Regularized MMSE ICI Equalization for OFDM Systems Over Doubly Selective Channels

Evangelos Vlachos
Dept. of Computer Engineering and Informatics
University of Patras & C.T.I RU-8
26500 Rio - Patra, Greece
vlaxose@ceid.upatras.gr

Aris S. Lalos
Dept. of Signal Theory and Communications
Universitat Politecnica de Catalunya (UPC)
08860 Castelldefels, Barcelona, Spain
aristeidis.lalos@tsc.upc.edu

Kostas Berberidis
Dept. of Computer Engineering and Informatics
University of Patras & C.T.I RU-8
26500 Rio - Patra, Greece
berberid@ceid.upatras.gr

***Abstract*—In this work, we consider a wireless OFDM system operating over doubly selective channels, where the Doppler effect destroys the orthogonality between subcarriers and hence, results into severe intercarrier interference (ICI). To mitigate this effect, computational demanding equalization schemes that require the inversion of the channel matrix, should be applied. In order to achieve linear complexity in the number of the subcarriers, a banded approximation of the channel matrix is usually adopted, whereas the performance of the equalizer is significantly degraded. To recover this performance loss, we propose a regularized estimation framework for MMSE ICI equalization in the frequency domain, where the complexity remains linear with respect to the number of the subcarriers. Simulation results verify the effectiveness of the proposed regularization.**

I. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) has been used for several wireless applications such as broadcasting of digital audio and digital video [1], [2] and mobile radio communication [3], [4]. In OFDM, the serial data stream is converted into parrallel substreams (OFDM blocks) transmitted over narrowband subchannels, which are generally known as subcarriers. The assumption of time-invariant frequency-selective multipath channels, ensures orthogonality between the subcarriers and thus, allows the use of a simple one-tap equalizer for recovering the transmitted symbol on each subcarrier.

In applications with high levels of mobility and rate, i.e., digital video broadcasting for hand-held terminals (DVB-H) [2], the experienced channels are usually time- and frequency-selective (so called doubly selective), and temporal variations within one OFDM block corrupt the orthogonality of different subcarriers, generating power leakage among them. This phenomenon makes the single-tap equalization unreliable [5], [6] since is not able to mitigate the interference between the non orthogonal subcarriers (Intercarrier Interference ICI).

To deal with the issue of ICI ,several approaches offering various complexity-performance tradeoffs have been proposed. Jeon et al. [7] truncated the channel matrix discarding a small number of coefficients, in order to reduce the dimension of matrix inversion for ZF equalizer. Choi et al. [8], presented linear and decision-feedback frequency domain equalizers. Cai and Giannakis [9], derived a low-complexity equalization scheme which was based on channel truncation and successive interference cancellation. This approach has also been exploited in [10], [11]. Schniter [12] derived an iterative equalizer which was composed by two stages, a first stage for ICI reduction and a second one for performing MMSE equalization. A standard approach used for the design of low-complexity equalizers in all the aforementioned works, is the approximation of the channel matrix in the frequency-domain with a banded one. This operation corresponds to the mitigation of interference of only a small number of selected subcarriers that is determined by a truncation factor.

The aforementioned approximation leads to reduction of both the computational complexity and equalization performance as well. In this paper we propose a regularization method for introducing additional information to the MMSE equalization design problem in order to minimize the performance degradation occured by matrix truncation. It is shown that by taking into account the frequency-domain correlation between different subcarriers we can significantly minimize the equalization performance degradation due to the use of the banded channel matrix, by achieving at the same time a linear complexity with respect to the number of the subcarriers.

The rest of this paper is organized as follows. We present the system model in Section II, where the channel model and the OFDM system are briefly described. In Section III, MMSE ICI equalization schemes are reviewed, while in Section IV the new regularized scheme is derived. In Section V, the proposed algorithms performance is evaluated through appropriate simulations. Finally, this paper is concluded in Section VI.

Notation: Lower-(upper-)case boldface letters are reserved for column vectors (matrices); the imaginary unit is denoted by $j = \sqrt{-1}$; $[A]_{i,j}$ denotes the component of the matrix $\mathbf{A}$ at the i-th row and the j-th column; $\delta(\cdot)$ denotes the Dirac's delta function; $(\cdot)^T, (\cdot)^H, (\cdot)^*$ denote the matrix transpose, complex conjugate transpose and complex conjugate respectively; $\mathcal{B}_Q(\mathbf{M})$ denotes the banded truncation of the matrix M, with $2Q+1$ non-zero elements at each row; $A_{|\Omega}$ denotes the submatrix with columns of $\mathbf{A}$ based on the index set Ω; $\mathbf{x}_{|\Omega}$

This work was supported in part by EU and national funds via the National Strategic Reference Framework (NSRF) - Research Funding Program Thales (ENDECON), and in part by the University of Patras.

978-1-4799-4796-6/13/$31.00 © 2013 IEEE

denotes the subvector with elements of $\mathbf{x}$ based on the index set Ω; $\mathbf{0}_{N\times N}$ denotes a $N \times N$ matrix with zeros and $\mathbf{I}_N$ the $N \times N$ identity matrix.

II. System Model

In this section we initially present the channel model and then we formulate the input output model of an OFDM wireless system that operates over double selective channels, as a set of linear equations.

A. Channel Model

In wireless communications, a doubly selective fading channel is often modeled as a wide sense stationary uncorrelated scattering (WSSUS) channel [13]. In discrete time model, the channel impulse response (CIR) of this WSSUS channel can be expressed as

$$h(n,\tau) = \sum_{l=0}^{L-1} a_l(n)\delta(\tau - l) \tag{1}$$

where $a_l(n)$ is complex zero mean Gaussian variable and L denotes the maximum channel cpan. The channel autocorrelation function is given by

$$E\{h(l_1, t+T)h^*(l_2, t)\} = \sigma_{l_1}^2 r_t(T)\delta(l_1 - l_2) \tag{2}$$

where $\sigma_{l_1}^2$ denotes the variance of the l_1 tap. According to Jakes' model [14], $r_t(T) = J_0(2\pi f_d T)$, where $J_0(\cdot)$ denotes the zeroth-order Bessel function of the first kind and f_d is the maximum Doppler frequency, and the Doppler power spectrum is defined as

$$r_f(\Delta f) = \sum_{l=0}^{L-1} \sigma_l^2 e^{-j2\pi\Delta f \tau_l} \tag{3}$$

B. OFDM Baseband Model

Let us consider an OFDM system with N subcarriers operating over a doubly-selective discrete-time baseband equivalent channel. Let $\mathbf{s} = [\ s_1 \ \ \ldots \ \ s_N\ ]^T$ be a set of N transmitted symbols where symbol s_k is transmitted via the k-th subcarrier. These symbols are forwarded at the input of the inverse discrete Fourier transform (DFT) unit, which output is denoted by $\mathbf{u} = \mathbf{F}^H\mathbf{s}$. At the cyclic prefix (CP) adder, the time-domain OFDM symbol $\hat{\mathbf{u}}$ of length $M = N + N_{cp}$ is formed by adding a CP of length N_{cp} at the beginning of vector $\mathbf{s}$. This operation may be described in matrix form as follows

$$\hat{\mathbf{u}} = \mathbf{C}^{cp}\mathbf{u} = \begin{bmatrix} \mathbf{0}_{(N-N_{cp})\times N} & \mathbf{I}_{N_{cp}} \\ \multicolumn{2}{c}{\mathbf{I}_N} \end{bmatrix} \mathbf{u}. \tag{4}$$

Assuming a causal channel with maximum delay spread $L \leq N_{cp}$, the received signal at the input of the OFDM demodulator may be written in matrix form as

$$\mathbf{x} \ = \ \mathbf{H}\hat{\mathbf{u}} + \mathbf{w} \tag{5}$$

where

$$[H]_{i,j} = \begin{cases} a_{i-j}(i), \ \text{for } i-j \in [0, L-1] \\ 0, \ \text{elsewere} \end{cases} \tag{6}$$

and $\mathbf{w}$ is a vector with complex Gaussian entries with zero mean and variance σ_w^2. The block of time-domain samples $\mathbf{x}$, passes through the CP removal unit of the OFDM demodulator, where the first N_{cp} samples are discarded. This operation may be written in matrix form as

$$\mathbf{z} = \mathbf{R}^{cp}\mathbf{x} = [\ \mathbf{0}_{(M-N)\times N} \quad \mathbf{I}_N\ ]\,\mathbf{x}. \tag{7}$$

The output of the CP removal unit passes through the DFT unit whose output reads as

$$\mathbf{y} = \mathbf{F}\mathbf{z} = \underbrace{\mathbf{F}\mathbf{R}^{cp}\mathbf{H}\mathbf{C}^{cp}\mathbf{F}^H}_{\mathbf{A}}\mathbf{s} + \mathbf{w} = \mathbf{A}\mathbf{s} + \mathbf{w} \tag{8}$$

where the entries of $\mathbf{w}$ are complex normal random variables with mean zero and variance σ_w^2 due to the unitary property of $\mathbf{F}$.

With a time-invariant channel, i.e. $a_i(1) = a_i(2) = \cdots = a_i(N)$, $i = 1, \ldots L$, the matrix $\mathbf{A}$ becomes diagonal since $\mathbf{R}^{cp}\mathbf{H}\mathbf{C}^{cp}$ has a circulant structure, and therefore equalization is possible with $\mathcal{O}(N)$ operations. On the contrary, with a time-varying channel, the matrix $\mathbf{A}$ is no longer diagonal due to the introduced ICI, and, hence, nontrivial equalization techniques are required.

III. MMSE Based Equalization schemes

Let us first review in this section some results and concepts concerning MMSE-based ICI mitigation. Although, the MMSE criterion leads, in general, to algorithms with higher implementation complexity compared to the zero-forcing one, it provides enhanced noise-suppression performance and stability. These properties are quite important in cases of doubly selective channels since the associated matrix becomes very often ill-conditioned. The results which are presented here will form the basis for the new techniques that will be derived in the next section

A. Conventional MMSE Equalizer

A straightforward way to mitigate ICI would be to apply MMSE equalization to the N parallel streams of (8). The $\mathbf{g}_k \in \mathbb{C}^{N\times 1}$ equalizer of the k-th stream may be computed as the minimizer of the cost function

$$\mathcal{J}(\mathbf{g}_k) \ = \ E\{\|s_k - \mathbf{g}_k^H\mathbf{y}\|_2^2\} \tag{9}$$

Assuming full knowledge of the channel state information and that data and noise signals are zero-mean uncorrelated sequences, the autocorrelation matrix can be expressed as $\mathbf{R} \triangleq E\{\mathbf{y}\mathbf{y}^H\} = (\mathbf{A}\mathbf{A}^H + \sigma_w^2\mathbf{I}_N)$. Recall that, since the autocorrelation matrix is Hermitian and positive definite, the minimizer of (10) has unique solution $\forall k$ and the symbols estimate is given by

$$\tilde{\mathbf{s}} = \mathbf{G}^H\mathbf{y} = \mathbf{R}^{-1}\mathbf{A}\mathbf{y}. \tag{10}$$

where $\mathbf{G} = [\mathbf{g}_1\ \mathbf{g}_2\ \ldots \mathbf{g}_N]$. The computational cost in order to solve (10) requires $\mathcal{O}(N^3)$ operations, in the general case.

B. Banded MMSE Equalizer

The conventional MMSE equalizer is impractical for systems with long OFDM symbols. In order to reduce the complexity, it has been proposed to approximate the channel matrix $\mathbf{A}$ with a banded one, i.e. $\hat{\mathbf{A}} = \mathcal{B}_Q(\mathbf{A})$, which captures the significant ICI effect depending on the truncation bandwidth $K = 2Q + 1$. In that case, the MMSE equalization matrix is computed by the following expression

$$\mathbf{G} = (\hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2 \mathbf{I}_K)^{-1}\hat{\mathbf{A}}\mathbf{y} \tag{11}$$

If K is the number of the non-zero elements for each row of $\hat{\mathbf{A}}$, the banded MMSE equalizer can be computed with $\mathcal{O}(NK^2)$ operations, using the $\mathbf{LDL}^H$ factorization which takes account the special structure of the system [].

The banded approximation provides a tradeoff between computational efficiency and performance, depending on the maximum Doppler spread f_d. Thus, for an increased f_d, the truncation parameter Q should also be increased appropriately, in order to capture the significant ICI leakage.

C. Iterative MMSE Equalizer

In order to increase the performance of the banded MMSE equalizer, an iterative scheme has been proposed [9], where the equalization is conducted in an recursive manner over the subcarriers. Let $\mathcal{I}_k$ be the $K \times 1$ index vector which is defined as follows for the i-th element, defined as follows for the i-th element,

$$\mathcal{I}_k(i) = (k - Q - 1 + i) \bmod N + 1, \quad i = 1, \ldots, N \tag{12}$$

where k is the subcarrier index and $Q \in \mathbb{N}$ is the truncation parameter. Then, the estimation of the k-th symbol is based on the $\mathbf{y}_{|\mathcal{I}_k}$ received subvector. The equalization vector for the subcarrier k, $\mathbf{g}_k \in \mathcal{C}^{K\times 1}$, can be computed as the solution of the following system,

$$\mathbf{R}_{\mathcal{I}_k}\mathbf{g}_k = \mathbf{A}_{|\mathcal{I}_k}^H \tag{13}$$

where $\mathbf{R}_{\mathcal{I}_k} = E\{\mathbf{y}_{|\mathcal{I}_k}\mathbf{y}_{|\mathcal{I}_k}^H\} = \mathbf{A}_{|\mathcal{I}_k}^H\mathbf{A}_{|\mathcal{I}_k} + \sigma_w^2\mathbf{I}_K$. Using the inversion lemma for the truncated system (13), the cost for the computation of the inverse of the autocorrelation matrix $\mathbf{R}_{\mathcal{I}_k}$ can be reduced to $\mathcal{O}(NK)$ per subcarrier iteration and to $\mathcal{O}(N^2K)$ for the whole OFDM block [9].

IV. Regularized MMSE Equalization

A. Preliminaries

Regularized estimation methods aim to extract as much information as possible from the observations in order to avoid overfitting due to inexact modeling. The naive solution that ignores the model mismatch can lead to significant loss in performance [15]. Let us consider the linear model of the OFDM communication system in the frequency-domain

$$\mathbf{y} = \mathbf{A}\mathbf{s} + \mathbf{w} \tag{14}$$

where $\mathbf{y}$ denotes the received signal vector, $\mathbf{A}$ denotes the channel matrix, $\mathbf{s}$ the transmitted symbols vector, and $\mathbf{w}$ the noise vector. In the case where the channel impulse response is perfectly known at the receiver, the MMSE linear estimator is given by (10). However the receivers' performance is based on the accuracy of the estimation of the channel matrix, which in practice may include errors due to several reasons, i.e. imperfect channel estimation, partial channel knowledge, roundoff errors, etc. In that case, the above linear estimator is no longer optimal for minimizing the MSE, resulting into severe performance degradation. Hence, a regularized estimator would be more suitable in this case.

A common regularization approach is the Tikhonov regularization [16]. Let $\mathbf{g}_k \in \mathbb{C}^{N\times 1}$ be the equalizer of the k-th stream, then the Tikhonov regularization cost function is written as

$$\mathcal{J}(\mathbf{g}_k) \;\; = \;\; E\{\|s_k - \mathbf{g}_k^H\mathbf{y}\|_2^2\} + \|\mathbf{\Gamma}\mathbf{g}_k\|_2^2 \tag{15}$$

for suitable chosen regularization matrix $\mathbf{\Gamma}$. The minimization of aforementioned problem $\forall k$ results in the following symbol recovery

$$\hat{\mathbf{s}}_R = (\mathbf{A}^H\mathbf{A} + \mathbf{\Gamma}^H\mathbf{\Gamma} + \sigma_w^2\mathbf{I}_N)^{-1}\mathbf{A}^H\mathbf{y}. \tag{16}$$

By appropriately selecting the matrix $\mathbf{\Gamma}$ in the penalized term in (15) we can significantly reduce the signal recovery error.

B. Proposed Method

In this section, we develop a regularized estimation framework for MMSE ICI equalization in frequency domain that significantly reduces the truncation effects to the MSE between the original and recovered signal. Initially, we consider the original OFDM input output model, and treat the neglected part of channel matrix due to truncation as an error term with specific second order statistics. Then, by selecting a regularizer that takes into account the statistics of the error term we end up to an efficient equalization scheme with linear complexity with respect to the number of the subcarriers.

As it has been described in Section III, in order to reduce the complexity of the MMSE ICI equalization, we consider a banded part of the channel matrix A, with only K terms at each row, neglecting the $N - K$ remaining ones.

Let us express the real channel matrix $\mathbf{A}$ as $\mathbf{A} \equiv \hat{\mathbf{A}} + \mathbf{\Delta}$, where $\hat{\mathbf{A}}$ is a banded matrix constructed from the Q upper and Q lower diagonals of $\mathbf{A}$, and $\mathbf{\Delta}$ the complement of $\hat{\mathbf{A}}$ which contains the remaining ICI terms. Here, we will make the assumption that the terms of $\mathbf{\Delta}$ are random variables with $E\{\mathbf{\Delta}\} = \mathbf{0}$ and covariance matrix $\mathbf{R}_{\mathbf{\Delta}} = E\{\mathbf{\Delta}\mathbf{Delta}^H\}$, where $\hat{\mathbf{A}}$ is deterministic within an OFDM block. The entries $[\mathbf{R}_{\mathbf{\Delta}}]_{k_1,k_2}$, may be computed as

$$\begin{aligned} E\{\delta_{k_1}\delta_{k_2}^*\} \;\; &= \;\; \frac{1}{N^2}\sum_{n_1,n_2=0}^{N-1}\sum_{l_1=,l_2=K}^{L-1} E\{\delta_{l_1}(n_1)\delta_{l_2}^*(n_2)\} \qquad (17)\\ &\quad \times e^{-j2\pi k_1 l_1/N}e^{-j2\pi k_2 l_2/N} \\ &= \;\; \frac{1}{N^2}\sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} r_t\big((n_1-n_2)T_s\big) \qquad (18)\\ &\quad \times \sum_{l=K}^{L-1}\sigma_l^2 e^{-j2\pi(k_1-k_2)l_2/N}. \qquad (19) \end{aligned}$$

Since the zeroth-order Bessel function of the first kind is an even function, it can be approximated by the second order Taylor expansion according to $J_0(2\pi f_d \Delta t) \approx 1 - (\pi x)^2$. Hence, for Jakes' model, the channel autocorrelation function can be approximated by the following expression

$$\begin{aligned} r_t((n_1 - n_2)T_s) &= J_0(2\pi f_d T(n_1 - n_2)) \\ &\simeq 1 - (\pi f_d T(n_1 - n_2)/N)^2. \end{aligned} \tag{20}$$

When $f_d T < 0.03$ the second term can be neglected, i.e. $r_t((n_1 - n_2)T) \simeq 1$, therefore

$$E\{\delta_{k_1}\delta_{k_2}^*\} \simeq \sum_{l=K}^{L-1} \sigma_l^2 e^{-j2\pi(k_1-k_2)l/N}. \tag{21}$$

The received symbols in the frequency-domain can be expressed as

$$\mathbf{y} = \underbrace{\hat{\mathbf{A}}\mathbf{s}}_{\text{signal}} + \underbrace{\mathbf{\Delta s} + \mathbf{w}}_{\text{noise term}} \tag{22}$$

where the AGWN and the model-error have been grouped together as the noise term.

By using Eq. (22) in (10), the block MMSE estimator is expressed as

$$E\left\{\left[(\hat{\mathbf{A}} + \mathbf{\Delta})\mathbf{s} + \mathbf{w}\right]\left[(\hat{\mathbf{A}} + \mathbf{\Delta})\mathbf{s} + \mathbf{w}\right]^H\right\}\mathbf{G} = \tag{23}$$

$$E\left\{\left[(\hat{\mathbf{A}} + \mathbf{\Delta})\mathbf{s} + \mathbf{w}\right]\mathbf{s}^H\right\} \tag{24}$$

$$\Rightarrow \big(\hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I} + E\{\mathbf{\Delta\Delta}^H\} \tag{25}$$

$$+E\{\mathbf{\Delta}\}\hat{\mathbf{A}}^H + \hat{\mathbf{A}}E\{\mathbf{\Delta}\}^H\big)\mathbf{G} = \hat{\mathbf{A}} + E\{\mathbf{\Delta}\} \tag{26}$$

$$\Rightarrow (\hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I} + E\{\mathbf{\Delta\Delta}^H\})\mathbf{G} = \hat{\mathbf{A}} \tag{27}$$

where $\mathbf{G}$ essentially expresses a regularized estimator of (15). In general, the matrix $\bar{\mathbf{R}} \triangleq \hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I} + E\{\mathbf{\Delta\Delta}^H\}$ is a Hermitian Toeplitz and full matrix, hence the solution of (27) requires at least $\mathcal{O}(N^3)$ operations, which is impractical for large N. On the other hand, the solution of the system $\big(\hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I}\big)\mathbf{G} = \hat{\mathbf{A}}$, which has a banded structure, can be obtained very efficiently with complexity $\mathcal{O}(NK^2)$ [17]. Therefore, in order to keep the linear complexity we should approximate the matrix $E\{\mathbf{\Delta\Delta}^H\}$ with a banded one.

Note that under proper conditions, the matrix $E\{\mathbf{\Delta\Delta}^H\}$ can be approximated by a banded matrix. By inspecting Eq. (21), it can be seen thtat the elements of matrix $E\{\mathbf{\Delta\Delta}^H\}$ follow a decaying function with respect to $k_1 - k_2$. Hence, $E\{\mathbf{\Delta\Delta}^H\}$ is a strongly diagonal matrix, which under proper conditions can be effectivelly approximated by a banded matrix, i.e. $E\{\mathbf{\Delta\Delta}^H\} \simeq \mathcal{B}(E\{\mathbf{\Delta\Delta}^H\}) \equiv \mathbf{BB}^H$.

Next, given that the matrix $E\{\mathbf{\Delta\Delta}^H\}$ is banded, we develop an efficient equalization scheme. In that case, Eq. (27) can be expressed as

$$\big(\hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I} + \mathbf{BB}^H\big)\mathbf{G}_b = \hat{\mathbf{A}} \tag{28}$$

Since $\mathbf{R} \triangleq \hat{\mathbf{A}}\hat{\mathbf{A}}^H + \sigma_w^2\mathbf{I} + \mathbf{BB}^H$ is a Hermitian band matrix, the solution of (28) can be obtained through the low-complexity banded $\mathbf{LDL}^H$ factorization [17]. The equalization algorithm is summarized in the following steps :

1) Construct the banded matrix $\mathbf{R}$.
2) Perform banded $\mathbf{LDL}^H$ factorization of $\mathbf{R}$.
3) Compute the equalization matrix $\mathbf{G}_b = (\mathbf{L}^H)^{-1}\left[\mathbf{D}^{-1}\left(\mathbf{L}^{-1}\hat{\mathbf{A}}\right)\right]$.
4) Compute the decision symbols $\hat{\mathbf{s}} = \Pi(\mathbf{G}_b^H\mathbf{y})$.

where $\Pi(\cdot)$ denotes the hard decision operation. At the first step, we construct the banded matrix $\mathbf{R}$, given the banded matrix $\mathbf{B}$. Since all the matrices are banded, with $2K$ off-diagonal elements for each row, Step 1 requires $\mathcal{O}(NK^2)$ operations. At Step 2, taking account that matrix $\mathbf{R}$ is a banded Hermitian matrix, we perform banded $\mathbf{LDL}^H$ factorization with $\mathcal{O}(NK^2)$ complexity order, while at Step 3, we solve (28) very efficiently with $\mathcal{O}(NK)$ operations. Finally, at Step 4 we compute the equalizer output. Therefore, the computational complexity order of the algorithm is $\mathcal{O}(NK^2)$.

V. Simulation Results

In this section, we provide some indicative simulation results in order to evaluate the performance of the proposed method. We consider an OFDM system with $N = 128$ and QPSK modulation. The doubly selective channel, which is given by Eq. (1), is generated according to Jakes' Doppler model [14], consisting by $L = 10$ uncorrelated Rayleigh channel taps, modeled as complex Gaussian variables with zero mean and variance $\sigma_q^2 = J_0(2\pi f_d q)$.

Fig. 1 compares compares bit error rates (BER) versus SNR obtained for the proposed regularized estimator, a conventional single-tap equalizer, the conventional MMSE equalizer, the banded MMSE equalizer with linear complexity [17] and the iterative MMSE equalizer [9]. We consider two cases for normalized Doppler spread $f_d = 0.008$ and $f_d = 0.02$. It can be seen that single-tap equalizer strongly suffers from residual ICI, while the conventional MMSE equalizer achieves the lowest BER for both cases. The banded and the iterative MMSE equalizers diverge from exact one for higher SNRs depending on the maximum normalized Doppler spread. We note that the proposed regularized MMSE equalizer with banded approximation achieves almost the same performance with the iterative MMSE equalizer, while the first one has linear and the second one has quadratic complexity order with respect to the number of the subcarriers.

Fig. 2 compares the BER versus the normalized Doppler spread f_d at a fixed SNR of 30dB. For normalized Doppler spreads below 0.01 all methods have essentially identical performance. For increasing Doppler spread, the BER which achieved by the methods is also increases. Though, the iterative MMSE, banded equalizer and the proposed method have identical performance which outperforms the block banded equalizer.

Fig. 3 compares the BER versus K, which is the length of the banded approximation matrix K. The comparison was

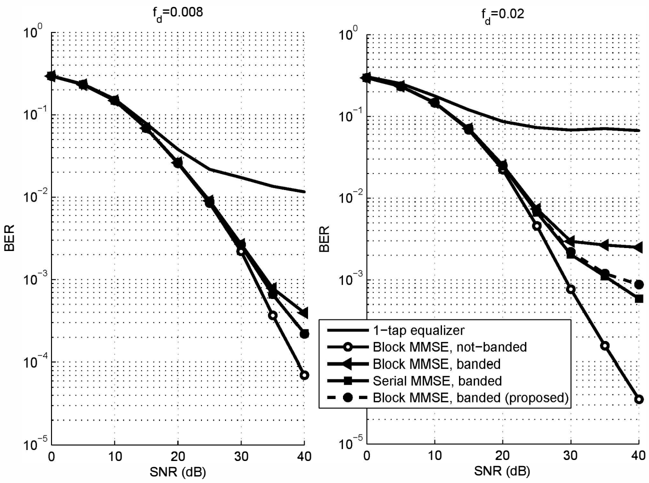

Fig. 1. Comparison of BER versus SNR.

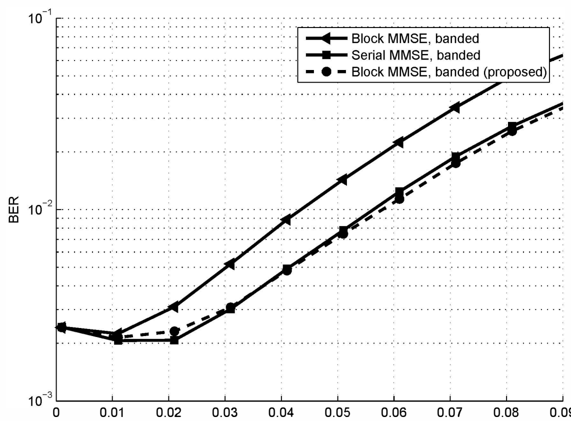

Fig. 2. Comparison of BER versus the normalized Doppler spread.

conducted at a fixed SNR of 30dB. For the affected with K methods, i.e. banded MMSE, iterative MMSE and the proposed method, we observe that performance increases with K. However, the banded MMSE has inferior performance, while the proposed method matches the performance of the iterative MMSE equalizer.

VI. Conclusion

It is known that banded MMSE equalizer, where the channel matrix is replaced with its banded approximation, offers linear complexity in the number of subcarriers at the expense of performance degradation. In order to reduce this performance loss, we have considered the channel truncation as a modeling error, therefore conventional MMSE equalizer is no longer optimal. We have proposed a regularized estimator for MMSE ICI equalization which accounts the mismatched modeling, retaining the banded structure of the system. Therefore, banded $\mathbf{LDL}^H$ factorization can be used in order to compute the regularized estimator with linear complexity. Simulation results verify that the regularized estimation can recover the performance of the exact MMSE equalization to a certain extent, offering performance similar to the iterative MMSE ICI equalizer which has quadratic complexity in the number of subcarriers.

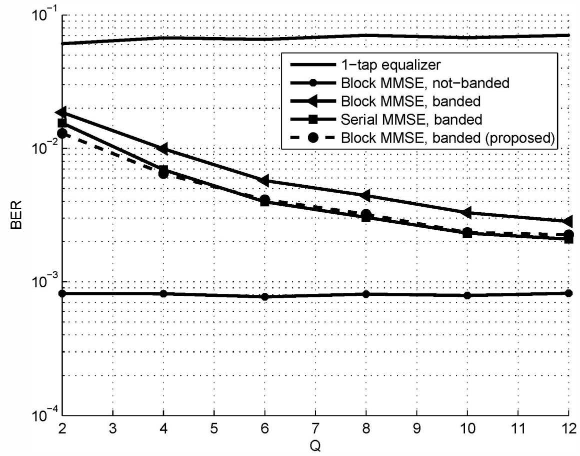

Fig. 3. Comparison of BER versus the length of the banded approximation, SNR=30dB

References

[1] ETSI, "Digital audio broadcasting (dab) to mobile, portable and fixed receivers," *ETS*, vol. 300 401, 1995. [Online]. Available: http://www.etsi.org.

[2] ——, "Digital video broadcasting (dvb): Transmission system for hand-held terminals (dvb-h)," *ETSI*, vol. EN 302 304 V1.1.1, Nov. 2004. [Online]. Available: http://www.etsi.org.

[3] I. P. L. Committee, "The working group on broadband wireless access standards." [Online]. Available: http://grouper.ieee.org/groups/802/16/index.html

[4] Agillent Technologies, "3GPP Long Term Evolution: System Overview, Product development and Test challenges," Application note, iterature number 5989-8139EN, May 19,2008.

[5] T. Wang, J. Proakis, E. Masry, and J. Zeidler, "Performance degradation of ofdm systems due to doppler spreading," *IEEE Transactions on Wireless Communications*, vol. 5, no. 6, pp. 1422 – 1432, Jun. 2006.

[6] P. Robertson and S. Kaiser, "The effects of doppler spreads in ofdm(a) mobile radio systems," in *Vehicular Technology Conference*, vol. 1, 1999, pp. 329 – 333.

[7] 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 Transactions on Communications*, vol. 47, no. 1, pp. 27 –32, Jan. 1999.

[8] 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 Transactions on Communications*, vol. 49, no. 8, pp. 1375 – 1387, 2001.

[9] X. Cai and G. Giannakis, "Bounding performance and suppressing intercarrier interference in wireless mobile ofdm," *IEEE Transactions on Communications*, vol. 51, no. 12, pp. 2047 – 2056, Dec. 2003.

[10] S. U. Hwang, H. Jeong Lee, and J. Seo, "Low complexity iterative ici cancellation and equalization for ofdm systems over doubly selective channels," *IEEE Transactions on Broadcasting*, vol. 55, no. 1, pp. 132 –139, march 2009.

[11] G. Taubock, M. Hampejs, P. Svac, G. Matz, F. Hlawatsch, and K. Grochenig, "Low-complexity ici/isi equalization in doubly dispersive multicarrier systems using a decision-feedback lsqr algorithm," *IEEE Transactions on Signal Processing*, vol. 59, no. 5, pp. 2432 –2436, may 2011.

[12] P. Schniter, "Low-complexity equalization of ofdm in doubly selective channels," *IEEE Transactions on Signal Processing*, vol. 52, no. 4, pp. 1002 – 1011, Apr. 2004.
[13] P. Bello, "Characterization of randomly time-variant linear channels," *IEEE Transactions on Communications Systems*, vol. 11, no. 4, pp. 360 – 393, Dec. 1963.
[14] W. C. Jakes, *Microwave Mobile Channels*. New York: Wiley, 1974.
[15] A. Liavas and D. Tsipouridou, "On the performance of the mismatched mmse and the ls linear equalizers," *IEEE Transactions on Signal Processing*, vol. 55, no. 7, pp. 3302–3311, 2007.
[16] A. N. Tikhonov and V. Y. Arsenin, *Solutions of ill-posed problems*. Washington, D.C.: John Wiley & Sons, New York: V. H. Winston & Sons, 1977.
[17] L. Rugini, P. Banelli, and G. Leus, "Simple equalization of time-varying channels for ofdm," *IEEE Communications Letters*, vol. 9, no. 7, pp. 619–621, 2005.