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Channel Code Aided Decision-Directed Channel Estimation for MIMO OFDM/SDMA Systems Based on the "Expectation-Conditional Maximization Either" Algorithm

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***Abstract*—In this paper, a Forward Error Coded (FEC) Decision-Directed (FEC-DD) channel estimation scheme is proposed for Multiple-Input Multiple-Output (MIMO) Orthogonal Frequency Division Multiplexing/Space Division Multiple Access (OFDM/SDMA) systems which is based on the Expectation-Conditional Maximization Either (ECME) algorithm. The proposed DD technique is combined with the Optimised Hierarchy Reduced Search Algorithm (OHRSA) based Multi-User Detector (MUD) and directly calculates the Maximization-Step (M-Step) by conditionally maximizing the logarithmic likelihood function of the "incomplete" data. We avoid the employment of matrix inversion, since only a single subcarrier's Frequency-Domain CHannel Transfer Function (FD-CHTF) is calculated at a time, since we assume that the other subcarriers' FD-CHTFs are the most recent estimates from the previous iteration of the proposed scheme. Our simulation results have demonstrated that the proposed scheme is capable of reducing the received power requirement by $4dB$ upon exploiting the error correction capability of a FEC decoder within the ECME loop.**

***Index Terms*—Multiple-Input Multiple-Output, Orthogonal Frequency Division Multiplexing/Space Division Multiple Access, channel estimation, Expectation-Conditinal Maximization Either.**

I. Introduction

Multiple-Input-Multiple-Output (MIMO) Othogonal Frequency Division Multiplexing/Space Division Multiple Access (OFDM/SDMA) systems have recently attracted substantial research interest, since they beneficially combine the advantages of OFDM and SDMA [1, 2]. More specifically, the transmitted signals of U simultaneous uplink mobile stations (MS), each of which employs a single transmit antenna, are received by an array of antennas at the base station (BS), where the superimposed signals are differentiated with the aid of their unique, user-specific Channel Impulse Responses (CIRs) [3, 4]. Hence, the performance of these MIMO SDMA/OFDM systems is critically dependent on the accuracy of the estimated channel knowledge. In the context of OFDM, the unique, user-specific spatial signature constituted by the users' Frequency-Domain CHannel Transfer Functions (FD-CHTFs) may be more conveniently estimated at the BS with the aid of FD pilots than in the Time-Domain (TD) [2, 5].

However, estimation of the FD-CHTFs in these systems is quite challenging, since the received signals are constituted by a superposition of signals transmitted from different MS antennas and the FD-CHTFs of all transmitter-receiver links have to be estimated simultaneously for each subcarrier. Numerous channel estimation techniques have been designed for MIMO systems, which employed different optimization criteria and imposed varying levels of implementational complexity [2]. As the affordable hardware complexity is increased according to Moore's law, it becomes more feasible to implement iterative receivers, allowing for substantial improvements of the physical layer functions. The iterative Expectation-Maximization (EM) algorithm [6] and the derivatives of this algorithm have been shown to strike an attractive trade-off between the performance attained and the complexity imposed.

Simplifying the Maximization-Step (***M-Step***) and accelerating the convergence to the best possible channel estimate constitute two different research directions for the EM-type algorithms. A classic EM based channel estimation algorithm and the so-called Space-Alternating Generalized Expectation-maximization (SAGE) based channel estimation algorithm were designed in [7] for OFDM systems invoking transmitter diversity and their convergence rates were compared. In [8], the so-called Unbiased Expectation-Maximization (UEM) and the Unbiased Expectation-Conditional-Maximization (UECM) channel estimator were designed by exploiting the similarity between the families of EM-type CIR estimators and Least-Square (LS) estimators. The authors of [9] derived an EM algorithm using low-rank approximation in order to avoid inverting large matrices, which substantially reduced the receiver's complexity. Choi [10] developed a robust EM-based channel estimation method which was applicable to diverse MIMO systems and it was capable of operating without requiring the Probability Density Function (PDF) of the channel parameters.

It is important to emphasize that EM based channel estimation methods typically carry out the ***M-Step*** by maximizing the conditional expectation of the logarithmic likelihood function (LLF) for all the available data. By contrast, **the novel contribution of this paper is that we design a Forward Error Code (FEC) Aided Decision-Directed (FEC-A-DD) FD-CHTF estimation technique based on the "Expectation-Conditional Maximization Either" (ECME) algorithm of [11], where the operations of the *M-Step* were replaced by steps that conditionally maximize the LLF of incomplete data**. The proposed ECME-based channel estimation scheme determines the FD-CHTF of a single subcarrier at a time, assuming that the other subcarriers' FD-CHTFs are known, for example, they are the most recent estimates from the previous iteration of the ECME algorithm.

The rest of the paper is organized as follows. In Section II, the classic MIMO OFDM/SDMA system model is described. Section III presents the proposed FEC-A-DD FD-CHTF estimation scheme based on the ECME algorithm. In Section IV, the performance of the proposed ECME-based channel estimation algorithm is evaluated by Monte-Carlo simulations both in terms of its average Mean-Squared Error (MSE) and Bit Error Ratio (BER) performance. Finally, our summary and concluding remarks are provided in Section V.

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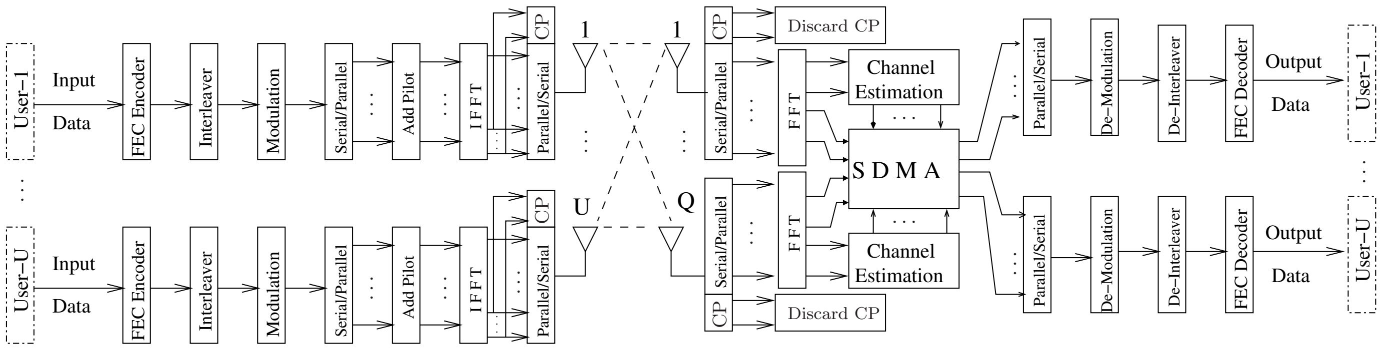

Fig. 1. Multi-user MIMO OFDM/SDMA uplink system model

II. System Model

The multi-user MIMO OFDM/SDMA UpLink (UL) system considered is shown in Fig. 1, where each of the U simultaneous users is equipped with a single transmission antenna, while the BS employs an array of Q antennas. All users simultaneously transmit their independent data streams, denoted by $\mathbf{b}^u, u = 1, 2, \cdots, U$. The information bits $\mathbf{b}^u$ are first encoded by the independent FEC encoder of Fig. 1 and interleaved. Then the data stream is grouped and Quadrature Phase Shift Keying (QPSK) or Quadrature Amplitude Modulated (QAM), followed by the classic K-point Inverse Fast Fourier Transform (IFFT) operator in order to obtain the TD modulated signal. After concatenating the Cyclic-Prefix (CP) of length N_{cp} samples, the resultant sequence is transmitted through the MIMO channel.

At the BS, the received signals $\mathbf{y}_q$ of antenna $q, q = 1, 2, \cdots, Q$, are constituted by the superposition of the indenpendently faded TD signals of the U users sharing the same space-frequency resource. The classic assumption is that we have $Q = U$, when the channel-matrix has a full rank and hence it is invertible. However, it is also important to consider the challening rank-deficient scenario [5, 12], when we have $U > Q$, because more users would like to access the system than the number of BS UL receiver antennas Q. The received signals are corrupted by the Gaussian noise at the array elements. After discarding the CP and performing the FFT-based demodulation of the received TD signals, we obtain Q separate received sequences for the n-th OFDM symbol $\mathbf{Y}_q[n], q = 1, 2, \cdots, Q$, which is given as the superposition of the different users' channel-impaired received signal contribution plus the Additive White Gaussian Noise (AWGN), which is expressed as:

$$\mathbf{Y}_q[n] = \sum_{u=1}^{U} \mathbf{X}^u[n]\mathbf{H}_q^u[n] + \mathbf{W}_q[n], \tag{1}$$

where we have $\mathbf{Y}_q[n] \in \mathbb{C}^{K\times1}, \mathbf{H}_q^u[n] \in \mathbb{C}^{K\times1}$ and $\mathbf{W}_q[n] \in \mathbb{C}^{K\times1}$ are column vectors in Equation (1) hosting the subcarrier-related variables $Y_q[n,k], H_q^u[n,k]$ and $W_q[n,k]$, respectively. Furthermore, $\mathbf{X}^u[n] \in \mathbb{C}^{K\times K}$ is a diagonal matrix with elements given by $X^u[n,k]$, where $k = 1, 2, \cdots, K$.

III. FEC-A-DD FD-CHTF Estimation Based on the ECME Algorithm

The structure of the proposed FEC-A-DD FD-CHTF estimation scheme, which is based on the ECME algorithm's philosophy [11] is illustrated in Fig. 2. The proposed channel estimation loop exploits the error correction capability of an arbitrary FEC decoder in order to mitigate the effects of noise and residual errors. It should be noted that our proposed solution is different from the conventional DD technique [13], which uses the current OFDM symbol's FD-CHTF estimate as the initial FD-CHTF estimate for the next OFDM symbol and then exploits the error correction capability of a FEC decoder to generate a more accurate FD-CHTF[1]. By contrast, the proposed scheme employs the FEC-A-DD technique for updating the channel estimates during the consecutive iterations of the ECME algorithm, as seen in Fig. 2. More specifically, the operation of the FEC-A-DD FD-CHTF estimation scheme is detailed as follows:.

- Step-1 Activate the Optimised Hierarchy Reduced Search Algorithm (OHRSA) based Multi-User Detector (MUD) [12] using the initial FD-CHTFs $\widehat{\mathbf{H}}^{(1)}$.
- Step-2 Reliable estimation of the transmitted signal is achieved by exploiting the error correction capability of the FEC decode of Fig. 2. The bit stream output by the FEC decoder is not delivered to the user before the FD-CHTF estimator's convergence, instead, it is reencoded and remodulated to generate $\widetilde{\mathbf{X}}^{(p)}$ of Fig. 2.
- Step-3 The reencoded and remodulated signal $\widetilde{\mathbf{X}}^{(p)}$ is then used in the "feedback loop" of Fig. 2 to perform FD-CHTF estimation based on the ECME algorithm.
- Step-4 The FD-CHTF estimate $\widehat{\mathbf{H}}^{(p+1)}$ is then transformed to the TD CIR by applying the IFFT and the insignificant noise-like CIR-taps are set to zero. Then remaining significant taps are then transformed back to the FD by the FFT, as shown in Fig. 2. The resultant FD-CHTF $\widetilde{\mathbf{H}}^{(p+1)}$ is then fed to the OHRSA MUD, according to Step-1, so that the process may continue interation-by-interation.

We will discuss the operations of the ECME-based FD-CHTF estimator of Step-3 and Fig. 2 in more detail below. The ECME algorithm [11] is a generalisation of the Expectation-Conditional-Maximization (ECM) algorithm [15], which is itself an extension of the EM algorithm [6]. The ECME algorithm may be arrived at upon replacing some or all of the Conditional-Maximization-Steps (CM-Steps) of the ECM algorithm, which maximizes the conditional expectation of the complete data's LLF, with steps that conditionally maximize the LLF of the incomplete data. Below, we will derive the operation of the ECME-based FD-CHTF estimator relying on the discrete-time received signal model of (1), which replaces the CM-Steps of the ECM algorithm with steps that conditionally maximize the LLF of the incomplete data. We assume that the initial estimate of the FD-CHTFs has been generated by the LS estimator of [16]. Additionally, we assume that the FD-CHTF remains time-invariant within S consecutive OFDM symbols, $S \geq 1$, hence we have

[1]We also note that other DD alternatives are also attractive, where the initial FD-CHTF estimate is derived using a low pilot-overhead and then exploiting within the same OFDM symbol for an improve second detection that in the absence of decision errors during the first detection. We now may assume the presence of 100% pilots [14].

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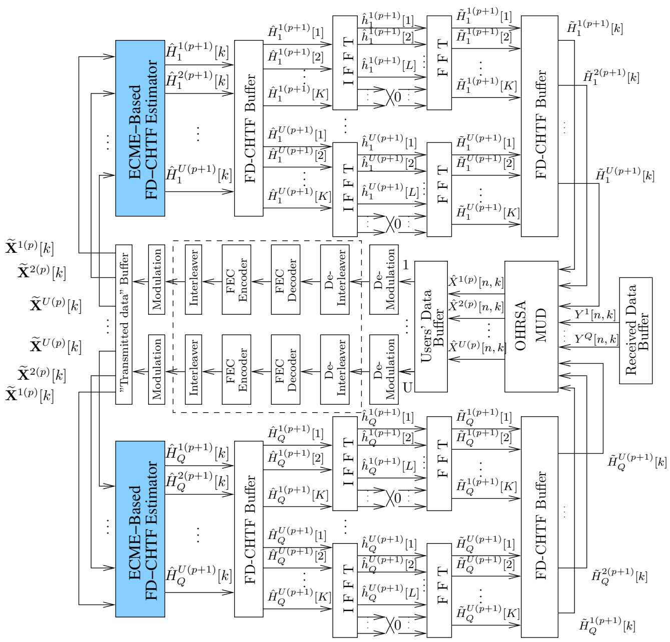

Fig. 2. Structure of the proposed FEC-A-DD FD-CHTF estimator scheme based on the ECME algorithm

$\mathbf{H}^u[1] = \mathbf{H}^u[2] = \cdots = \mathbf{H}^u[S] = \mathbf{H}^u$ for $u = 1, 2, \cdots, U$. By contrast, $S = 1$ implies that the FD-CHTF is time-varying from one OFDM symbol to the next. To simplify our notation without any loss of generality, we will omit the BS receiver antenna's index q from now on. Then the discrete-time received signal model associated with one of the BS antennas can be rewritten as,

$$\mathbf{Y}[s] = \widetilde{\mathbf{X}}^T[s]\mathbf{H} + \mathbf{W}[s], \tag{2}$$

where in Equation (2) $\mathbf{Y}[s] \in \mathbb{C}^{K\times 1}$ and $\mathbf{W}[s] \in \mathbb{C}^{K\times 1}$ are column vectors hosting the subcarrier-related variables $Y[s,k]$ and $W[s,k]$, respectively, and $\widetilde{\mathbf{X}}[s] \in \mathbb{C}^{UK\times K}, \mathbf{H} \in \mathbb{C}^{UK\times 1}$ which may be readily given by

$$\mathbf{Y}[s] = \Big[Y[s,1], Y[s,2], \cdots, Y[s,K]\Big]^T, \tag{3}$$

$$\mathbf{W}[s] = \Big[W[s,1], W[s,2], \cdots, W[s,K]\Big]^T, \tag{4}$$

$$\widetilde{\mathbf{X}}[s] = \Big[\widetilde{\mathbf{X}}^1[s], \widetilde{\mathbf{X}}^2[s], \cdots, \widetilde{\mathbf{X}}^U[s]\Big]^T, \tag{5}$$

$$\mathbf{H}[s] = \Big[\mathbf{H}^{1T}[s], \mathbf{H}^{2T}[s], \cdots, \mathbf{H}^{UT}[s]\Big]^T, \tag{6}$$

$$\widetilde{\mathbf{X}}^u[s] = diag\left\{\widetilde{X}^u[s,1], \cdots, \widetilde{X}^u[s,K]\right\}, \tag{7}$$

$$\mathbf{H}^u = \Big[H^u[1], H^u[2], \cdots, H^u[K]\Big]^T, \tag{8}$$

with $\widetilde{\mathbf{X}}[s]$ representing the reliable estimate of the transmitted signal referred to in "Step-2". Following the terminology of the ECME algorithm, an obvious choice for the "incomplete" data is the interference-contaminated composite received data $\mathbf{Y} = \big[\mathbf{Y}[1], \mathbf{Y}[2], \cdots, \mathbf{Y}[S]\big]^T$ of the U users. Then we estimate the FD-CHTF by maximizing the conditional likelihood function $f\left(\mathbf{Y}\middle|\mathbf{H}\right)$. More specifically, let us maximize $\log\ f\left(\mathbf{Y}\middle|\mathbf{H}\right)$ instead of $f\left(\mathbf{Y}\middle|\mathbf{H}\right)$, because using the logarithm allows us to convert the associated products to additions. Then we can conditionally maximize the LLF of the incomplete data as follows:

Once a reliable estimate of the transmitted signal was generated according to the above-mentioned Step-2, the PDF of $\mathbf{Y}$ given $\mathbf{H}$ may be readily formulated, invoking the assumption that the noise $\mathbf{W}[s]$ in Equation (2) is Gaussian, yielding:

$$\log\ f^{(p)}\left(\mathbf{Y}\middle|\mathbf{H}\right) = -KS\log(\pi\sigma^2) - \frac{1}{\sigma^2}\sum_{s=1}^{S}\left\|\mathbf{Y}[s] - \widetilde{\mathbf{X}}^{(p)T}[s]\mathbf{H}\right\|^2. \tag{9}$$

where $\widetilde{\mathbf{X}}^{(p)}[s]$ is the reencoded and remodulated version of $\widehat{\mathbf{X}}^{(p)}[s]$, while $\widehat{\mathbf{X}}^{(p)}[s]$ is the estimate of the transmitted signal at the output of OHRSA MUD, which was generated with the aid of the FD-CHTF estimate $\widehat{\mathbf{H}}^{(p)} = \left[\widehat{\mathbf{H}}^{1(p)T}, \widehat{\mathbf{H}}^{2(p)T}, \cdots, \widehat{\mathbf{H}}^{U(p)T}\right]^T$.

Jointly estimating all the FD-CHTFs of all users by directly maximizing $f^{(p)}\left(\mathbf{Y}\middle|\mathbf{H}\right)$ is a computationally demanding task. Hence below we simplify the estimation procedure as follows. The FD-CHTF of a single subcarrier is determined at a time, while assuming that the other subcarriers' FD-CHTFs are the most recent estimates generated during the previous iteration of the ECME algorithm. All the already estimated values may be grouped in a vector $\bar{\mathbf{H}}^{u(p)}[k]$ for the calculation of the estimate of $H^u[k]$ at iteration p, which is denoted by $\widehat{H}^{u(p)}[k]$. More specifically, the vector $\bar{\mathbf{H}}^{u(p)}[k]$ may be readily formulated as

$$\bar{\mathbf{H}}^{u(p)}[k] = \Big[\widehat{H}^{1(p)}[k], \cdots, \widehat{H}^{u-1(p)}[k], H^u[k], \widehat{H}^{u+1(p)}[k], \cdots, \widehat{H}^{U(p)}[k]\Big]^T. \tag{10}$$

Then we can conditionally maximize the LLF of the incomplete data, $\log\ f^{(p)}\left(\mathbf{Y}[k]\middle|\bar{\mathbf{H}}^{u(p)}[k]\right)$, for $u = 1, 2, \cdots, U, k = 1, 2, \cdots, K$ as follows

$$\widehat{H}^{u(p+1)}[k] = \arg\max_{H^u[k]}\ \log\ f^{(p)}\left(\mathbf{Y}[k]\middle|\bar{\mathbf{H}}^{u(p)}[k]\right). \tag{11}$$

Since we wish to maximize $\log\ f^{(p)}\left(\mathbf{Y}[k]\middle|\bar{\mathbf{H}}^{u(p)}[k]\right)$ with respect to $H^u[k]$, we may omit the expected value of the constant $-KS\log(\pi\sigma^2)$ and of all those terms that do not depend on $H^u[k]$. Hence we can rewrite Equation (11) in a more convenient form as follows:

$$\widehat{H}^{u(p+1)}[k] = \arg\min_{H^u[k]}\sum_{s=1}^{S}\left\|\mathbf{Y}[s,k] - \widetilde{\mathbf{X}}^{(p)T}[s,k]\bar{\mathbf{H}}^{u(p)}[k]\right\|^2. \tag{12}$$

The $(p+1)$st iteration estimates for $\widehat{H}^{u(p+1)}[k]$ can be obtained by directly differentiating Equation (12) with respect to $H^u[k]$, and setting it to zero. After further manipulations, we arrive at the following more explicit expression derived from Equation (11),

$$\widehat{H}^{u(p+1)}[k] = \frac{\sum_{s=1}^{S}\left(Y[s,k] - \widetilde{\mathbf{X}}^{(p)T}[s,k]\check{\mathbf{H}}^{u(p)}[k]\right)\widetilde{X}^{u(p)*}[s,k]}{\sum_{s=1}^{S}\widetilde{X}^{u(p)}[s,k]\widetilde{X}^{u(p)*}[s,k]}, \tag{13}$$

where the superscript * of $[\cdot]^*$ denotes conjugation, while $\widetilde{\mathbf{X}}^{(p)}[s,k]$ and $\check{\mathbf{H}}^{u(p)}[k]$ in Equation (13) are given by

$$\widetilde{\mathbf{X}}^{(p)}[s,k] = \Big[\widetilde{X}^{1(p)}[s,k], \widetilde{X}^{2(p)}[s,k], \cdots, \widetilde{X}^{U(p)}[s,k]\Big]^T, \tag{14}$$

$$\check{\mathbf{H}}^{u(p)}[k] = \Big[\widehat{H}^{1(p)}[k], \cdots, \widehat{H}^{u-1(p)}[k], 0, \widehat{H}^{u+1(p)}[k], \cdots, \widehat{H}^{U(p)}[k]\Big]^T. \tag{15}$$

Equations (9) and (13) constitute the kernel operations of the ECME-based FD-CHTF estimator. The stopping criterion of the ECME

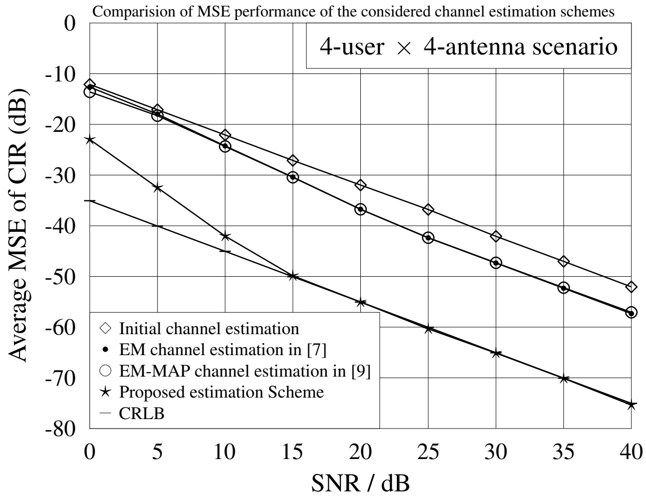

Fig. 3. Average MSE performance comparison for various channel estimation algorithms.

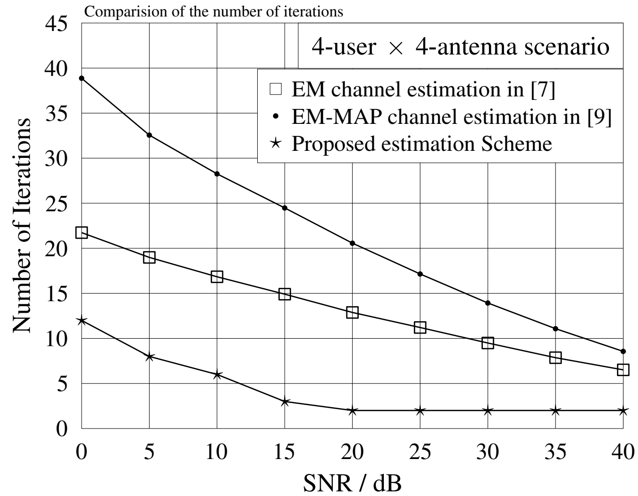

Fig. 4. Comparision of the iteration times versus SNR for the proposed ECME-based channel estimation and the existing EM-type channel estimation.

iterations was that the incremental changes of $\widehat{H}^{u(p)}[k]$ in consecutive iterations became smaller than ε, i. e.

$$\left\| \widehat{H}^{u(p+1)}[k] - \widehat{H}^{u(p)}[k] \right\|^2 \leq \varepsilon \tag{16}$$

was met for all users $u = 1, 2, \cdots, U$, where ε is a system parameter.

IV. Simulation Results

In this section, we will evaluate the performance of the proposed FEC-A-DD FD-CHTF estimation scheme, which is based on the ECME algorithm invoked for the multi-user MIMO OFDM/SDMA UL. Owning to space limitations, here we only provide results for $U = 4$ users and $Q = 4$ UL receiver antennas at the BS. In order to investigate the robustness of the proposed channel estimation scheme for diverse number of users, the achievable BER performance of the proposed scheme is also presented for $u = 1, 2, \cdots, 4$ simultaneous users, when the BS employs $Q = 4$ antennas. The parameters of each UL transmitter are set to values similar to those of the IEEE 802. 11n WLAN using $N_c = 64$ subcarriers and a single sample per subcarrier. The length of the CP is assumed to be $N_{cp} = 16$ samples and QPSK modulation is used for each user. In fact, the different users may employ different modulation schemes, but for simplicity, in this paper, we assume that all users employ QPSK. The channel parameters are similar to the IEEE 802.11n *channel model B* [17], which are as follows: the negative exponentially decaying Rayleigh fading CIR has $L = 9$ taps for each transmitter-receiver-antenna link, where the tap delays are $0, 1, \cdots, L-1$ samples. There are two clusters, which correspond to overlapping subsets of the tap delays: cluster 1 corresponds to the tap delays of 0 to 4 samples, while cluster 2 has tap delays spanning from 2 to 8 samples. Hence the CIR used in our simulations is given by

$$h(n) = \sum_{l=0}^{L-1} e^{-j\theta_l} \alpha_l \delta(n-l), \tag{17}$$

where $\alpha_l, 0 \leq l \leq L-1$ represents the total average path gain corresponding to the lth tap, while θ_l is the phase corresponding to the lth tap. The complex-valued channel envelope is assumed to be constant within $S = 50$ consecutive OFDM symbols duration.

The Cramer-Rao Lower Bound (CRLB) [18] characterizes the best achievable performance of an unbiased estimator. Since the Maximum Likelihood (ML) estimator is an unbiased estimator and the proposed FEC-A-DD FD-CHTF estimation scheme based on the ECME algorithm constitutes an iterative method of finding the ML estimate of parameters, we may characterize the proposed scheme by comparing the average MSE to the average CRLB, which is given by $\overline{CRLB}(h_l^u) = \sigma^2/A\cdot K\cdot S$, where A is the modulus of the modulated signals, for example, we have $A = 2$ for QPSK modulation.

Fig. 3 shows the attainable MSE performance versus the SNR for the different channel estimation schemes considered. The average MSE is defined by

$$\overline{MSE} = \frac{1}{UL} \sum_{u=1}^{U} E\left\{ \left\| \hat{\mathbf{h}}^u - \mathbf{h}^u \right\|^2 \right\}. \tag{18}$$

Observe in Fig. 3 that the average MSE performance of the proposed FEC-A-DD FD-CHTF estimation scheme is capable of approaching the CRLB in the Signal-to-Noise Ratio (SNR) range above $10dB$ and it substantially outperforms the EM-based channel estimation methods of [7, 9]. More specifically, the SNR gain is up to $12dB$ for the range of SNRs below $10dB$ and about $20dB$ in the SNR range above $10dB$.

In Fig. 4, the number of iterations required for the estimates $\widehat{H}^{u(p)}[k]$ to converge were compared. The value of ε in Equation (16) was set as $\varepsilon = 5 \times 10^{-4}$ for $SNR < 10dB$ and $\varepsilon = 10^{-4}$ for $SNR \geq 10dB$, respectively. Observe from Fig. 4 that the number of iterations required by the proposed ECME algorithm is lower than that of the classic EM algorithms, especially in the SNR range above $10dB$.

In order to provide an overall performance assessment, in Fig. 5, we evaluated the system's BER performance both with and without convolutional FEC coding, as shown using solid and dashed lines, respectively. Observe in Fig. 5 that our scheme approaches the BER performance of the ideal case associated with perfect channel information, both with and without FEC coding. More specificially, the SNR gain extracted from the BER comparisions is up to about $4dB$ and $2.5dB$ compared to the EM-based benchmarkers of [7, 9], with and without convolutional FEC coding, respectively.

Fig. 6 shows the resultant BER versus SNR curves for $u = 1, 2, \cdots, 4$ simultaneous users, while the BS employs an array of $Q = 4$ antennas. The BER performance of the conventional single-user, single-antenna based OFDM scheme is marked by the hearts and stars for the proposed channel estimation scheme and for the

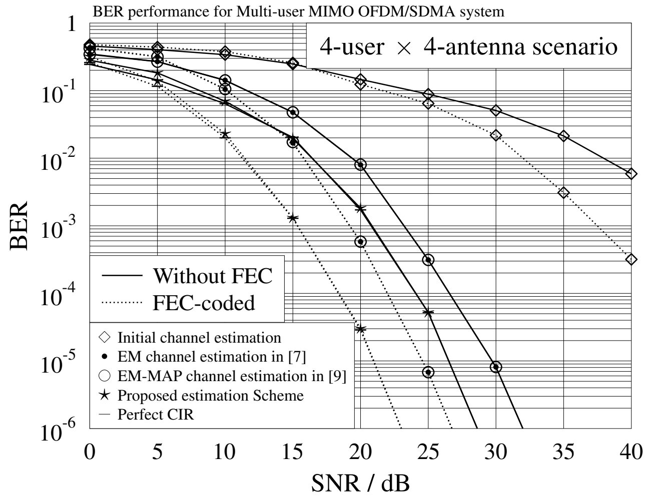

Fig. 5. BER performance comparison for various channel estimation algorithms. The system's BER performance recorded both with and without convolutional FEC coding, namely at the output of the de-modulation and FEC decoder, respectively, when using the schematic of Fig. 1.

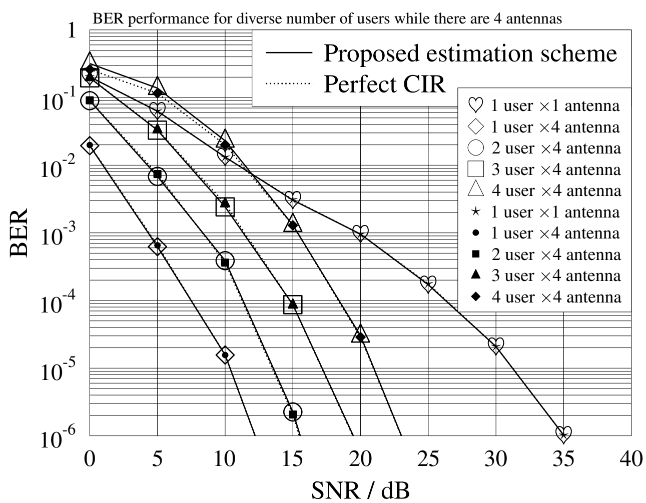

Fig. 6. BER performance versus SNR curves for one to four simultaneous users while the BS employs an array of $Q = 4$ antennas.

perfect CIR scenario, respectively. Observe from this figure that the proposed channel estimation scheme consistently approaches the ideal case associated with a perfect CIR, regardless of the number of users supported. An additional important observation is that the system employing $Q = 4$ antennas achieved a substantial performance gain as a benefit of spatial diversity, especially in the SNR range above $10dB$.

V. Conclusion

In this paper we presented a FEC-A-DD FD-CHTF estimation scheme, which further improved the ECME algorithm by exploiting the error correction capability of a FEC decoder for iteratively exchanging information between the decoder and the ECME algorithm. Our simulation results demonstrated that the proposed channel estimation scheme is capable of approaching the CRLB, while attaining a BER performance close to the ideal scenario associated with perfect CIR, while supporting multiple users.

References

[1] P. Vandenameele, L. Van der Perre, M. Engels, and H. De Man, "A novel class of uplink OFDM/SDMA algorithms for WLAN," in *Global Telecommunications Conference, 1999. GLOBECOM'99*, vol. 1, pp. 6–10, 1999.

[2] L. Hanzo, M. Münster, B. J. Choi, and T. Keller, *OFDM and MC-CDMA for broadband multi-user communications, WLANs, and broadcasting*. Piscataway, NJ: IEEE Press, 2003.

[3] P. Vandenameele, L. Van Der Perre, M. Engels, B. Gyselinckx, and H. De Man, "A combined OFDM/SDMA approach," *IEEE Journal on Selected Areas in Communications*, vol. 18, no. 11, pp. 2312–2321, 2000.

[4] S. Wu and Y. Bar-Ness, "Multiple phase noise correction for OFDM/SDMA," in *IEEE Global Telecommunications Conference, 2003. GLOBECOM'03*, vol. 3, pp. 1311–1315, 2003.

[5] M. Jiang, J. Akhtman, and L. Hanzo, "Iterative joint channel estimation and multi-user detection for multiple-antenna aided OFDM systems," *IEEE Transactions on Wireless Communications*, vol. 6, no. 8, pp. 2904–2914, 2007.

[6] A. Dempster, N. Laird, and D. B. Rubin, "Maximum likelihood from incomplete data via the EM algorithm," *Journal of the Royal Statistical Society. Series B (Methodological)*, vol. 39, no. 1, pp. 1–38, 1977.

[7] Y. Z. Xie and C. N. Georghiades, "Two EM-type channel estimation algorithms for OFDM with transmitter diversity," *IEEE Transactions on Communications*, vol. 51, no. 1, pp. 106–115, 2003.

[8] X. Wautelet, C. Herzet, A. Dejonghe, J. Louveaux, and L. Vandendorpe, "Comparison of EM-based algorithms for MIMO channel estimation," *IEEE Transactions on Communications*, vol. 55, no. 1, pp. 216–226, 2007.

[9] J. Gao and H. Liu, "Low-complexity MAP channel estimation for mobile MIMO-OFDM systems," *IEEE Transactions on Wireless Communications*, vol. 7, no. 3, pp. 774–780, 2008.

[10] J. Choi, "An EM based joint data detection and channel estimation incorporating with initial channel estimate," *IEEE Communications Letters*, vol. 12, no. 9, pp. 654–656, 2008.

[11] C. Liu and D. B. Rubin, "The ECME algorithm: A simple extension of EM and ECM with faster monotone convergence," *Biometrika*, vol. 81, no. 4, pp. 633–648, 1994.

[12] J. Akhtman, A. Wolfgang, S. Chen, and L. Hanzo, "An optimized-hierarchy-aided approximate log-MAP detector for MIMO systems," *IEEE Transactions on Wireless Communications*, vol. 6, no. 5, pp. 1900–1909, 2007.

[13] V. Mignone and A. Morello, "CD3-OFDM: A novel demodulation scheme for fixed and mobile receivers," *IEEE Transactions on Communications*, vol. 44, no. 9, pp. 1144–1151, 1996.

[14] M. Münster and L. Hanzo, "Parallel-interference-cancellation-assisted decision-directed channel estimation for OFDM systems using multiple transmit antennas," *IEEE Transactions on Wireless Communications*, vol. 4, no. 5, pp. 2148–2161, 2005.

[15] X. Meng and D. Rubin, "Maximum likelihood estimation via the ECM algorithm: A general framework," *Biometrika*, vol. 80, no. 2, pp. 267–278, 1993.

[16] Y. Li, "Simplified channel estimation for OFDM systems with multiple transmit antennas," *IEEE Transactions on Wireless Communications*, vol. 1, no. 1, pp. 67–75, 2002.

[17] V. Erceg, L. Schumacher, P. Kyritsi, *et al.*, "IEEE P802. 11 wireless LANs: TGn channel models," *IEEE 802.11 document 03/940r1*, 2003.

[18] S. M. Kay, *Fundamentals of statistical signal processing: estimation theory*. Englewood Cliffs, NJ: Prentice-Hall, 1993.