REDUCED-COMPLEXITY EQUALIZATION FOR MC-CDMA SYSTEMS OVER TIME-VARYING CHANNELS

Luca Rugini [1], *Paolo Banelli* [1], *and Geert Leus* [2]

[1] DIEI, University of Perugia, Via G. Duranti 93, 06125 Perugia, Italy
[2] Dept. of Elect. Eng., Delft University of Technology, 2628 CD Delft, The Netherlands

ABSTRACT

We present a low-complexity equalizer for multicarrier code-division multiple-access (MC-CDMA) downlink systems over time-varying (TV) multipath channels with non-negligible Doppler spread. The equalization algorithm, which is based on a block minimum mean-squared error (MMSE) approach, exploits the band structure of the frequency-domain channel matrix by means of a band LDL^H factorization. The complexity of the proposed block MMSE equalizer is linear in the number of subcarriers, and smaller with respect to a serial MMSE equalizer characterized by a similar performance.

1. INTRODUCTION

Multicarrier techniques based on orthogonal frequency-division multiplexing (OFDM), such as MC-CDMA, gained a lot of attention for wireless mobile communications [1][2]. Indeed, thanks to the cyclic prefix (CP), OFDM systems allow for an easy equalization of time-invariant (TI) frequency-selective channels by means of 1-tap equalizers [1]. Moreover, MC-CDMA systems are able to collect the frequency diversity gain offered by TI frequency-selective channels, since the users' data are spread in the frequency domain. Also in this case the equalization can be very simple, especially if the despreading operation is decoupled from the equalization [2].

However, the request for communications in high-mobility environments suggests that future MC-CDMA designs should take into account not only the frequency-selectivity of the multipath channel, but also its time variability. This requirement greatly complicates the equalization, because the Doppler spread associated with the TV channel generates intercarrier interference (ICI), and hence destroys the orthogonality of the subcarriers [3], as it happens for OFDM [4][5].

Recently, various techniques have been proposed to counteract such ICI effects in multicarrier systems [6]-[11]. It has been shown that nonlinear equalizers based on ICI cancellation generally outperform linear approaches, since they are able to collect the diversity introduced by the time-selectivity of the channel [7]-[10]. Anyway, linear schemes still preserve their importance. First, linear equalizers are usually less complex, and hence more suitable for the downlink. Second, nonlinear schemes often employ a linear equalizer to obtain the tentative decisions used to cancel out the ICI. Third, the cancellation of the ICI due to the interfering users requires the knowledge about the spreading codes of the other users, whereas in the downlink a mobile user generally does not know which users are active.

By focusing on MC-CDMA downlink systems, we present an MMSE equalizer that, in order to reduce complexity, takes advantage of the band structure of the frequency-domain channel matrix. The proposed algorithm firstly performs the LDL^H factorization [12] of the band matrix to be inverted, and then directly solves the associated linear system. The overall complexity, evaluated in number of complex operations, is linear in the number of subcarriers, and quadratic in the channel matrix bandwidth [12], which can be selected to trade off performance for complexity.

The proposed scheme is a block equalizer, which jointly equalizes all the subcarriers, as in [7] and [10]. Those schemes, such as [6][8][9], which separately equalize each subcarrier discarding the data received on the faraway subcarriers, are usually called serial equalizers. The comparison between the proposed block MMSE equalizer and a serial MMSE equalizer derived from [9] evidences a reduction in complexity, while preserving performance.

2. MC-CDMA SYSTEM MODEL

We consider an MC-CDMA downlink system with K users, N subcarriers, N_{A} of which are active and $N_{\mathrm{V}} = N - N_{\mathrm{A}}$ inactive, and a cyclic prefix of length L. Using a notation similar to [1], the ith transmitted block can be expressed as

$$\mathbf{u}[i] = \mathbf{T}_{\mathrm{CP}}\mathbf{F}^{H}\underline{\mathbf{a}}[i], \tag{1}$$

where $\mathbf{u}[i]$ is a column vector of dimension $P = N + L$, $\mathbf{F}$ is the $N \times N$ unitary FFT matrix, $\underline{\mathbf{a}}[i]$ is the $N \times 1$ vector that contains the frequency-domain data, and $\mathbf{T}_{\mathrm{CP}} = [\mathbf{I}_{\mathrm{CP}}^{T}\ \mathbf{I}_{N}^{T}]^{T}$ is the $P \times N$ matrix that inserts the cyclic prefix, where $\mathbf{I}_{\mathrm{CP}}$ contains the last L rows of the identity matrix $\mathbf{I}_{N}$. We assume that at both edges $N_{\mathrm{V}}/2$ subcarriers are reserved as guard frequency bands, as expressed by

$$\underline{\mathbf{a}}[i]^{T} = [\mathbf{0}_{1\times N_{\mathrm{V}}/2}\ \ \mathbf{a}[i]^{T}\ \ \mathbf{0}_{1\times N_{\mathrm{V}}/2}]. \tag{2}$$

The data contained in the $N_{\mathrm{A}} \times 1$ vector $\mathbf{a}[i]$ are obtained by multiplexing the symbols of different users, as expressed by

$$\mathbf{a}[i] = \mathbf{C}\mathbf{s}[i], \tag{3}$$

where $\mathbf{s}[i]$ is the $K \times 1$ vector that contains the data symbols of

This work was partially supported by the Italian Ministry of University and Research under the project "MC-CDMA: an air interface for the 4th generation of wireless systems." Geert Leus was supported in part by NWO-STW under the VICI program (DTC.5893).

the K users, and $\mathbf{C}$ is the $N_{\mathrm{A}} \times K$ matrix whose kth column $\mathbf{c}_k$ contains the unit-norm spreading code of the kth user.

After the parallel-to-serial conversion of (1), the stream $u[n]$ is transmitted through a time-varying multipath channel, where $h[n,l]$ denotes the lth path of the impulse response at time n (including the pulse-shaping effects at transmitter and receiver). Throughout the paper, we assume that the maximum delay spread $\tau_{\max}$ is smaller than the cyclic prefix duration $\tau_{\mathrm{CP}} = LT_{\mathrm{S}}$, where $T_{\mathrm{S}} = T/N$ is the sampling period, and $\Delta_f = 1/T$ is the subcarrier spacing.

At the receiver side, assuming time and frequency synchronization, the received sequence obtained after sampling can be expressed as

$$x[n] = \sum_{l=0}^{L} h[n,l]u[n-l] + w[n] , \qquad (4)$$

where $w[n]$ represents the additive white Gaussian noise (AWGN). By serial-to-parallel conversion, the P received samples relative to the ith block are grouped in the vector $\mathbf{x}[i]$, thus obtaining [1]

$$\mathbf{x}[i] = \mathbf{H}_0[i]\mathbf{u}[i] + \mathbf{H}_1[i]\mathbf{u}[i-1] + \mathbf{w}[i] , \qquad (5)$$

where $\mathbf{H}_0[i]$ is a $P \times P$ band matrix, defined by $[\mathbf{H}_0[i]]_{m,p} = h[m+iP, m-p]$, with upper bandwidth 0 and lower bandwidth L [12], and $\mathbf{H}_1[i]$ is a $P \times P$ upper triangular matrix defined by $[\mathbf{H}_1[i]]_{m,p} = h[m+iP, m-p+P]$. By applying the matrix $\mathbf{R}_{\mathrm{CP}} = [\mathbf{0}_{N\times L}\ \mathbf{I}_N]$ to $\mathbf{x}[i]$ in (5), the cyclic prefix, and hence the interblock interference, is eliminated, thus obtaining, by (1), the $N \times 1$ vector [1]

$$\mathbf{y}[i] = \mathbf{R}_{\mathrm{CP}}\mathbf{x}[i] = \mathbf{H}[i]\mathbf{F}^H \underline{\mathbf{a}}[i] + \mathbf{v}[i] , \qquad (6)$$

where $\mathbf{H}[i] = \mathbf{T}_{\mathrm{CP}}\mathbf{H}_0[i]\mathbf{R}_{\mathrm{CP}}$ is the $N \times N$ time-domain channel matrix defined by $[\mathbf{H}[i]]_{m,p} = h[m+iP, (m-p)_{\mathrm{mod}\,N}]$, and $\mathbf{v}[i] = \mathbf{R}_{\mathrm{CP}}\mathbf{w}[i]$. By applying the FFT at the receiver, we obtain $\underline{\mathbf{z}}[i] = \mathbf{F}\mathbf{y}[i]$, which by (6) can be rewritten as

$$\underline{\mathbf{z}}[i] = \underline{\mathbf{\Lambda}}[i]\underline{\mathbf{a}}[i] + \underline{\mathbf{n}}[i] , \qquad (7)$$

where $\underline{\mathbf{\Lambda}}[i] = \mathbf{F}\mathbf{H}[i]\mathbf{F}^H$ is the $N \times N$ frequency-domain channel matrix, while $\underline{\mathbf{n}}[i] = \mathbf{F}\mathbf{v}[i] = \mathbf{F}\mathbf{R}_{\mathrm{CP}}\mathbf{w}[i]$.

It is easy to verify that if the multipath channel is time invariant, i.e., if $h[n,l]$ does not depend on n, the matrices $\mathbf{H}[i]$ in (6) and $\underline{\mathbf{\Lambda}}[i]$ in (7) will not depend on the block index i. In this case, $\mathbf{H}[i]$ will be circulant, and, accordingly, $\underline{\mathbf{\Lambda}}[i]$ will be diagonal [1]. Anyway, in our case the time variation of the channel cannot be neglected, and hence $\underline{\mathbf{\Lambda}}[i]$ is not diagonal. Specifically, each diagonal of $\underline{\mathbf{\Lambda}}[i]$ is associated with a discrete Doppler frequency, which introduces ICI. Consequently, nontrivial equalization techniques are required.

3. SIMPLE EQUALIZATION OF TV CHANNELS

In this paper, we assume that the receiver of the user k performs equalization and detection separately. Indeed, since we are considering the downlink situation, our main aim is to limit the receiver complexity, which would be higher in case of joint equalization and detection. Moreover, by using the separate approach, the receiver does not require the spreading codes of the other users. In addition, we assume that $\underline{\mathbf{\Lambda}}[i]$ is known to the receiver. In practice, since (7) is similar to that for OFDM systems [6]-[10], $\underline{\mathbf{\Lambda}}[i]$ could be estimated as in [8][10][11].

Let us assume that the equalizer does not make use of the data received on the N_{V} virtual subcarriers, which contain little signal power. In this case, by dropping the block index i for the sake of simplicity, (7) becomes

$$\mathbf{z} = \mathbf{\Lambda}\mathbf{a} + \mathbf{n} , \qquad (8)$$

where $\mathbf{z}$ and $\mathbf{n}$ are $N_{\mathrm{A}} \times 1$ vectors obtained by selecting the middle part of $\underline{\mathbf{z}}[i]$ and $\underline{\mathbf{n}}[i]$, respectively, and $\mathbf{\Lambda}$ is the $N_{\mathrm{A}} \times N_{\mathrm{A}}$ matrix obtained by selecting the central block of $\underline{\mathbf{\Lambda}}[i]$. In order to recover $\mathbf{a}$, several options are possible [7]. In this paper, we will only consider linear approaches. Indeed, since $\mathbf{a}$ is obtained by multiplexing the symbol of different users, nonlinear equalizers that exploit the finite-alphabet property of $\mathbf{a}$ are usually complex, and also require the knowledge of the spreading codes of the other users. Specifically, we focus on linear MMSE equalization, expressed by

$$\hat{\mathbf{a}}_{\mathrm{MMSE}} = \mathbf{\Lambda}^H \left(\mathbf{\Lambda}\mathbf{\Lambda}^H + \gamma^{-1}\mathbf{I}_{N_{\mathrm{A}}}\right)^{-1} \mathbf{z} , \qquad (9)$$

where the signal-to-noise ratio (SNR) $\gamma = \sigma_a^2/\sigma_n^2$ is assumed known to the receiver. Subsequently, by using orthogonal spreading codes, we can recover the transmitted symbol s_k by a simple despreading operation, as expressed by

$$\hat{s}_k = \mathbf{c}_k^H \hat{\mathbf{a}} . \qquad (10)$$

Although the MMSE outperforms other linear criteria [7], the matrix inversion in (9) requires $O(N_{\mathrm{A}}^3)$ flops [12], which represent a significant burden when N_{A} is high. However, in multicarrier systems, as already documented in [6] and [9], TV multipath channels produce a nearly banded channel matrix $\mathbf{\Lambda}$. This behavior can be intuitively explained as follows. The nth column of $\mathbf{\Lambda}$ is obtained by sampling the convolution between the channel Doppler spectrum and the spectrum of the nth subcarrier [9]. Since the subcarrier spectrum decays as $1/\omega^2$ and the Doppler spectrum is bounded by the maximum Doppler frequency f_{D}, the result of this convolution will be practically limited, with significant values only in a neighborhood of the nth element. Therefore, the elements of $\mathbf{\Lambda}$ faraway from the main diagonal are characterized by an almost-zero magnitude. This property can be exploited to further reduce complexity by means of an $\mathrm{LDL^H}$ factorization [12] of Hermitian band matrices.

3.1. Equalization by Band $\mathrm{LDL^H}$ Factorization

Let us approximate the channel matrix $\mathbf{\Lambda}$ with the band matrix $\mathbf{B}$ obtained by selecting the main diagonal, the Q subdiagonals and Q superdiagonals, of $\mathbf{\Lambda}$. Thus $\mathbf{B} = \mathbf{\Lambda} \circ \mathbf{T}^{(Q)}$, where $\circ$ denotes Hadamard (element-wise) product, and $\mathbf{T}^{(Q)}$ is a matrix with lower and upper bandwidth Q [12] and all ones within its band. Accordingly, (9) can be approximated by

$$\hat{\mathbf{a}}_{\mathrm{LDL^H}} = \mathbf{B}^H(\mathbf{B}\mathbf{B}^H + \gamma^{-1}\mathbf{I}_{N_{\mathrm{A}}})^{-1}\mathbf{z} . \qquad (11)$$

Since $\mathbf{M} = \mathbf{B}\mathbf{B}^H + \gamma^{-1}\mathbf{I}_{N_{\mathrm{A}}}$ is a Hermitian band matrix with lower and upper bandwidth $2Q$, $\mathbf{M}^{-1}$ can be obtained by using low-complexity decompositions such as the Cholesky or the $\mathrm{LDL^H}$ factorizations, which are also characterized by a small sensitivity to rounding errors [12]. We consider the $\mathrm{LDL^H}$ factorization because it does not require square roots. The steps of the equalization algorithm are described in Table I.

The parameter Q can be chosen to trade off performance for complexity. Obviously, a larger Q implies a smaller approximation error and hence a performance improvement. On the other hand, the complexity increases due to the higher bandwidth of $\mathbf{M}$. As a rule of thumb, we can adopt $Q \geq \lceil f_{\mathrm{D}}/\Delta_f \rceil + 1$ [9].

TABLE I
EQUALIZATION ALGORITHM

0) Choose Q, and construct the band matrix $\mathbf{B} = \mathbf{\Lambda} \circ \mathbf{T}^{(Q)}$;
1) Construct the band matrix $\mathbf{M} = \mathbf{B}\mathbf{B}^H + \gamma^{-1}\mathbf{I}_{N_A}$;
2) Perform the band LDL^H factorization of $\mathbf{M}$, as expressed by $\mathbf{M} = \mathbf{L}\mathbf{D}\mathbf{L}^H$, where $\mathbf{D}$ is diagonal, and the triangular factor $\mathbf{L}$ has lower bandwidth $2Q$;
3) Solve the system $\mathbf{M}\,\mathbf{d} = \mathbf{z}$ by solving firstly the triangular system $\mathbf{L}\,\mathbf{f} = \mathbf{z}$, secondly the diagonal system $\mathbf{D}\mathbf{g} = \mathbf{f}$, and thirdly the triangular system $\mathbf{L}^H\mathbf{d} = \mathbf{g}$;
4) Calculate $\hat{\mathbf{a}}_{\mathrm{LDL^H}} = \mathbf{B}^H\mathbf{d}$.

3.2. Complexity Analysis

We evaluate the computational cost of the proposed algorithm in terms of complex additions (CA), complex multiplications (CM), and complex divisions (CD). Since $\mathbf{B}$ has lower and upper bandwidth Q, the computation of $[\mathbf{B}\mathbf{B}^H]_{m,p}$ in Step 1 requires $2Q+1-|m-p|$ CM and $2Q-|m-p|$ CA. Hence, taking into account that $\mathbf{B}\mathbf{B}^H$ is Hermitian and neglecting some smaller terms in the complexity expression, Step 1 requires at most $(2Q^2+3Q+1)N_A$ CM and $(2Q^2+Q+1)N_A$ CA. The band LDL^H factorization algorithm of Step 2, reported in Table II, requires $(2Q^2+3Q)N_A$ CM, $(2Q^2+Q)N_A$ CA, and $2QN_A$ CD. This result is obtained by observing that the submatrix $[\mathbf{L}]_{j+1:M,m:j-1}$ is strictly upper triangular. The two band triangular systems of Step 3 can be solved by band forward and backward substitution [12]. Since each algorithm requires $2QN_A$ CM and CA, Step 3 requires $4QN_A$ CM, $4QN_A$ CA, and N_A CD. Moreover, each element of $\hat{\mathbf{a}}_{\mathrm{LDL^H}} = \mathbf{B}^H\mathbf{d}$ needs $2Q+1$ CM and $2Q$ CA. Therefore, the whole algorithm requires approximately $(4Q^2+12Q+2)N_A$ CM, $(4Q^2+8Q+1)N_A$ CA, and $(2Q+1)N_A$ CD, leading to a total of $(8Q^2+22Q+4)N_A$ complex operations.

We now compare the complexity of the proposed block MMSE equalizer with the serial MMSE equalizer expressed by

$$[\hat{\mathbf{a}}_{\mathrm{serial}}]_n = \bar{\mathbf{b}}_n^H(\bar{\mathbf{B}}_n\bar{\mathbf{B}}_n^H + \gamma^{-1}\mathbf{I}_{2Q+1})^{-1}\bar{\mathbf{z}}_n\,, \qquad (12)$$

where n is the subcarrier index, $\bar{\mathbf{z}}_n = [\mathbf{z}]_{n-Q:n+Q}$, $\bar{\mathbf{B}}_n = [\mathbf{B}]_{n-Q:n+Q,\,n-2Q:n+2Q}$, and $\bar{\mathbf{b}}_n = [\bar{\mathbf{B}}_n]_{:,\,2Q+1}$. This equalizer is similar to the one adopted to initialize the iterative symbol MMSE estimator [9] in OFDM systems. Note that Equation (12) is valid only for the active subcarriers with index from $n = 2Q+1$ to $n = N_A - 2Q$. For the edge subcarriers, two approaches are possible, depending on the use of the virtual subcarriers. For simplicity, let us assume that $N_V/2 \geq 2Q$. If we allow for the use of the virtual subcarriers, then the matrix $\bar{\mathbf{B}}_n$ should be extracted from the banded version of $\underline{\mathbf{\Lambda}}$. Moreover, in this case the product $\bar{\mathbf{B}}_n\bar{\mathbf{B}}_n^H$ in (12) should be replaced by $\bar{\mathbf{B}}_n\mathbf{J}_n\bar{\mathbf{B}}_n^H$, where $\mathbf{J}_n$ is a square matrix of size $4Q+1$ obtained from the identity matrix by setting to zero the diagonal elements that correspond to the virtual subcarriers. If we do not allow for the use of the virtual subcarriers, then we can still use (12) if we assume in this formula that indices smaller than 1 or larger than N_A are rounded to 1 or N_A, respectively. In the following, we assume that serial MMSE equalizers make use of the virtual subcarriers.

TABLE II
BAND LDL^H FACTORIZATION

```
L = I_{N_A} ; D = M ∘ I_{N_A} ; v = 0_{N_A×1} ;
for j = 1 : N_A
    m = max{1, j − 2Q} ; M = min{j + 2Q, N_A} ;
    for i = m : j − 1
        [v]_i = [L]*_{j,i} [D]_{i,i} ;
    end
    [v]_j = [M]_{j,j} − [L]_{j,m:j−1} [v]_{m:j−1} ; [D]_{j,j} = [v]_j ;
    [L]_{j+1:M,j} = ([M]_{j+1:M,j} − [L]_{j+1:M,m:j−1} [v]_{m:j−1}) / [v]_j ;
end
```

By using the LDL^H factorization, a careful complexity analysis shows that the serial MMSE equalizer (12) [9] requires $(2Q^2+3Q+1)N_A$ CD, $(4/3\,Q^3+10Q^2+26/3\,Q+2)N_A$ CM, and $(4/3\,Q^3+8Q^2+17/3\,Q+1)N_A$ CA, leading to a total of $(8/3\,Q^3+20Q^2+52/3\,Q+4)N_A$ complex operations. Alternatively, the matrix inversion for index n can be done by reusing the inverse computed for index $n-1$, similarly to the recursive inversion algorithm of [8]. For banded channel matrices, by neglecting the complexity of the first inversion, the serial MMSE equalizer (12) [9] requires about $(14Q^2+15Q+3)N_A$ CM, $(14Q^2+7Q+1)N_A$ CA, and $(2Q+1)N_A$ CD, leading to a total of $(28Q^2+24Q+5)N_A$ complex operations. Therefore, for the serial equalizer (12), the recursive inversion approach is cheaper than LDL^H factorization for $Q \geq 4$. Hence, with respect to the proposed block equalizer, the complexity of the serial equalizer (12) is 1.75 and 2.50 times higher for $Q = 2$ and $Q = 4$, respectively, and asymptotically ($N_A, Q \to \infty$) 3.5 times higher.

3.3. Performance Comparison

We compare by simulations the BER performance of some serial and block equalizers [6]-[10], assuming a banded channel matrix or not. We consider an MC-CDMA system with $N = 128$, $N_A = 96$, $L = 8$, QPSK modulation, and Walsh-Hadamard spreading codes. We assume Rayleigh fading channels with an exponential power delay profile and a Jakes' Doppler spectrum.

In Fig. 1, we assume $K = 32$, $Q = 4$, and $f_D/\Delta_f = 0.15$. Note that this assumption corresponds to a high Doppler scenario. As an example, for a subcarrier spacing $\Delta_f = 20$ kHz, the maximum Doppler frequency is $f_D = 3$ kHz, which is obtained with a mobile speed of $V = 324$ Km/h and a carrier frequency of $f_C = 10$ GHz. Among the banded approaches, the serial MMSE equalizer [9] and the proposed block MMSE equalizer have the best performance. Anyway, the proposed approach allows for a complexity reduction of 60%. Worse performances are obtained using the serial zero-forcing (ZF) equalizer [6], the conventional 1-tap MMSE equalizer [3], introduced in [2] for TI channels, and an approximated block MMSE equalizer with first-order truncation of the Taylor series of $\mathbf{M}^{-1}$ (Eq. 28 of [10]). On the other hand, the non-banded approaches [7][8] outperform the banded ones. However, the complexity is quadratic [8] or even cubic [7] instead of linear in N_A.

It is worth noting that the BER of the serial MMSE non-banded equalizer [8] is lower with respect to the proposed block banded equalizer. This fact points out that the modeling error due to the band matrix approximation $\mathbf{\Lambda} \approx \mathbf{B}$ is higher than the error caused by using a serial instead of a block equalizer. However, in a more realistic scenario, the exact channel matrix $\mathbf{\Lambda}$ is not known, and only an estimated version is available. Although the effect of channel estimation errors is beyond the scope of this paper, we expect that non-banded equalizers should be more sensitive to channel estimation errors than banded equalizers. Indeed, for practical SNR values, the small Doppler components (i.e., those that fall outside the channel matrix band) are difficult to be estimated, and hence it could be better to neglect them.

Fig. 2 shows that the BER performance of the proposed equalizer improves by increasing Q, which can be selected depending on the performance-complexity requirements. Moreover, Fig. 3 illustrates that the considerations deduced from Fig. 1 are valid also when $K = 64$, $f_D / \Delta_f = 0.1$, and $Q = 2$.

4. CONCLUSIONS

We have proposed a block MMSE equalizer for MC-CDMA systems over TV multipath channels. By making use of a band LDL^H factorization, we have shown that the complexity of the proposed equalizer is significantly smaller than for other MMSE approaches, while the performance is still quite good. The sensitivity to different channel estimation techniques could be the subject of future work.

5. REFERENCES

[1] Z. Wang and G. B. Giannakis, "Wireless multicarrier communications: where Fourier meets Shannon," *IEEE Signal Processing Mag.*, vol. 17, pp. 29-48, May 2000.

[2] S. Hara and R. Prasad, "Overview of multicarrier CDMA," *IEEE Commun. Mag.*, vol. 35, pp. 126–133, Dec. 1997.

[3] J.-P. M. G. Linnartz, "Performance analysis of synchronous MC-CDMA in mobile Rayleigh channel with both delay and Doppler spreads," *IEEE Trans. Veh. Technol.*, vol. 50, pp. 1375-1387, Nov. 2001.

[4] M. Russell and G. L. Stüber, "Interchannel interference analysis of OFDM in a mobile environment," in *Proc. IEEE Veh. Tech. Conf.*, Chicago, IL, July 1995, pp. 820-824.

[5] P. Robertson and S. Kaiser, "The effects of Doppler spreads on OFDM(A) mobile radio systems," in *Proc. IEEE Veh. Tech. Conf. Fall*, Amsterdam, The Netherlands, Sept. 1999, pp. 329-333.

[6] 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, pp. 27-32, Jan. 1999.

[7] 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, pp. 1375-1387, Aug. 2001.

[8] X. Cai and G. B. Giannakis, "Bounding performance and suppressing intercarrier interference in wireless mobile OFDM," *IEEE Trans. Commun.*, vol. 51, pp. 2047-2056, Dec. 2003.

[9] P. Schniter, "Low-complexity equalization of OFDM in doubly selective channels," *IEEE Trans. Signal Processing*, vol. 52, pp. 1002-1011, Apr. 2004.

[10] A. Gorokhov and J.-P. Linnartz, "Robust OFDM receivers for dispersive time-varying channels: Equalization and channel acquisition," *IEEE Trans. Commun.*, vol. 52, pp. 572-583, Apr. 2004.

[11] A. Stamoulis, S. N. Diggavi, and N. Al-Dhahir, "Intercarrier interference in MIMO OFDM," *IEEE Trans. Signal Processing*, vol. 50, pp. 2451-2464, Oct. 2002.

[12] G. H. Golub and C. F. Van Loan, *Matrix Computations*, 3rd ed., Johns Hopkins Univ. Press, 1996.

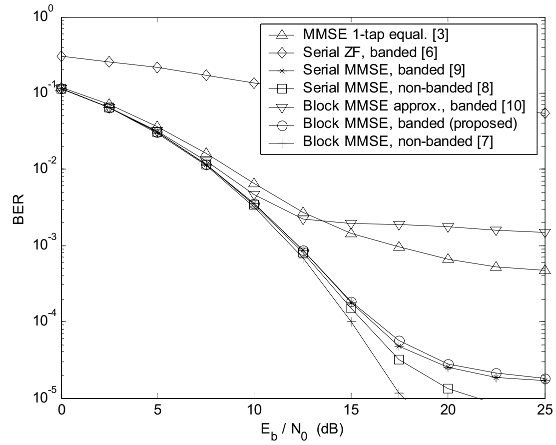

Fig. 1. BER comparison for $K = 32$ and $f_D / \Delta_f = 0.15$ ($Q = 4$ for serial and banded approaches).

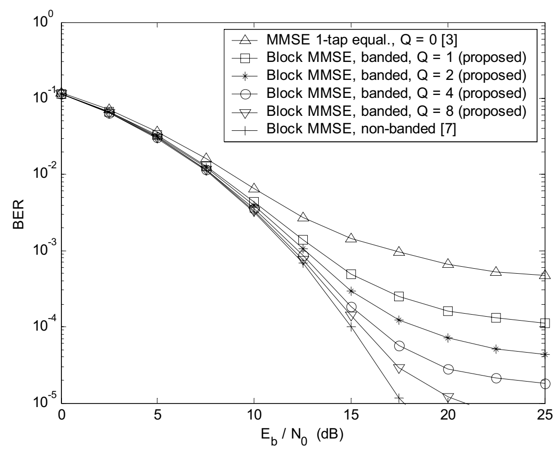

Fig. 2. BER performance of the proposed equalizer as a function of Q ($K = 32$ and $f_D / \Delta_f = 0.15$).

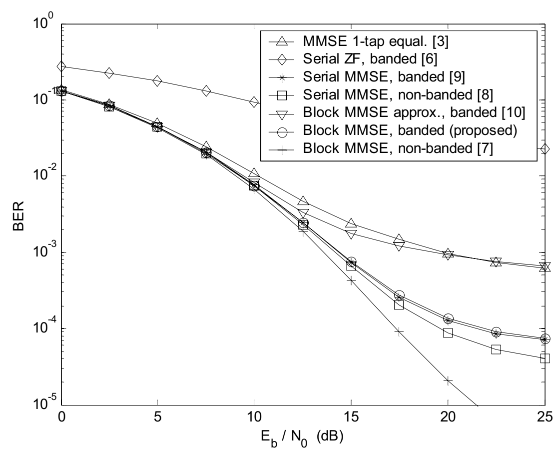

Fig. 3. BER comparison for $K = 64$ and $f_D / \Delta_f = 0.1$ ($Q = 2$ for serial and banded approaches).