations are involved in classical time domain approaches for typical ITU indoor, pedestrian and vehicular channel profiles with maximal delay of 310ns, 410ns and 2510ns respectivley, while 4.5469e+004, 4.5857e+004 and 1.0374e+005 operations for up links. The results of the proposed methods, are given as 1.8063e+005, 1.8084e+005 and 1.8724e+005 operations for down links, and 4.5158e+004, 4.5211e+004, 4.6809e+004 operations for down links. However, the computational costs of classical frequency schemes are 3.93216e+005 and 9.8304e+004 multiplication operations for down and up links respectively.

As listed in Table.2, numerical results of bandwidth overhead and computational complexity, indicate that the proposed approach cosume consume less channel resources with less complexity than the classical time and frequency methods, when typical cellular wireless channel scenarios are considered.

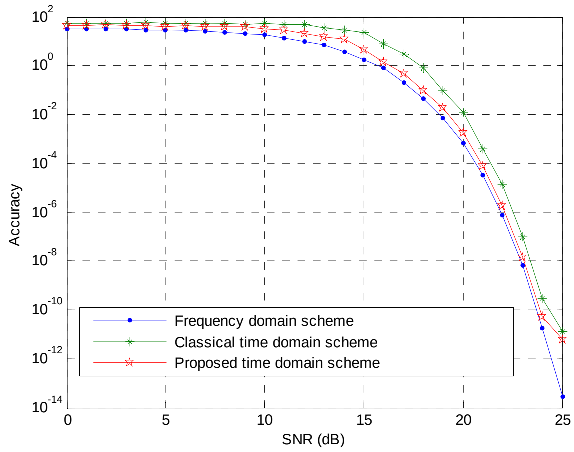

Fig.3 Accuracy of the proposed scheme, the classical time and frequency domain approaches in down links of MIMO OFDM with 8 transmit antennas for typical ITU vehicle profiles at 200Hz Doppler frequency.

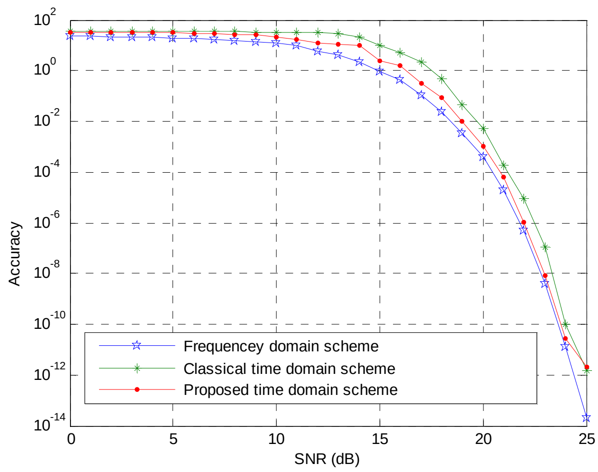

Fig.4 Accuracy of the proposed scheme, the classical time and frequency domain approaches in up links of MIMO OFDM with 2 transmit antennas for typical ITU vehicle profiles at 200Hz Doppler frequency.

Table.2 Channel estimation cost of the proposed method, the classical time and frequency approaches in the scenarios of MIMO OFDM systems with 8x2 antennas configuration at BS and MS, where the typical ITU indoor, pedestrian and vehicular channel profiles are used to numerically evaluate channel estimation cost

Channel Estimation Cost			Time Scheme	Frequency Scheme	Proposed Scheme
Bandwdith occupancy(%)	Down links	Indoor	0.3674	145.6	0.3674
		pedestrian	0.4859	145.6	0.4859
		vehicular	2.9748	145.6	2.9748
	Up links	Indoor	0.0919	36.41	0.0919
		pedestrian	0.1215	36.41	0.1215
		vehicular	0.7437	36.41	0.7437
Computational Complexity (operations)	Down links	Indoor	1.9126e+005	3.93216e+005	1.8063e+005
		pedestrian	2.0163e+005	3.93216e+005	1.8084e+005
		vehicular	1.7484e+006	3.93216e+005	1.8724e+005
	Up links	Indoor	4.5469e+004	9.8304e+004	4.5158e+004
		pedestrian	4.5857e+004	9.8304e+004	4.5211e+004
		vehicular	1.0374e+005	9.8304e+004	4.6809e+004

Subseuqently, their channel estimation accuracys with the same system configuration parameters, are further validated the typical ITU vehicle channel profiles with maximal doppler frequency of 200Hz, where the 2-norm of spatial channel matrix is used as the accuracy metric. Here, pseudo noise (PN) sequences are used as training sequences for different transmit antennas in classical time domain schemes, and orthogonal Hadamard sequences as frequency pilot symbols. Fig.3 and Fig.4 show the numerically simulated results in down and up links respectively. Clearly, the proposed channel estimation approach can obtain better channel estimation accuracy than that of classical time approach, as the channel pilot matrix constructed by the proposed method has better reversibility than that by PN sequences. But its different performance from the classical frequency methods can be compensated by its simply implementation with less bandwidth overhead..

6 Conclusions

In this paper, aiming at the challenges in current common frequency domain channel estimation schemes for MIMO OFDM systems, we analyzed the typical time and frequency domain channel estimation schemes, and proposed one new channel estimation scheme. Inspired from the Steiner channel estimation scheme in multi-user CDMA uplink radios, the scheme is conducted in time domain with elaborately designed training sequences. For the mobile fast frequency selective channels, abundant bandwidth overhead can be saved when the maximum delay of channel profile is less than that of one MIMO OFDM word symbol. Furthermore, it can also achieve good channel estimation accuracy when compared with classical frequency and time domain approaches, respectively.

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