http://ansinet.com/itj

ITJ

ISSN 1812-5638

INFORMATION TECHNOLOGY JOURNAL

ANSInet
Asian Network for Scientific Information
308 Lasani Town, Sargodha Road, Faisalabad - Pakistan

Information Technology Journal 6 (8): 1238-1244, 2007
ISSN 1812-5638

A New Steiner Channel Estimation Method in MIMO OFDM Systems

[1]Lu Zhaogan, [1,2]Wang Liejun, [1]Zhang Taiyi and [1]Runping Yun
[1]Department of Information and Communication Engineering,
Xi'an Jiaotong University, Peoples's Republic of China
[2]School of Information Science and Engineering, Xinjiang University, Peoples's Republic of China

Abstract: The combination of Multiple-Input Multiple-Output (MIMO) signal processing with 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. However, when the length of MIMO OFDM symbols is larger than that of wireless channel delay, two intractable issues should be resolved before their applications in cellular fast fading channel scenarios with large numbers of users, i.e., the bandwidth overhead of channel estimation and the challenge to construct large numbers of orthogonal training sequences. So, a new design scheme of training sequence in time domain, is adopted to conduct channel estimation in MIMO OFDM systems, which works as a generations of Steiner method in multi-user CDMA uplink scenarios. 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 one MIMO OFDM symbol. Numerical results of bandwidth overheads and channel estimations, 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

INTRODUCTION

The wireless communication system (Hongwei, 2005) coupled with multiple transmit/receive antennas (Nanda *et al.*, 2005) and Orthogonal Frequency-Division Multiplexing (OFDM) (Zelst and Schenk, 2004), 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 (Ogawa *et al.*, 2004; Li and Wang, 2003; Stuber *et al.*, 2004; Shenghao and Yuping, 2004), which could be put into two categories such as frequency domain (Stuber *et al.*, 2004; Shenghao and Yuping, 2004) and time domain (Ogawa *et al.*, 2004; Li and Wang, 2003) approaches. Nevertheless, for the scenarios with large numbers of users in cellular fast fading channels, there are two difficult problems (Barhumi *et al.*, 2002; Jagannatham and Rao, 2006) 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., the equivalents of links in multi-user CDMA uplinks. So, we generalize the Steiner channel estimation in up CDMA wireless links (Steiner and Jung, 1994; Steiner and Baier, 1993) to estimate MIMO OFDM spatial channels. 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. Subsequently, high dimensional matrix inversion can be avoided via diagonalization of pilot matrix by unitary DFT matrices (Proakis, 1995). 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 one MIMO OFDM symbol. Channel information is firstly estimated in time domain and then its frequency version at different sub-carriers is obtained by its fourier transformation.

Corresponding Author: Lu Zhaogan, Department of Information and Communication Engineering, Xi'an Jiaotong University, Peoples's Republic of China

SYSTEM MODEL

For a MIMO-OFDM system with M_t transmits and M_r receive antennas, one OFDM modulator is employed on each transmit antenna. Firstly, a data stream is divided into N parallel sub-streams, typical for any multi-carrier modulation scheme. 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 form of 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 a 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 via 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, frequency selective, Rayleigh fading channel, modeled by a tapped delay line with L_h nonzero taps (Proakis, 1995). 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:

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

Where, 1/T is sub-carrier spacing, channel gain $h^{(v,u)}_{n,l}$ of the l-th tap at time delay $\tau_l^{(v,u)}$ 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 the motion of vehicles, $h^{(v,u)}_{n,l}$ will be time-variant and band-limited according to the maximum Doppler frequency v_{max}.

DESIGN TRAINING SEQUENCES

Denote L_m and W as the length of training sequence and spatial CIRs and $L = W*M_t$ is the length of one basic sequence, which can be delineated as:

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

Its circular extension version is given by:

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

Where, the first L elements are consistent with corresponding elements of the basic sequence and other elements are determined by:

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

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

$$m^{(u)} = (m_1^{(u)}, m_2^{(u)}, ..., m_{L_m}^{(u)}) \tag{6}$$

Where:

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

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.

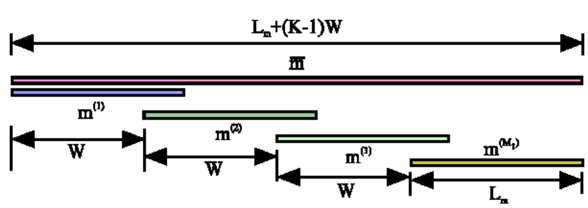

Fig. 1: Construction of training sequences from a circular extended sequence, where the training sequence for every transmit antenna is obtained via truncating a circular extended sequence according to its position in the circular extended sequence

However, the training sequences given by Eq. 6 are generally described as binary sequences and should be converted into complex number. Firstly, they are represented as bi-polar sequences and then further converted into complex numbers. Denote $\vec{m}^{(u)}$ as one bi-polar sequence, $m_c^{(u)}$ as its correspondent complex form, which can be determined by:

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

Where:
j = Unit of imaginary number.

STEINER CHANNEL ESTIMATION

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:

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

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

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

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:

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

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

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

According to linear convolution theory (Proakis, 1995), 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:

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

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

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

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

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

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

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

which could be further expressed in matrix form according to Eq. 15, i.e.,

$$e = Gh + n \tag{18}$$

Simultaneously, other terms in Eq. 18 expression are given by:

$$G = \left(G^{(M_t)}, G^{(M_t-1)}, \cdots, G^{(1)}\right);\ n = \sum_{u=1}^{M_t} n^{(u)};$$
$$h = \left(h^{(M_t)T}, h^{(M_t-1)T}, \cdots, h^{(1)T}\right)^T \tag{19}$$

Assume the noise vector in Eq. 18 is a Wide-Sense Stationary (WSS) complex Gaussian vector with zero means and covariance matrix given by $R_n = \sigma^2 I$, where σ^2 is noise variance and I is n-order unit matrix. Then, according to Eq. 18, the spatial channels between all transmit antennas and the receive antenna, could be estimated by:

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

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

$$\hat{h} = h + G^{-1} n \tag{21}$$

Moreover, according to (16) and (19), G is actually a L-order circular matrix, which could be diagonalized by an unitary DFT matrix (Proakis, 1995), that is:

$$G = F \Delta F^H \tag{22}$$

Where:

F = A L-order DFT matrix,

Δ = A L-order diagonal matrix whose elements are given by the DFT of the basic training sequence.

Subsequently, substitute Eq. 22 into 20, then we will get:

$$\hat{h} = F \Delta^{-1} F^H e \tag{23}$$

Where, operator $F(\cdot)$ and $F^H(\cdot)$ could be explained as performing DFT and Inverse Discrete Fourier Transformation (IDFT) to one vector, respectively. So, we can rewritten Eq. 23 into an alternative form as:

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

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

BANDWIDTH OVERHEAD AND COMPLEXITY

Here, we use the time intervals when channel estimations are conducted as bandwidth overhead of channel estimation, which are quantitated 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 (Stuber *et al*., 2004; Shenghao and Yuping, 2004), pilot sequences transmitted by every transmit antenna are designed to be orthogonal to each other and carried by at least M_t OFDM symbols 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 (Ogawa *et al*., 2004; Li and Wang, 2003). 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 M_t*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 Steiner 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 generalized Steiner

Table 1: Bandwidth overhead and complexity for Steiner scheme, classical frequency and time approaches, respectively

Scheme candidate	Bandwidth overhead (sample period)	Complexity (multiplication operations)
Frequency Scheme	$5M_tN/4$	$2M_t N(\log_2 N+1)$
Time Scheme	M_tL	$M_tN\log_2 N+(M_t*L)^{2.37}$
Steiner Scheme	M_tL	$3M_tL\log_2 L+M_tN\log_2 N$

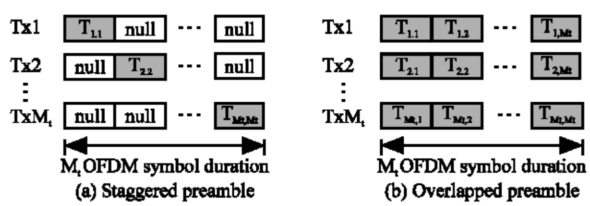

Fig. 2: Two typical pilot patterns for channel estimation in MIMO OFDM systems

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.

NUMERICAL RESULTS

In order to evaluate the performance of the proposed Steiner channel estimation scheme, we consider a MIMO OFDM system with 2048 carriers at carrier frequency of 2.4 GHz, which has 20 MHz bandwidth and a 1/4 OFDM symbols as guard intervals, which can eliminate intersymbol interference (ISI) caused by frequency selective channels. When the transceivers are equipped with 8×2 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 (Kan Zheng *et al*., 2005), the interval of channel estimation 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 310, 410 and 2510 ns, respectivley, while the result for frequency approaches would be 145.64%, i.e., the interval is not sufficient to estimate channel information in down links. However, the bandwidth occupancy for classical frequency approaches is 36.41% in up links with 2 transmit antennas. The bandwidth occupancy of time domain approaches in up links with 2 transmit antennas, is only one quarter of that in down links, but the result for frequency methods is kept unchanged. 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 if the length of channel profiles is still less than that of MIMO OFDM symbols. This results indicate that the channel estimation schemes in time domain can save a lot of bandwidth resouces when compared with its equivalents in frequency domains.

Furthermore, as showed in the analysis of computational complexity for the proposed steiner 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 310, 410 and 2510 ns, respectivley, while 4.5469e+004, 4.5857e+004 and 1.0374e+005 operations for up links. The results of the proposed steiner 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.

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

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

Table 2: Channel estimation cost of the proposed Steiner method, the classical time and frequency approaches in the scenarios of MIMO OFDM systems with 8×2 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	Steiner 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

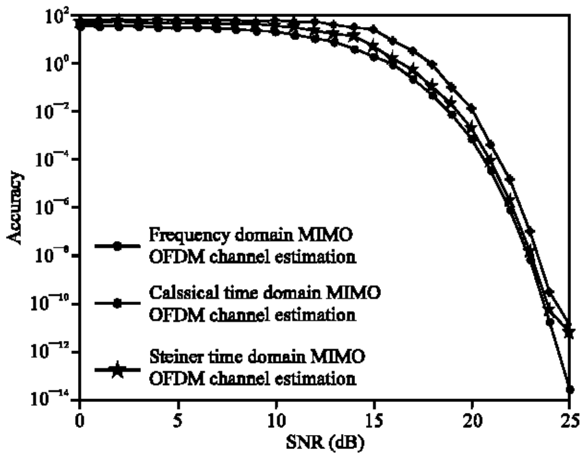

Fig. 3: Accuracy of the proposed steiner scheme, the classical time and frequency domain approaches in down links of MIMO OFDM with 8 transmit antennas for typical ITU vehicle profiles at 200 Hz Doppler frequency

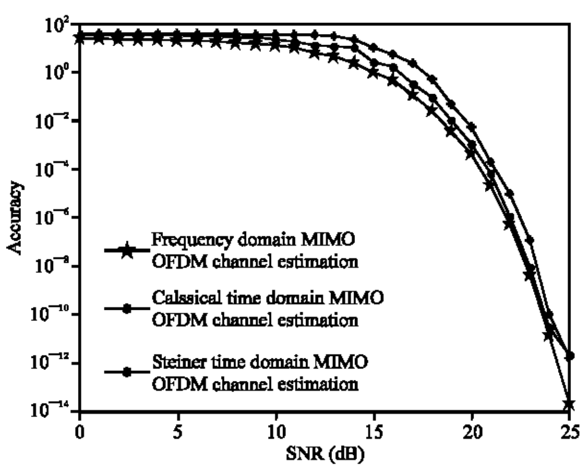

Fig. 4: Accuracy of the proposed steiner scheme, the classical time and frequency domain approaches in up links of MIMO OFDM with 2 transmit antennas for typical ITU vehicle profiles at 200 Hz Doppler frequency

classical time domain schemes and orthogonal Hadamard sequences as frequency pilot symbols. Figure 3 and 4 show the numerically simulated results in down and up links respectively. Clearly, the proposed Steiner channel estimation approach can obtain better channel estimation accuracy than that of classical time approach, as the channel pilot matrix constructed by the proposed Steiner 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.

CONCLUSION

In this study, aiming at two problems in MIMO OFDM systems, i.e., the bandwidth overhead of channel estimation and the challenge to construct large numbers of orthogonal training sequences, we adopt the Steiner channel estimation method for multi-user CDMA uplink radios for the scenarios of MIMO OFDM systems. The training sequences for different transmit antennas, are obtained by truncating the circular extension version of a basic sequences, which simplifies the construction of training sequences in time domain approaches. Furthermore, the pilot matrix assembled by these training sequences is a circular matrix and the high dimension matrix inversion operation can be avoid by FFT of the basic sequences. By reason of the good reversibility of the last pilot matrix, the proposed scheme has better channel estimation accuracy than the classical time approaches. Although its different performance from the classical frequency methods, this can be compensated by its simple implementation with less bandwidth overhead.

REFERENCES

Barhumi, I., G. Leus and M. Moonen, 2002. Optimal training sequences for channel estimation in MIMO OFDM systems in mobile wireless channels. Proceedings of International Zurich Seminar on Broadband Communications, Zurich, Switzerland.

Hongwei, Y., 2005. A road to future broadband wireless access: MIMO-OFDM-Based air interface. IEEE Commun. Mag., 43: 53-60.

Jagannatham, A.K. and B.D. Rao, 2006. Whitening-Rotation-Based Semi-Blind MIMO Channel Estimation. IEEE Trans. Signal Process, 54: 861-869.

Li, Y. and H. Wang, 2003. Channel estimation for MIMO-OFDM wireless communications, Personal, Indoor and Mobile Radio Communications. IEEE Proceedings of PIMRC, Beijing, China.

Nanda, S., R. Walton and J. Ketchum *et al*., 2005. A high-performance MIMO OFDM wireless LAN. IEEE Commun. Mag., 43: 101-109.

Ogawa, Y., K. Nishio and T. Nishimura *et al*., 2004. Channel and frequency offset estimation for a MIMO-OFDM system. Proceedings of Vehicular Technology Conference, Los Angeles, USA.

Proakis, J.G., 1995. Digital Communications. 3rd Edn. McGraw-Hill, New York, USA.

Shenghao, Y. and Z. Yuping, 2004. Channel estimation method for 802.11a WLAN with multiple-antenna. Proceedings of 10th Asia-Pacific Conference on Communications and 5th International Symposium on Multi-Dimensional Mobile Communications, Beijing China.

Steiner, B. and P.W. Baier, 1993. Low cost channel estimation in the uplink receiver of CDMA mobile radio systems. Fre-quenz, 47: 292-298.

Steiner, B. and P. Jung, 1994. Optimum and suboptimum channel estimation for uplink of CDMA mobile radio systems with joint detection. Eur. Trans. Telecommun. Related Technol., 5: 39-49.

Stuber, G.L., J.R. Barry and S.W. McLaughlin *et al*., 2004. Broadband MIMO-OFDM wireless communications. IEEE Proc., 92: 271-294.

Zelst, A. and T.C.W. Schenk, 2004. Implementation of a MIMO OFDM-based wireless LAN system. IEEE Trans. Signal Process, 52: 483-494.

Zheng, K., L. Huang and W. Wang *et al*., 2005. TD-CDM-OFDM: Evolution of TD-SCDMA toward 4G. IEEE Commun. Mag., 43: 45-52.