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Semiblind Joint Channel Estimation and Equalization for OFDM Systems in Rapidly Varying Channels

Habib ŞENOL[1], Erdal PANAYIRCI[2], H. Vincent POOR[3],
Onur OĞUZ[4], Luc VANDENDORPE[5]

[1]*Kadir Has University, Dept. of Computer Engineering, Istanbul, 34083, Turkey*
Tel: +90 212 5336532-1410, Fax: +90 212 5335753, Email: hsenol@khas.edu.tr
[2]*Kadir Has University, Dept. of Electronics Engineering, Istanbul, 34083, Turkey*
Tel: +90 212 5336532-1404, Fax: +90 212 5335753, Email: eepanay@khas.edu.tr
[3]*Princeton University, Dept. of Electrical Engineering, Princeton, NJ 08544, USA*
Tel: +1 609 2583500, Fax: +1 609 2583745, Email: poor@princeton.edu
[4]*Université Catholique de Louvain, Communications and Remote Sensing Laboratory,*
Place du Levant 2, 1348 Louvain-la-Neuve, Belgium
Tel: +32 10 478071 , Fax: +32 10 472089, Email: onur.oguz@uclouvain.be
[5]*Université Catholique de Louvain, Communications and Remote Sensing Laboratory,*
Place du Levant 2, 1348 Louvain-la-Neuve, Belgium
Tel: +32 10 472312 , Fax: +32 10 472089, Email:vandendorpe@tele.ucl.ac.be

Abstract: We describe a new joint iterative channel estimation and equalization algorithm for joint channel estimation and data detection for orthogonal frequency division multiplexing (OFDM) systems in the presence of frequency selective and rapidly time-varying channels. The algorithm is based on the expectation maximization-maximum a posteriori (EM-MAP) technique which is very suitable for the multicarrier signal formats. The algorithm leads to a receiver structure that yields the equalized output, using the channel estimates. The pilot symbols are employed to estimate the initial channel coefficients effectively and unknown data symbols are averaged out in the algorithm. The band-limited, discrete cosine serial expansion of low dimensionality is employed to represent the time-varying fading channel. In this way, the resulting reduced dimensional channel coefficients are estimated iteratively with tractable complexity. The extensive computer simulations show that the algorithm has excellent symbol error rate (SER) and mean square error (MSE) performances for very high mobility even during the initialization step.

Keywords: OFDM Systems, Time-Varying Channel, Channel Estimation, Channel Equalization, EM Algorithm

1. Introduction

Orthogonal frequency-division multiplexing (OFDM) with a cyclic prefix (CP) has been shown to be an effective method to overcome inter-symbol interference (ISI) effects due to frequency-selective fading with a simple transceiver structure. Consequently, it is becoming a key air interface of next-generation wireless communications systems such as the IEEE 802.16 family - better known as Mobile Worldwide Interoperability Microwave Systems for Next-Generation Wireless Communication Systems (WiMAX) - and by the Third-Generation Partnership Project (3GPP) in the form of its Long-Term Evolution (LTE) project. OFDM eliminates ISI and simply uses a one-tap equalizer to compensate for multiplicative channel distortion in quasi-static channels. However, in fading channels with very high mobility, the time variation of the channel over an OFDM symbol period results in a loss of subchannel orthogonality which leads to inter-carrier interference (ICI). Since mobility support is widely considered to be one of the key features in wireless communication systems, and in this case ICI degrades the performance of OFDM systems, OFDM transmission over very rapidly time varying multipath fading channels has been considered in a number of recent works [1, 2, 3, 4].

To reduce the effects of ICI, a time-domain channel estimator was proposed in [1] which assumed that the channel impulse response (CIR) varies in a linear fashion within the symbol duration. However, this assumption no longer holds when the normalized Doppler frequency takes substantially higher values. In a rapidly time-varying channel, the time-domain channel estimation method proposed in [5] is a potential candidate for the channel estimator, in order to mitigate ICI. This technique estimates the fading channel by exploiting the time-varying nature of the channel as a provider of time diversity and reduces the computational complexity using the singular-value decomposition (SVD) method. In [3], to handle rapid variation within an OFDM symbol, the pilot-based estimation scheme using channel interpolation was proposed. Moreover, coupled with the proposed channel estimation scheme, a simple Doppler frequency estimation scheme was proposed. In [4], two methods to mitigate ICI in an OFDM system with coherent channel estimation were proposed. Both methods employed a piece-wise linear approximation to estimate channel time-variations in each OFDM symbol. The first method extracted channel time-variation information from the cyclic prefix while the second method estimated these variations using the next symbol. Moreover, a closed-form expression for the improvement in average signal-to-interference ratio (SIR) was derived for a narrowband time-varying channel.

In this work, expectation maximization-maximum a posteriori (EM-MAP) based, a new joint iterative channel estimation and equalization algorithm is proposed for joint channel estimation and data detection for OFDM systems in rapidly varying channels. We employ orthogonal discrete cosine transform (DCT) basis functions to represent the time-varying fading channel having Jake's Doppler profile. The DCT basis functions are well suited to represent such a low-pass channel and have also the advantages of being independent of the channel statistics. In this way, the resulting reduced dimensional channel coefficients are estimated iteratively with tractable complexity. The algorithm leads to a receiver structure that yields the equalized output, using the channel estimates. The pilot symbols are employed to estimate the initial channel coefficients effectively and unknown data symbols are averaged out in the algorithm. Computer simulations show that the algorithm has excellent SER and MSE performance for very high mobility even during the initialization step.

2. Signal Model

We consider an OFDM system with N subcarriers. At transmitter, K out of N subcarriers are actively employed to transmit data symbols and noting is transmitted from the remaining $N-K$ carriers. The time-domain transmitted symbols are denoted as $s(n,k)$, where n is the OFDM symbol discrete-time index and $k \in \{0,1,\cdots,K\}$ is the subcarrier discrete-frequency index. A cyclic prefix of length L_c is then added. We assume a time-varying multipath mobile radio channel with discrete-time impulse response $h(n,\ell)$, $\ell = 0,1,\cdots,L-1$ where L is the maximum channel length and it is assumed that $L \leq L_c$. At the receiver, after matched filtering, symbol-rate sampling and discarding the symbols falling in the cyclic prefix, the received signal at the input of the discrete Fourier transform (DFT) can be expressed as

$$r(n) = \sum_{\ell=0}^{L-1} h(n,\ell)\, s(n-\ell) + w(n) \quad , \quad n = 0,1,2,\ldots,N-1 \tag{1}$$

where $w(\cdot)$ is a zero-mean complex additive Gaussian noise with variance N_0. By collecting receive signal samples in a vector, the above model can be expressed in vectorial form as follows

$$\mathbf{r} = \sum_{\ell=0}^{L-1} diag(\mathbf{s}_\ell)\,\mathbf{h}_\ell + \mathbf{w} \quad \in \mathcal{C}^N, \tag{2}$$

where $\mathbf{r}=[r(0), r(1), \cdots, r(N-1)]^T \in \mathcal{C}^N$, and $\mathbf{h}_\ell=[h(0,\ell), h(1,\ell), \cdots, h(N-1,\ell)]^T \in \mathcal{C}^N$, $\ell = 0, 1, \cdots, L-1$, represents the L-path wide sense stationary uncorrelated scattering (WSSUS) rayleigh fading coefficients. We assume Jake's channel model having an exponentially decaying normalized multipath channel powers described by $\sigma_\ell^2 = e^{-\ell/L}/(\sum_{m=0}^{L-1} e^{-m/L})$, $\forall\ \ell$. Note that due to the cyclic prefix (CP) employed at the transmitter, $s(-\ell) = s(N-\ell)$ for $\ell = 0, 1, \cdots, L-1$. We now define

$$\begin{aligned} \mathbf{s}_\ell &= [\, s(-\ell), s(-(\ell-1)), \cdots, s(N-(\ell+1))\,]^T \in \mathcal{C}^N \\ &= vshift(\mathbf{s}, \ell)\ , \end{aligned} \tag{3}$$

where $vshift(\mathbf{s}, \ell)$ denotes ℓ-step circular shift operator for a column vector $\mathbf{s} = [\mathbf{s}(0),\ \mathbf{s}(1),\ \cdots,\ \mathbf{s}(N-1)]^T \sim \mathcal{CN}(\mathbf{s}_P, \mathbf{\Sigma}_\mathbf{s}^{(0)})$. Defining $\mathbf{S}_\ell = diag(\mathbf{s}_\ell)$, $\mathbf{S}=[\mathbf{S}_0, \mathbf{S}_1, \cdots, \mathbf{S}_{L-1}] \in \mathcal{C}^{N\times LN}$ and $\mathbf{h}=[\mathbf{h}_0^T, \mathbf{h}_1^T, \cdots, \mathbf{h}_{L-1}^T]^T \in \mathcal{C}^{LN}$, the receive signal model in (2) can be rewritten as

$$\mathbf{r} = \mathbf{S}\mathbf{h} + \mathbf{w}\ . \tag{4}$$

3. DCT Expansions of the Multipath Channels

The number of unknown channel parameters to be estimated within one OFDM symbol interval is NL and it seems that the estimation of those coefficients is impossible even with pilot symbols since there are more unknowns to be determined than known equations. However, due to the banded character of the channel matrix in frequency domain, it is possible to reduce the number of unknown channel parameters, substantially, by representing the channel by a suitable orthogonal series expansion and taking only significant expansion coefficients for estimation. To reduce number of unknowns from N to D, for each multipath, we employ a discrete cosine transformation (DCT) for expansion of the ℓth multipath of the channel as

$$\mathbf{h}_\ell = \mathbf{\Psi}\mathbf{c}_\ell\ , \tag{5}$$

where $\mathbf{\Psi} \in \mathcal{C}^{N\times D}$ is the expansion matrix, and $\mathbf{c}_\ell \in \mathcal{C}^D$ is the coefficient vector for $\mathbf{h}_\ell$. Accordingly, DCT expansion of the overall channel vector is given as

$$\mathbf{h} = \mathbf{\Phi}\mathbf{c}\ , \tag{6}$$

where $\mathbf{c} = [\mathbf{c}_0^T, \mathbf{c}_1^T, \cdots, \mathbf{c}_{L-1}^T]^T \in \mathcal{C}^{LD}$, $\mathbf{\Phi} = \mathbf{I}_L \otimes \mathbf{\Psi} \in \mathcal{C}^{LN\times LD}$, $\otimes$ denotes kronecker product and $\mathbf{I}_L$ is an $L\times L$ identity matrix. So, substituting (6) in (4) we have the following observation model

$$\mathbf{r} = \mathbf{S}\mathbf{\Phi}\mathbf{c} + \mathbf{w}\ . \tag{7}$$

The dimension D of the basis expansion fulfills $\tilde{D} \le D \le N$. The lower bound is given by $\tilde{D} = [2(f_D)_{max} + 1]$, where $(f_D)_{max}$ is the maximum (one-sided) Doppler bandwidth defined by

$$(f_D)_{max} = \frac{v_{max} f_c}{c} T \tag{8}$$

where v_{max}, f_c, c are maximum supported velocity, the carrier frequency and the speed of light, respectively, and T is the OFDM symbol duration.

4. Joint Channel Estimation and Equalization Algorithm

Expectation maximization- maximum a posteriori (EM-MAP) channel estimation algorithm is optimal in minimizing symbol error rate and implemented in two steps. In the first step, called the expectation step (E-step), the auxiliary function

$$Q\big(\mathbf{c}|\mathbf{c}^{(i)}\big) = E_{\mathbf{s}}\big[\log p(\mathbf{r}|\mathbf{c},\mathbf{s}) \,\big|\, \mathbf{r},\mathbf{c}^{(i)}\big] + \log p(\mathbf{c}) \tag{9}$$

is computed where $\mathbf{c}^{(i)}$ is the estimation of $\mathbf{c}$ at the ith iteration. The conditional expectation in (9) is taken with respect to $\mathbf{s}$ given the observation $\mathbf{r}$ and assumes that $\mathbf{c}$ equals its estimate calculated at iteration (i). In a second step, called the maximization step (M-step), the unknown channel parameter vector $\mathbf{c}$ is updated according to

$$\mathbf{c}^{(i+1)} = \arg\max_{\mathbf{c}} Q\big(\mathbf{c}|\mathbf{c}^{(i)}\big) \ . \tag{10}$$

After going through the mathematical details, the final form of the updating rule of the DCT coefficients (reduced dimensional channel coefficient vector) can be obtained as follows

$$\mathbf{c}^{(i+1)} = \mathbf{G}^{(i)^{-1}}\mathbf{F}^{(i)} \ , \tag{11}$$

where

$$\begin{aligned} \mathbf{G}^{(i)} &= \boldsymbol{\Sigma}_{\mathbf{c}}^{-1} + \frac{1}{N_0}\boldsymbol{\Phi}^\dagger \mathbf{A}^{(i)}\boldsymbol{\Phi} \quad \in \mathcal{C}^{LD\times LD}, \\ \mathbf{F}^{(i)} &= \boldsymbol{\Phi}^\dagger \mathbf{B}^{(i)^\dagger}\mathbf{r} \quad \in \mathcal{C}^{LD} \ , \end{aligned} \tag{12}$$

and $\boldsymbol{\Sigma}_{\mathbf{c}}$ is the covariance matrix of $\mathbf{c}$ which can be determined easily from the channel correlation matrix. $\mathbf{A}^{(i)}{=}E_{\mathbf{s}}\big[\mathbf{S}^\dagger\mathbf{S}\big|\,\mathbf{r},\mathbf{c}^{(i)}\big] \ \in \mathcal{C}^{LN\times LN}$ and $\mathbf{B}^{(i)}{=}E_{\mathbf{s}}\big[\mathbf{S}\big|\,\mathbf{r},\mathbf{c}^{(i)}\big] \ \in \mathcal{C}^{N\times LN}$ can be determined after some algebra as follows

$$\mathbf{B}^{(i)} = \Big[\mathit{diag}\Big(\mathit{vshift}\big(\boldsymbol{\mu}_{\mathbf{s}}^{(i)},0\big)\Big), \mathit{diag}\Big(\mathit{vshift}\big(\boldsymbol{\mu}_{\mathbf{s}}^{(i)},1\big)\Big), \cdots, \mathit{diag}\Big(\mathit{vshift}\big(\boldsymbol{\mu}_{\mathbf{s}}^{(i)},L-1\big)\Big)\Big] \ , \tag{13}$$

where

$$\boldsymbol{\mu}_{\mathbf{s}}^{(i)} = E_{\mathbf{s}}\big[\mathbf{s}\big|\,\mathbf{r},\mathbf{c}^{(i)}\big] \quad \in \mathcal{C}^{N}. \tag{14}$$

It can be shown that

$$\boldsymbol{\mu}_{\mathbf{s}}^{(i)} = \mathbf{s}_P + \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)}\,\mathbf{H}^{(i)^\dagger}\,\big(\mathbf{H}^{(i)}\boldsymbol{\Sigma}_{\mathbf{s}}^{(0)}\,\mathbf{H}^{(i)^\dagger} + N_0\mathbf{I}_N\big)^{-1}\big(\mathbf{r} - \mathbf{H}^{(i)}\mathbf{s}_P\big) \ , \tag{15}$$

where $\mathbf{s}_P{=}\mathbb{F}^{-1}\,\mathbf{d}_P$, $\mathbf{d}_P = [d(0),0,\cdots,0,d(\Delta),0,\cdots,0,d(2\Delta),0,\cdots,0,d(P\Delta),0,\cdots]^T$ $\in \mathcal{C}^N$ denotes the pilot symbol vector containing P pilots with pilot spacing Δ, and

$$\mathbf{H}^{(i)}{=}\sum_{\ell=0}^{L-1}\mathit{mshift}\big(\mathit{diag}(\mathbf{h}_\ell^{(i)}),0,-\ell\big) \ . \tag{16}$$

Accordingly, $\mathbf{A}^{(i)}$ is given as follows

$$\mathbf{A}^{(i)} = \begin{bmatrix} \boldsymbol{\rho}_{0,0}^{(i)} & \cdots & \boldsymbol{\rho}_{0,L-1}^{(i)} \\ \vdots & \ddots & \vdots \\ \boldsymbol{\rho}_{L-1,0}^{(i)} & \cdots & \boldsymbol{\rho}_{L-1,L-1}^{(i)} \end{bmatrix} \ , \tag{17}$$

where

$$\boldsymbol{\rho}_{p,q}^{(i)} = diag\Big(dg\big(mshift(\mathbf{R}_{\mathbf{s}}^{(i)}, q, p)\big)\Big) \ , \tag{18}$$

$mshift(\mathbf{R}_{\mathbf{s}}^{(i)}, q, p)$ represents row-wise q-step and column-wise p-step circular shift of matrix $\mathbf{R}_{\mathbf{s}}^{(i)}$, and $dg(\cdot)$ returns the main diagonal vector of a matrix. In Eq. (18), the posterior autocorrelation matrix of $\mathbf{s}$ given $\mathbf{c}^{(i)}$ is obtained as

$$\mathbf{R}_{\mathbf{s}}^{(i)} = \boldsymbol{\mu}_{\mathbf{s}}^{(i)} \boldsymbol{\mu}_{\mathbf{s}}^{(i)\dagger} + \boldsymbol{\Sigma}_{\mathbf{s}}^{(i)} \ , \tag{19}$$

together with

$$\boldsymbol{\Sigma}_{\mathbf{s}}^{(i)} = \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} - \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} \mathbf{H}^{(i)\dagger} \big(\mathbf{H}^{(i)} \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} \mathbf{H}^{(i)\dagger} + N_0 \mathbf{I}_N\big)^{-1} \mathbf{H}^{(i)} \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)\dagger} \ . \tag{20}$$

Note that the estimation of the channel coefficients $\mathbf{c}$ given by (11) at $(i+1)$th EM iteration is a blind MMSE since, the unknown data is averaged out through $\mathbf{A}^{(i)}$ and $\mathbf{B}^{(i)}$.

4.1 Initialization of the EM-MAP Algorithm

The initial value of the reduced dimensional channel vector c can be determined from the received signal model

$$\mathbf{r} = \mathbf{Z}\,\mathbf{c} + \mathbf{w} \ , \tag{21}$$

where

$$\mathbf{Z} = \sum_{q=0}^{K-1} d(q) \mathbf{U}_q \ , \tag{22}$$

$d(q)$ is the data symbol transmitted on qth subcarrier, and

$$\mathbf{U}_q = \mathbb{F}_L^T(q) \otimes \Big(\big(\mathbf{1}_D^T \otimes \frac{1}{N} \mathbb{F}_N^*(q)\big) \odot \boldsymbol{\Psi}\Big) \ . \tag{23}$$

Here, $\mathbb{F}_L(q)$ represents the first L term of the qth column of the DFT matrix $\mathbb{F}$, $\odot$ denotes the element by element product and $\mathbf{1}_D$ stands for all-one column vector with length D. In Eq. (22), we consider $\mathbf{Z} = \mathbf{Z}_P + \mathbf{Z}_D$, where $\mathbf{Z}_P = \sum_{q \in \mathcal{I}_P}^{K-1} d(q) \mathbf{U}_q$ and $\mathbf{Z}_D = \sum_{q \in \mathcal{I}_D}^{K-1} d(q) \mathbf{U}_q$ are the matrices obtained from pilot and data symbols, respectively. Then the initial channel vector $\mathbf{c}^{(0)}$ can be obtained by a linear MMSE estimation technique as follows

$$\mathbf{c}^{(0)} = \boldsymbol{\Sigma}_{\mathbf{c}}\, \mathbf{Z}_P^\dagger \Big(\mathbf{Z}_P \boldsymbol{\Sigma}_{\mathbf{c}} \mathbf{Z}_P^\dagger + \sum_{q \in \mathcal{I}_D} \mathbf{U}_q \boldsymbol{\Sigma}_{\mathbf{c}} \mathbf{U}_q^\dagger + N_0 \mathbf{I}_N\Big)^{-1} \mathbf{r} \ . \tag{24}$$

5. Complexity Analysis

The computation complexity of the algorithm is presented in Table-1 under the assumption that $K = N$. Note that, in the initialization step of the algorithm in (24), $\boldsymbol{\Sigma}_{\mathbf{c}}\, \mathbf{Z}_P^\dagger \big(\mathbf{Z}_P \boldsymbol{\Sigma}_{\mathbf{c}} \mathbf{Z}_P^\dagger + \sum_{q \in \mathcal{I}_D} \mathbf{U}_q \boldsymbol{\Sigma}_{\mathbf{c}} \mathbf{U}_q^\dagger + N_0 \mathbf{I}_N\big)^{-1} \in \mathcal{C}^{DL \times N}$ is a precomputed matrix. Therefore, the initialization step requires only a multiplication of this precomputed matrix with $N \times 1$ $\mathbf{r}$ vector resulting in DLN complex multiplications (CMs). On the other hand, the covariance matrix $\boldsymbol{\Sigma}_s^{(0)}$, necessary for computation of (15) and (20), is a block matrix whose submatrices are diagonal with constant entries. Also, the convolution matrix $\mathbf{H}^{(i)}$ in (15) and (20) is a sparse matrix whose columns have only L non-zero entries. Consequently, in the computation of $\boldsymbol{\mu}_{\mathbf{s}}^{(i)}$ and $\boldsymbol{\Sigma}_{\mathbf{s}}^{(i)}$ in (15) and (20), the terms $\boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} \mathbf{H}^{(i)\dagger}$, $\mathbf{H}^{(i)} \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} \mathbf{H}^{(i)\dagger}$ and $\big(\mathbf{H}^{(i)} \boldsymbol{\Sigma}_{\mathbf{s}}^{(0)} \mathbf{H}^{(i)\dagger} + N_0 \mathbf{I}_N\big)^{-1}$ can be approximated by block matrices whose submatrices are diagonal with constant entries resulting in a reduced complexity algorithm. As a result, it follows from Table-1 that total computational complexity per detected symbol of the channel estimation algorithm presented in this work is approximately $(N+1)/2 + D^2L^2 + DL^2 + 2DL + L \ \propto \ \mathcal{O}(N)$.

Table 1: Computational Complexity Details

Eq. No	Variable	Complexity (CMs)
(24)	$\mathbf{c}^{(0)}$	NLD
(16) (using (5))	$\mathbf{H}^{(i)}$	NDL^2
(15)	$\boldsymbol{\mu}_{\mathbf{s}}^{(i)}$	$NL + 2\Delta^2 + 2\Delta L$
(20)	$\boldsymbol{\Sigma}_{\mathbf{s}}^{(i)}$	$3\Delta^2 + 2\Delta L$
(19)	$\mathbf{R}_{\mathbf{s}}^{(i)}$	$N(N+1)/2$
(17) and (13)	$\mathbf{A}^{(i)}$ and $\mathbf{B}^{(i)}$	0
(12)	$\mathbf{G}^{(i)}$ $\mathbf{F}^{(i)}$	$ND^2L(L+1)/2$ $ND(L+1)$
(11)	$\mathbf{c}^{(i+1)}$	$2D^2L^2$

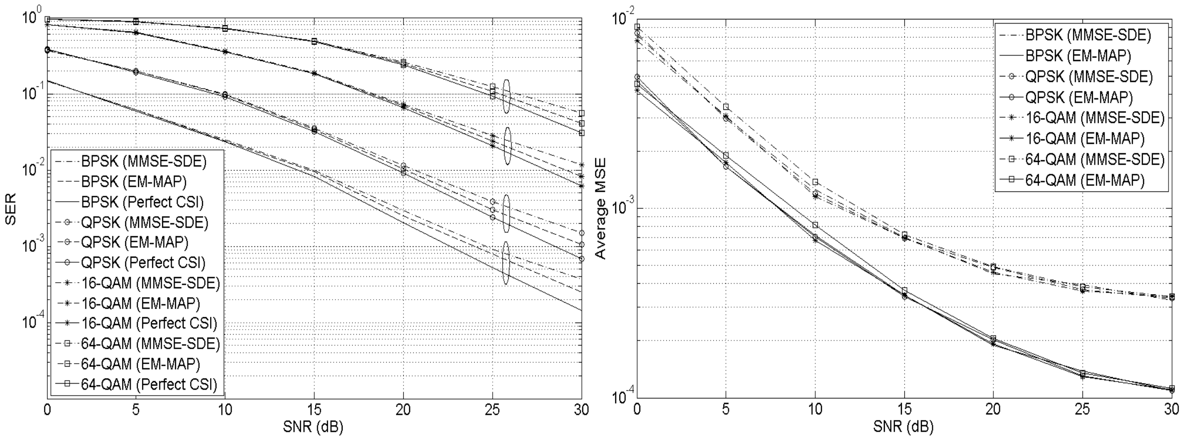
6. Simulation Results

In this section, we present computer simulation results to assess the performance of the OFDM systems operating with the proposed joint channel estimation and equalization algorithm. Simulation parameters are chosen as in Table-2. The initial estimate of the channel

Table 2: Simulation Parameters

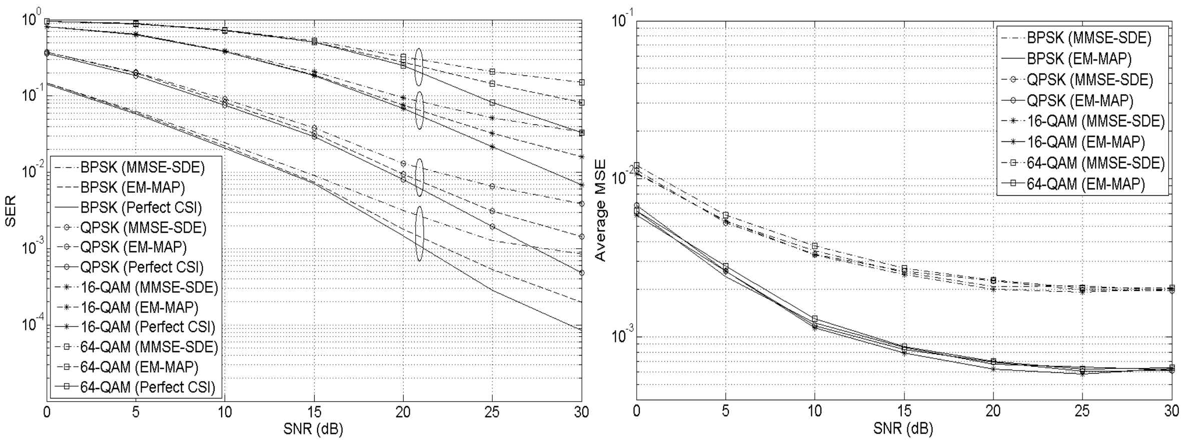
Bandwidth (BW)	10 MHz
Carrier Frequency (f_c)	2.5 GHz
Number of Subcarriers (N)	1024
Number of Multipaths (L)	3
Number of DCT Coefficients (D)	3 , 5
Number of Iterations (i_{max})	5
Pilot Spacing (Δ)	8 , 12
Modulation Formats	BPSK , QPSK , 16QAM , 64QAM

is performed by the reduced-complexity linear MMSE estimation techniques based on the pilot symbols.We refer to this method for obtaining the initial channel and data estimates as MMSE separate detection and estimation (MMSE-SDE) scheme. The solid and the dashed curves in Figures 1 and 2 represent the SER and MSE performance curves of the EM-MAP and MMSE-SDE algorithms, when the pilot spacing is chosen as $\Delta = 8$ and the corresponding mobilities are $f_DT = 0.0284$ ($v = 120$ km/h) and $f_DT = 0.0852$ ($v = 360$ km/h). The multipath wireless channel having an exponentially decaying power delay profile with the normalized powers, $\sigma_0^2 = 0.448$, $\sigma_1^2 = 0.321$, and $\sigma_2^2 = 0.230$, is chosen. It was observed that a maximum of three iterations were sufficient in order for the EM-MAP algorithm to converge. We conclude from these curves that even when the number of DCT coefficients is chosen to be fairly small as compared to the total number of coefficients, the performance loss in SER is not significant when CSI not available. We also observed that the symbol error rate (SER) performance of the EM-MAP algorithm obtained at the end of the third iteration step is better than that of the MMSE-SDE and the performance difference becomes more significant at higher mobilities. On the other hand, we observe that the average mean square error (MSE) performance of the EM-MAP algorithm is substantially better than that of the MMSE-SDE. In Fig. 3, effects of channel estimation on the average MSE and on the SER performance are investigated as functions of the pilot spacing (Δ) with the mobility $f_DT = 0.0284$ ($v = 120$ km/h). It is concluded from Fig. 3 that the SER and MSE performances do not change significantly for pilot spacing 8 and 12.



(a) (b)

Figure 1: SER and MSE performances of the EM-MAP and MMSE-SDE algorithms for $f_D T = 0.0284$ ($v = 120$ km/h), $N = K = 1024$, $\Delta = 8$, $L = 3$

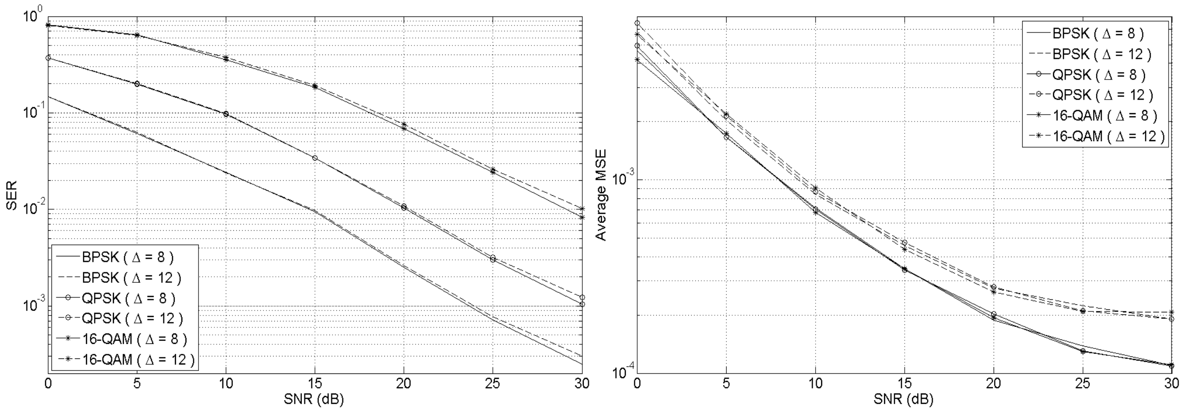


(a) (b)

Figure 2: SER and MSE performances of the EM-MAP and MMSE-SDE algorithms for $f_D T = 0.0852$ ($v = 360$ km/h), $N = K = 1024$, $\Delta = 8$, $L = 3$

7. Conclusions

In this work, the problem of joint iterative channel estimation and equalization algorithm for joint channel estimation and data detection in OFDM systems operating over frequency selective and very rapidly time-varying channels has been investigated. We have presented an iterative algorithm based on the EM-MAP algorithm for channel estimation that incorporates the channel equalization and the data detection. The band-limited cosine orthogonal basis functions have been applied to describe the rapidly time-varying channel. Initial channel coefficients are effectively obtained by the pilot aided MMSE estimator and unknown data symbols are averaged out in the algorithm. It has been shown by computer simulation that the proposed algorithm has excellent symbol error rate and channel estimation performance even with a very small number of channel expansion coefficients, resulting in reduction of the computational complexity substantially.



(a) (b)

Figure 3: SER and MSE performances of the EM-MAP algorithm with different pilot spacing for $f_D T = 0.0284$ ($v = 120$ km/h), $N = K = 1024$, $L = 3$

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References

[1] W. G. Jeon, K. H. Chang and Y. S. Cho, "An equalization technique for orthogonal frequency-division multiplexing systems in time-variant multipath channels," *IEEE Trans. Commun.*, vol. 47, no. 1, pp. 27–32, Jan. 1999.

[2] X. Cai and G. B. Giannakis, "Low-complexity ICI suppression for OFDM over time- and frequency-selective Rayleigh fading channels," in *Proc. 36th Asilomar Conf. Signals, Systems, Computers*, (Pacific Grove, CA), pp. 1822–1826, 2002.

[3] W. Song and J. Lim, "Pilot-symbol aided channel estimation for OFDM with fast fading channels," *IEEE Tran. Broadcas.*, vol. 49, no. 4, pp. 398–402, 2003.

[4] Y. Mostofi and D. Cox, "ICI mitigation for pilot-aided OFDM mobile systems," *IEEE Trans. Wireless Commun.*, vol. 4, no. 2, pp. 765–774, 2005.

[5] Y.-S. Choi, P. J. Voltz and F. A. Cassara, "On channel estimation and detection for multicarrier signals in fast and selective Rayleigh fading channels," *IEEE Trans. Commun.*, vol. 49, no. 8, pp. 1375–1387, Aug. 2001.

