MMSE Multiuser Detection for Array Multicarrier DS-CDMA in Fading Channels

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***Abstract*—Reception of asynchronous, multicarrier direct-sequence-code division multiple access (DS-CDMA) in time-varying, multipath radio channels with use of a receiving antenna array is investigated. Interference reducing minimum mean squared error (MMSE) receivers are discussed, and by considering the time-variation of the channel, a modified structure is derived which is efficient for channels experiencing small-scale fading. A blind implementation of this receiver is then proposed. Subspace concepts are applied to formulate a tracking, composite channel vector estimator which operates effectively in fading situations, even when high levels of interference are present. Both the modified MMSE weight matrix and diversity combining weights are generated from these channel estimates. Simulations of the proposed receiver show it to have superior performance over a standard MMSE receiver which is periodically re-evaluated to permit it to follow the channel variations due to small-scale fading. Furthermore, a hybrid MMSE receiver is proposed which applies different processing methods depending on each transmitters mobility, resulting in improved performance.**

***Index Terms*—Antenna arrays, blind estimators, diversity methods, interference suppression, multicarrier CDMA, small-scale fading, space-time processing.**

NOTATION

a	Scalar.
$\underline{a}$	Column vector.
$\mathbb{A}$	Matrix.
$\mathbb{I}_N$	$N \times N$ identity matrix.
$\underline{0}_N$	N element column vector of zeros.
$(\cdot)^T$	Transpose.
$(\cdot)^H$	Hermitian transpose.
$(\cdot)^\dagger$	Pseudoinverse.
$\exp(\underline{a})$	Elementwise exponential.
$\text{diag}\{\cdot\}$	Generate a block diagonal matrix.
$\mathcal{E}\{\cdot\}$	Expectation.
$\otimes$	Kronecker product.
$\odot$	Hadamard product.
$\lfloor\cdot\rfloor$	Round down to integer.
$\mathcal{N}$	Set of natural numbers.
$\mathcal{C}$	Field of complex numbers.

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I. INTRODUCTION

MULTICARRIER direct-sequence code-division multiple access (MC-DS-CDMA) systems are designed to yield a frequency diversity gain, as opposed to a path diversity gain as achieved by single-carrier (SC) DS-CDMA systems. In [1], an MC system is analyzed under the assumption of bandlimited subchannels which experience independent, frequency nonselective fading. For this situation, the performance of MC and SC systems is identical, although the MC receiver realizes a diversity improvement without the use of a RAKE structure.

Considering the more general case of frequency selective subchannels, it was demonstrated in [2] that both path and frequency diversity can be beneficial in an MC system with overlapping subcarrier spectra, although a careful balance between the two sources of diversity is required for optimum performance. This is explained by realizing that the multipath creates intersubcarrier interference which increases the bit error rate (BER), even though the additional path diversity will tend to reduce the BER. Nonetheless, MC-DS-CDMA can outperform SC-DS-CDMA, particularly for channels with a decaying multipath intensity profile, where longer delay paths contain relatively little power [3] or when noncontiguous spectrum for the different subcarriers is available [4]. Hence, MC-DS-CDMA is of interest for future radio systems.

Receiver performance in the presence of cochannel signals is significantly improved if interference cancellation is employed either before or after maximal ratio combining. Minimum mean squared error (MMSE) detection suppresses the multiple access interference (MAI) while still providing frequency diversity [5], [6]. Antenna arrays provide a more powerful alternative means to reduce MAI whilst providing spatial diversity. Joint space-time processing schemes are discussed in [7] and [8]; however, it is assumed that the subchannels are frequency nonselective. In contrast, the authors of [9] consider an array MC-DS-CDMA system which is transmitted through a fully frequency selective radio channel. A blind zero-forcing (ZF) receiver is proposed which cancels both intersymbol interference (ISI) and MAI.

The work presented in this paper removes the assumption that the channel is quasistationary—a partial discussion of the concepts involved was published in [10]. Now, the channel is considered to be slow fading, where the channel coherence time is greater than one symbol period. However, even for a slow fading channel, there may be significant time-variation within the duration of a received burst of data, especially for mobile terminals moving at vehicular speeds. When users have high mobility, the channel coherence time is reduced, and so, the linear ZF and

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MMSE receivers become rapidly time-varying. One effort to address this problem for SC systems applies a subspace tracking receiver to blindly and adaptively implement the ZF or MMSE receivers in an efficient manner [11]. However, the weight vector generated by this approach is still rapidly time-varying, which means that accuracy suffers as the fading rate increases. Alternative approaches attempt to reduce the rate of change of the detector by performing interference cancellation prior to diversity combination. The adaptive approaches in [12]–[14] use a modified MMSE criterion to improve performance in fading situations. Differences between the papers occur according to whether only phase variation is compensated or both phase and amplitude variation are compensated. The disadvantage of the systems described in [12]–[14] is that training signals are assumed to be regularly available for channel estimation purposes, which reduces the rate of payload data transfer.

The first receiver described in the sequel is designed for efficient reception of array MC-DS-CDMA for frequency selective subchannels which experience time-variation. In Section II, a comprehensive model of the system is developed, while the construction of the receiver weight matrix used to detect all of the transmitting users' signals is discussed in Section III. This weight matrix is made up of two parts. The first operation is calculated infrequently in a block processing mode and is designed to reduce interference. Second, there is a low complexity operation which is updated every symbol period and applied to coherently combine the signal contributions associated with each user. A blind implementation of the receiver is proposed in Section IV, which uses a recursive channel estimator that applies subspace techniques.

Section V considers the more general case when the transmitting users move with a range of speeds so that there is a variety of fading rates. A hybrid MMSE receiver is proposed which applies the most suitable processing for each user depending on their rate of channel time-variation. For practical application, this hybrid receiver can be implemented using blind channel estimation techniques. Both receiver designs are evaluated by a selection of simulation studies in Section V, and finally, the paper is concluded in Section VI.

II. System Model

Considering transmission from the ith mobile user, the baseband SC-DS-CDMA signal is modeled as

$$m_i(t) = \sum_{n=-\infty}^{+\infty} b_i[n] c_{PN,i}(t - nT_{cs}). \tag{1}$$

In this expression, $b_i[n]$ is the nth transmitted symbol, and T_{cs} is the symbol period. Additionally, $c_{PN,i}(t)$ is the pseudo-noise (PN) signal for the ith user, which is created by convolving the PN-sequence $\{\alpha_i[m],\ m = 0, 1, \ldots, N_c - 1\}$ of length N_c with a rectangular chip waveform. Therefore, the chip period is $T_c = T_{cs}/N_c$ for this short code system. The SC-DS-CDMA signal now modulates N_{sc} subcarriers, with the frequency of the kth subcarrier denoted as F_k. A MC-DS-CDMA signal suitable for transmission is generated by summing the modulated subcarriers and upconverting to the carrier frequency F_c, as described by

$$y_i(t) = \sum_{k=0}^{N_{sc}-1} \sqrt{P_i} \exp(j(2\pi F_c t + \zeta_i)) \exp(j2\pi F_k t) m_i(t) \tag{2}$$

where P_i is the transmitted power, and ζ_i is a random phase offset relative to the base station receiver.

Each users transmission propagates through a radio channel, which is assumed to be time-varying and multipath dispersive. At the base station, there is an array, with a given geometry, of N receiving antennas so that the channel impulse response vector for the kth subcarrier and jth path of the ith user is represented by

$$\underline{c}_{ijk}(t) = \beta_{ijk} \exp(j2\pi f_{ijk} t) \underline{S}_{ijk} \delta(t - \tau_{ij}) \in \mathcal{C}^{N \times 1}. \tag{3}$$

β_{ijk} is the complex path coefficient which sets the path magnitude and encompasses a random phase shift, while $\exp(j2\pi f_{ijk} t)$ explicitly models the Doppler shift due to motion of the transmitter. If v_{ij} is the speed of motion for the ith user's jth path toward the base station, then the Doppler frequency for the kth subcarrier is $f_{ijk} = (F_c + F_k)(v_{ij}/c)$, where c is the speed of light. The path delay is denoted as τ_{ij} and will be the same for all subcarriers for a particular path. Finally, $\underline{S}_{ijk}$ is the array manifold vector (array response vector), which is dependent on the array geometry, subcarrier frequency, and direction of arrival of the particular path.

Generally, the individual paths modeled by (3) are not all expected to be temporally resolvable, e.g., when several paths arrive within one sample period. In this case, a single temporally resolvable path is composed of several unresolvable paths. However, each path has a different direction of arrival; therefore, temporally unresolvable paths can be resolved in the spatial domain by the array system. With regard to time-variation, the parameters β_{ijk}, f_{ijk}, $\underline{S}_{ijk}$, and τ_{ij} are slowly changing, but they are considered to be constant over a block of B symbols. In contrast, the phase of each path is subject to rapid changes due to the Doppler shift; therefore, the channel cannot be considered to be quasistationary, even for the duration of a short block of symbols.

All of the superimposed radio signals transmitted from a total of M asynchronous users are received at the base station array, and the resultant signal is downconverted to baseband. If the radio channel for the ith user consists of K_i propagating paths then the received signal vector at point F in Fig. 1 is

$$\underline{x}(t) = \sum_{i=1}^{M} \sum_{j=1}^{K_i} \sum_{k=0}^{N_{sc}-1} \beta_{ijk} \exp(j2\pi f_{ijk} t) \underline{S}_{ijk} \cdot \exp(j2\pi F_k (t - \tau_{ij})) m_i(t - \tau_{ij}) + \underline{n}(t). \tag{4}$$

Here, the path coefficient β_{ijk} has absorbed the complex factor $\sqrt{P_i} \exp(j(\zeta_i - 2\pi F_c \tau_{ij}))$, and $\underline{n}(t)$ represents complex additive white Gaussian noise (AWGN) of power σ^2.

The continuous-time received signal from each antenna is discretized by a sampler operating at a rate of $1/T_s$, where the

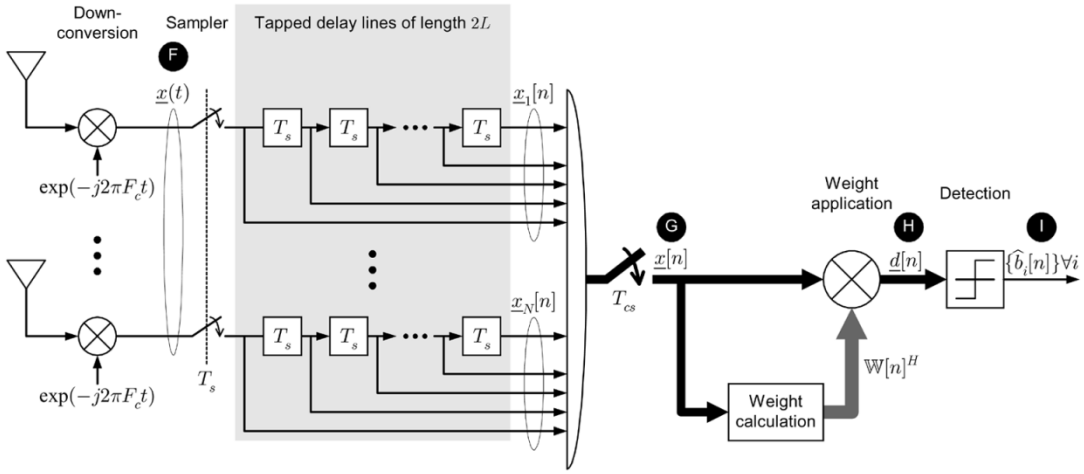


Fig. 1. Base station array receiver block diagram.

sampling period is $T_s = T_c/qN_{sc}$, and $q \in \mathcal{N}$ is the oversampling factor. Each path delay τ_{ij} is now comprised of two parts: the integer delay $l_{ij}T_s$ and the fractional delay $\rho_{ij}T_s$, where $l_{ij} \in \{0, 1, \ldots, L-1\}$, and $\rho_{ij} \in [0, 1)$. Note that $L = qN_cN_{sc}$ is the number of temporal samples per symbol. For simplicity of representation, we assume that the channel delay spread is less that one symbol period. To accommodate the lack of synchronization, a bank of N tapped delay lines of length $2L$ store the data prior to the formation of the discrete received signal vector $\underline{x}[n] \in \mathcal{C}^{2NL\times 1}$ at point G in Fig. 1. This is achieved by concatenating the contents of the tapped delay lines every symbol period.

Before an expression for the discrete vector $\underline{x}[n]$ can be written, the time-varying multicarrier spatio-temporal array (STAR) manifold vector for the kth subcarrier, jth path of the ith user is defined as

$$\underline{\mathfrak{h}}_{ijk}[n] \triangleq \underline{S}_{ijk} \otimes \left(\mathbb{J}^{l_{ij}}\underline{\mathfrak{a}}_{ik}[l_{ij}] \odot \underline{\xi}_{ijk}[n]\right) \quad \in \mathcal{C}^{2NL\times 1} \tag{5}$$

where $\otimes$ is the Kronecker product, and $\odot$ is the Hadamard product. The fundamental property of the STAR manifold vector is that it maps the channel parameters associated with a particular path to a vector in the $2NL$-dimension observation space. Within (5), $\mathbb{J}^{l_{ij}}\underline{\mathfrak{a}}_{ik}[l_{ij}]$ is the delayed temporal vector, and $\underline{\xi}_{ijk}[n]$ is the Doppler vector, and both will now be described in more detail. First, the temporal vector for paths arriving with an integer delay of ℓ sample periods is

$$\underline{\mathfrak{a}}_{ik}[\ell] = \left[\mathfrak{a}_{ik}[0,\ell], \mathfrak{a}_{ik}[1,\ell], \ldots, \mathfrak{a}_{ik}[L-1,\ell], \underline{0}_L^T\right]^T \tag{6}$$

with elements given by

$$\mathfrak{a}_{ik}[m,\ell] = \alpha_i \left[\left\lfloor \frac{m}{qN_{sc}} \right\rfloor\right] \exp(j2\pi F_k(m-\ell)T_s). \tag{7}$$

This temporal vector accounts for the unique PN-sequence of the ith user, as well as for the phase progression for the kth subcarrier. To account for the propagation delay, the temporal vector is shifted by l_{ij} elements, which is an amount equal to the duration of the integer delay for the particular path. This is achieved by premultiplying the temporal vector with the shift matrix $\mathbb{J}^{l_{ij}}$, the $2L \times 2L$ matrix $\mathbb{J}$ being defined as

$$\mathbb{J} = \begin{bmatrix} \underline{0}_{2L-1}^T & 0 \\ \mathbb{I}_{2L-1} & \underline{0}_{2L-1} \end{bmatrix}. \tag{8}$$

As an example, applying $\mathbb{J}^\ell$ to a column vector shifts the elements down by ℓ elements, whereas premultiplying with $(\mathbb{J}^T)^\ell$ shifts the elements up by ℓ elements. The outstanding definition required for (5) is the Doppler vector, which is a time-varying quantity expressed as

$$\underline{\xi}_{ijk}[n] = \begin{bmatrix} 1 \\ \exp(j2\pi f_{ijk}T_s) \\ \vdots \\ \exp(j2\pi f_{ijk}(2L-1)T_s) \end{bmatrix} \exp(j2\pi f_{ijk}nT_{cs}). \tag{9}$$

This vector models the sampled Doppler shift for the particular path.

Using the multicarrier STAR manifold vector, the discrete received signal vector can be written as

$$\underline{x}[n] = \sum_{i=1}^{M}\sum_{j=1}^{K_i}\sum_{k=0}^{N_{sc}-1} \beta_{ijk} \left(\overbrace{\mathbb{J}^{(prev)}\underline{\mathfrak{h}}_{ijk}[n-1]b_i[n-1]}^{\text{previous}} + \overbrace{\underline{\mathfrak{h}}_{ijk}[n]b_i[n]}^{\text{current}} + \underbrace{\mathbb{J}^{(next)}\underline{\mathfrak{h}}_{ijk}[n+1]b_i[n+1]}_{\text{next}} \right) + \underline{n}[n] \tag{10}$$

where it is noted that due to user asynchronism $\underline{x}[n]$ contains contributions from the previous, current, and next symbols for all users. Time shifts for the previous and next symbols are generated by expanding the shift matrix so that $\mathbb{J}^{(prev)} = \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^L\right)$ and $\mathbb{J}^{(next)} = \left(\mathbb{I}_N \otimes \mathbb{J}^L\right)$. A simpler representation of $\underline{x}[n]$ is formulated by defining matrix $\mathbb{H}_i[n] \in \mathcal{C}^{2NL\times K_iN_{sc}}$ associated with the ith user, which has columns containing the multicarrier STAR manifold vectors $\underline{\mathfrak{h}}_{ijk}[n]\ \forall\ j,k$. Furthermore, if vector $\underline{\beta}_i \in \mathcal{C}^{K_iN_{sc}\times 1}$ contains

the path coefficients $\beta_{ijk}\ \forall\, j,k$, then $\underline{x}[n]$ may be written in the form of (11).

$$\underline{x}[n] = \sum_{i=1}^{M} \left[\mathbb{J}^{(prev)} \mathbb{H}_i[n-1]\underline{\beta}_i, \mathbb{H}_i[n]\underline{\beta}_i, \mathbb{J}^{(next)} \mathbb{H}_i[n+1]\underline{\beta}_i \right] \cdot \begin{bmatrix} b_i[n-1] \\ b_i[n] \\ b_i[n+1] \end{bmatrix} + \underline{n}[n]. \quad (11)$$

A final simplification of the model is possible by defining the composite channel vector for the ith user as

$$\underline{h}_i[n] \triangleq \mathbb{H}_i[n]\underline{\beta}_i \quad \in \mathcal{C}^{2NL\times 1}. \quad (12)$$

This important vector contains all of the effects of the radio channel applied to symbol $b_i[n]$. The discrete received signal vector is now given by

$$\underline{x}[n] = \mathbb{H}[n]\underline{b}[n] + \underline{n}[n] \quad (13)$$

and contains the terms

$$\begin{aligned} \mathbb{H}[n] &= \left[\mathbb{J}^{(prev)}\overline{\mathbb{H}}[n-1], \overline{\mathbb{H}}[n], \mathbb{J}^{(next)}\overline{\mathbb{H}}[n+1] \right], \\ \overline{\mathbb{H}}[n] &= [\underline{h}_1[n], \underline{h}_2[n], \ldots, \underline{h}_M[n]] \\ \underline{b}[n] &= \left[\underline{b}_M[n-1]^T, \underline{b}_M[n]^T, \underline{b}_M[n+1]^T \right]^T \\ \underline{b}_M[n] &= [b_1[n], b_2[n], \ldots, b_M[n]]^T. \end{aligned} \quad (14)$$

At point H in Fig. 1, a decision variable vector $\underline{d}[n] = \mathbb{W}[n]^H \underline{x}[n]$ is observed, and at point I, the decision variables are resolved into detected symbols.

III. MMSE Receivers for Fading Channels

Standard multiuser receivers [15] are applicable for both quasistationary and fading environments, but in the case of a fading channel, the weight matrix becomes time-varying. Based on (13), the MMSE receiver during the nth symbol period is

$$\mathbb{W}[n] = \mathbb{H}[n]\left(\mathbb{H}[n]^H\mathbb{H}[n] + \sigma^2\mathbb{I}_{3M}\right)^{-1} \in \mathcal{C}^{2NL\times 3M}. \quad (15)$$

The time-varying nature of the weight matrix is a problem because of the complexity of calculating $\mathbb{W}[n]$ every symbol period. $\mathbb{W}[n]$ can be considered to be constant for the channel coherence time, but this may only be of the order of ten symbol periods when the radio channels are rapidly changing due to high transmitter mobility. Therefore, (15) is considered an expensive solution for fading environments. In this section, an alternative MMSE receiver is described which is valid for a time period significantly greater than the channel coherence time, thus reducing the update rate of the receiver.

First, we consider the composite channel vector for the ith user $\underline{h}_i[n] = \mathbb{H}_i[n]\underline{\beta}_i$, which can be rewritten in the following form:

$$\underline{h}_i[n] = \mathbb{C}_i\underline{\mathfrak{s}}_i[n] \quad (16)$$

where the code matrix for the ith user is $\mathbb{C}_i \in \mathcal{C}^{2NL\times NN_{sc}L}$ and consists of the concatenation of all the code submatrices $\mathbb{C}_{ik}[\ell]$ for $k = 0,1,\ldots,N_{sc}-1$ and $\ell = 0,1,\ldots,L-1$. Additionally, the code submatrices are defined by

$$\mathbb{C}_{ik}[\ell] = \mathbb{I}_N \otimes \left(\mathbb{J}^{\ell}\underline{\mathfrak{a}}_{ik}[\ell]\right) \quad \in \mathcal{C}^{2NL\times N}. \quad (17)$$

The vector $\underline{\mathfrak{s}}_i[n]$ is formed by stacking up all of the subvectors $\underline{\mathfrak{s}}_{ik}[n,\ell]$ for $k = 0,1,\ldots,N_{sc}-1$ and $\ell = 0,1,\ldots,L-1$ with

$$\underline{s}_{ik}[n,\ell] = \sum_{\forall j \text{ where } l_{ij}=\ell} \beta_{ijk}\underline{S}_{ijk}\exp(j2\pi f_{ijk}nT_{cs}). \quad (18)$$

It should be noted that for any particular integer delay ℓT_s, there may be no paths present, a single path, or multiple paths. Furthermore, in (16), the Doppler shift within a single symbol period is considered negligible and has been ignored. This is reasonable because fast fading (where the channel coherence time is less than one symbol period) is not expected to occur.

Assuming there to be D temporally resolvable paths for every user, each being made up of multiple unresolvable paths with different azimuth directions and Doppler shifts, then (16) is simplified to

$$\underline{h}_i[n] = \mathbb{V}_i\underline{\mathfrak{u}}_i[n] \quad (19)$$

where $\underline{\mathfrak{u}}_i[n] \in \mathcal{C}^{NN_{sc}D\times 1}$ consists of the stacked nonzero elements of $\underline{\mathfrak{s}}_i[n]$, and $\mathbb{V}_i \in \mathcal{C}^{2NL\times NN_{sc}D}$ contains the corresponding columns of $\mathbb{C}_i$ at the D time delays. Consequently, $\underline{\mathfrak{u}}_i[n]$ contains the channel coefficients for each antenna and subcarrier at each of the D temporally resolvable delays.

The total received signal vector can now be written as

$$\underline{x}[n] = \mathbb{V}\mathbb{U}[n]\underline{b}[n] + \underline{n}[n] \quad (20)$$

with $\mathbb{V}$ and $\mathbb{U}[n]$ defined by

$$\begin{aligned} \mathbb{V} &= [\mathbb{J}^{(prev)}\overline{\mathbb{V}}, \overline{\mathbb{V}}, \mathbb{J}^{(next)}\overline{\mathbb{V}}] \\ \overline{\mathbb{V}} &= [\mathbb{V}_1, \mathbb{V}_2, \ldots, \mathbb{V}_M] \\ \mathbb{U}[n] &= \text{diag}\{[\overline{\mathbb{U}}[n-1], \overline{\mathbb{U}}[n], \overline{\mathbb{U}}[n+1]]\} \\ \overline{\mathbb{U}}[n] &= \text{diag}\{[\underline{\mathfrak{u}}_1[n], \underline{\mathfrak{u}}_2[n], \ldots, \underline{\mathfrak{u}}_M[n]]\}. \end{aligned} \quad (21)$$

Substituting for $\underline{u}[n] = \mathbb{U}[n]\underline{b}[n]$ in (20) produces

$$\underline{x}[n] = \mathbb{V}\underline{u}[n] + \underline{n}[n]. \quad (22)$$

This equation is in the general linear form so the appropriate modified MMSE receiver is formulated as

$$\mathbb{W} = \mathbb{V}\left(\mathbb{V}^H\mathbb{V} + \sigma^2\mathbb{R}_{uu}^{-1}\right)^{-1} \in \mathcal{C}^{2NL\times 3MNN_{sc}D}. \quad (23)$$

Notice that this weight matrix does not change over periods of time when the radio channel undergoes fading so that it does not need to be recalculated every symbol period. In fact, the weight matrix is valid for as long as the time of arrivals (TOAs) of the temporally resolvable paths remain approximately constant relative to the sampling period, i.e., a time duration significantly greater than the channel coherence time. Under the assumption that the transmitted symbols have unit energy, then covariance matrix $\mathbb{R}_{uu} = \mathcal{E}\{\underline{u}[n]\underline{u}[n]^H\} = \mathcal{E}\{\mathbb{U}[n]\mathbb{U}[n]^H\}$ may be estimated from the long term properties of the channel and depends

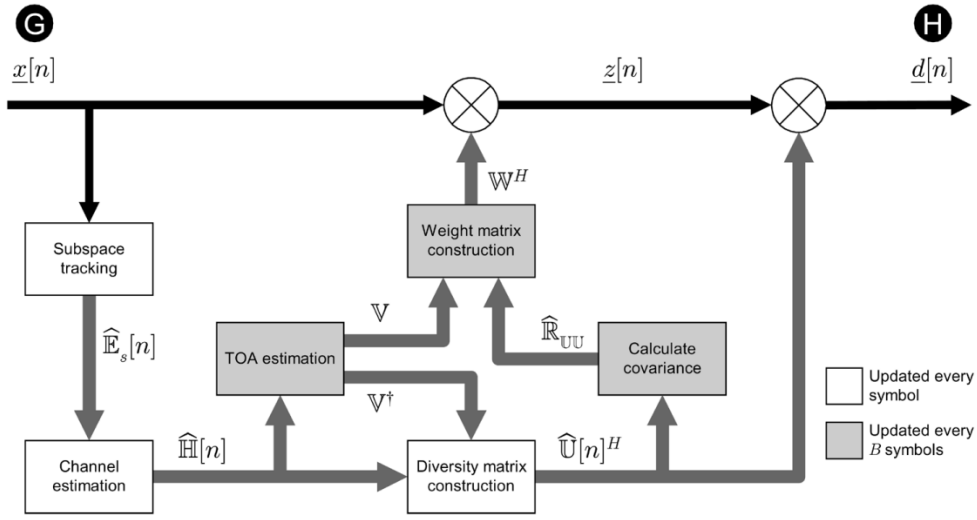


Fig. 2. Structure of the blind modified MMSE receiver for small-scale fading channels.

on the average powers of each users path coefficients. Generally, $\mathbb{R}_{uu}$ will have a block diagonal structure, but if the subcarrier fading is decorrelated and the array antennas are sufficiently separated to decorrelate the fading then $\mathbb{R}_{uu}$ becomes diagonally dominated. In this case, $\mathbb{R}_{uu}$ can be simplified to a diagonal matrix, which is then trivial to invert.

The output of the modified MMSE receiver is given by

$$\underline{z}[n] = \mathbb{W}^H \underline{x}[n] \in \mathcal{C}^{3MNN_{sc}D\times 1}. \tag{24}$$

From $\underline{z}[n]$, the decision variables for the previous, current, and next symbols corresponding to the M users are calculated by diversity combining the signals from different antennas, paths, and subcarriers,

$$\underline{d}[n] = \mathbb{U}[n]^H \underline{z}[n] \in \mathcal{C}^{3M\times 1}. \tag{25}$$

The complete receiver structure consists of a fading invariant MMSE multiuser detector followed by a time-varying diversity combining matrix. Therefore, the computationally intensive part of the receiver only needs to be occasionally updated while the simple combining matrix requires updating every symbol period.

IV. Blind Implementation of the Modified MMSE Receiver

In this section, the blind multiuser receiver shown in Fig. 2 is proposed. There are two matrices required to formulate the modified MMSE receiver, namely, $\mathbb{V}$ and $\mathbb{U}[n]$. It is claimed that both matrices can be found from the estimated composite channel vectors for all users $\widehat{\underline{h}}_i[n]\ \forall\ i$. In [9], it was demonstrated that the composite channel vector for each user could be estimated in a quasistationary environment by using subspace techniques and applying the theorem of alternating projection [16], [17]. A modified, tracking version of this procedure is presented here to provide estimates under time-varying conditions. The method employs two constraint subspaces; the first is provided by the code matrix $\mathbb{C}_i$, where the projector into this space is

$$\mathbb{P}_i = \mathbb{C}_i \left(\mathbb{C}_i^H \mathbb{C}_i\right)^\dagger \mathbb{C}_i^H. \tag{26}$$

A second constraint from the signal subspace of $\underline{x}[n]$ produces the following time-varying projection operation

$$\mathbb{P}_s[n] = \widehat{\mathbb{E}}_s[n]\widehat{\mathbb{E}}_s[n]^H. \tag{27}$$

In this expression, $\widehat{\mathbb{E}}_s[n]$ is an estimated orthonormal basis of the signal subspace which spans the channel matrix $\mathbb{H}[n]$ and is valid during the nth symbol period. Many subspace tracking algorithms are available to generate $\widehat{\mathbb{E}}_s[n]$, but a particularly suitable technique is refinement-only fast subspace tracking (RO-FST), which has low complexity and is robust in operation [18]. This algorithm computes a precisely orthonormal basis $\widehat{\mathbb{E}}_s[n]$ from the previous estimate $\widehat{\mathbb{E}}_s[n-1]$ and the current received signal $\underline{x}[n]$. In operation, RO-FST applies a subspace decomposition to the exponentially weighted data covariance matrix. An alternative approach is to directly track the signal subspace projector $\mathbb{P}_s[n]$, which can result in lower complexity [19].

Recursive estimation of the composite channel vector for the ith user is calculated from

$$\widehat{\underline{h}}_i[n] = \mathbb{P}_i\mathbb{P}_s[n]\widehat{\underline{h}}_i[n-1] \tag{28}$$

where the initial value of $\widehat{\underline{h}}_i[n]$ is any arbitrarily chosen nonzero vector. It is noted that $\widehat{\underline{h}}_i[n]$ only estimates $\underline{h}_i[n]$ up to a complex scalar. Equation (28) is repeatedly applied for all M users so that an estimate of the total composite channel matrix, $\widehat{\mathbb{H}}[n]$ is generated.

Matrix $\mathbb{V}$ is based entirely on the PN-codes of the active users, the structure of the multicarrier signalling, and the TOAs of the D temporally resolvable paths for each user, which are generally unknown. It is sensible to estimate the TOAs for the ith user directly from $\widehat{\underline{h}}_i[n]$ rather than $\underline{x}[n]$ because $\widehat{\underline{h}}_i[n]$ does not contain any MAI in the ideal case, whereas in practice, only a small residual amount of interference will remain. If the vector $\widehat{\underline{h}}_i^{(p)}[n] \in \mathcal{C}^{2L\times 1}$ denotes the subvector of $\widehat{\underline{h}}_i[n]$ containing elements only concerned with the pth antenna, then the vector of correlation coefficient magnitudes between $\widehat{\underline{h}}_i^{(p)}[n]$ and the delayed temporal vector for the kth subcarrier is given by

$$\underline{\nu}_{ik}^{(p)}[n] = \left[\nu_{ik}^{(p)}[n,0], \nu_{ik}^{(p)}[n,1], \ldots, \nu_{ik}^{(p)}[n,L-1]\right]^T \tag{29}$$

where

$$\nu_{ik}^{(p)}[n,\ell] = \left| \left(\mathbb{J}^{\ell} \underline{\mathfrak{a}}_{ik}[\ell] \right)^{H} \widehat{\underline{h}}_{i}^{(p)}[n] \right|. \tag{30}$$

Averaging over all subcarriers and antennas—so that diversity is provided in a fading situation—produces a spectrum where peaks correspond to estimated TOAs

$$\underline{\nu}_{i}[n] = \frac{1}{NN_{sc}} \sum_{p=1}^{N} \sum_{k=0}^{N_{sc}-1} \underline{\nu}_{ik}^{(p)}[n]. \tag{31}$$

This spectrum is expected to be noisy due to it being calculated over a single symbol period using the *estimated* composite channel vector. Consequently, the D strongest peaks of $\underline{\nu}_i[n]$ may not be located with sufficient accuracy for reliable TOA estimation. However, it is reasonable to assume that the TOAs remain constant over a block of B symbols of total duration greater than the channel coherence time. This is because of the relatively slow rate of change of path TOA compared to the fading. Consequently, a histogram of peak positions in $\underline{\nu}_i[n]$ can be built up. The final accurately estimated TOAs for the ith user are taken from this histogram. This procedure is repeated for all active users, and combined with knowledge of $\mathbb{C}_i \; \forall \, i$ means that $\mathbb{V}_i \; \forall \, i$, and hence, $\mathbb{V}$ can be formed.

Considering the second unknown matrix $\mathbb{U}[n]$, this is a rapidly time-varying quantity that must be tracked as the channel fades. The component vectors of the estimated matrix $\widehat{\mathbb{U}}[n]$ are calculated using

$$\widehat{\underline{\mathfrak{u}}}_{i}[n] = \mathbb{V}_{i}^{\dagger} \widehat{\underline{h}}_{i}[n]. \tag{32}$$

This is justified by (19), and it is necessary to use the pseudoinverse because $\mathbb{V}_i$ is a tall but thin matrix. The pseudoinverse operation is not trivial in terms of complexity, but because $\mathbb{V}_i$ is constant over a block of B symbols, $\mathbb{V}_i^{\dagger}$ only needs to be periodically updated. Therefore, the computational requirement to update $\widehat{\mathbb{U}}[n]$ every symbol period is limited to a single matrix/vector multiplication per user. In Fig. 2, the white boxes represent operations which are updated every symbol period, and the shaded boxes are for operations updated once per block of B symbols.

With estimates of both $\mathbb{V}$ and $\mathbb{U}[n]$, it is possible to use (23) to form the modified MMSE weight matrix, which is applied to the received signal vector according to (24). To generate the receiver weight matrix, additional knowledge of the AWGN power is required, and this is provided by the updating process of the RO-FST subspace tracking algorithm. After application of the weight matrix, the output signal is diversity combined following (25) to produce the required decision variable vector $\underline{d}[n]$. Differentially coherent modulation must be used because the vector of diversity combining weights for each user is only known up to a complex scalar when blind estimation is used. However, the use of differential modulation further increases the receivers robustness to time-variation.

V. Mixed High and Low Mobility Transmitters

In a general mobile communications system, the transmitters will have a range of speeds ranging from pedestrian to fast vehicular. Ideally, the radio receiver will take into account a user's mobility and process its signal accordingly. Assuming the processing to be applied to blocks of B received symbols, then (15) is most appropriate if all users are "slow" and assumed to have quasistationary channels, whereas (23) is more suitable if all users are "fast" with significant channel variation within the block duration. Now, assume that the M active users can be partitioned so that M_s users are considered slowly moving and the other M_f users are fast, where the threshold between fast and slow depends on the time spanned by B symbols and the channel characteristics. Using the subscripts s and f to denote terms related to slower and faster moving users, respectively, the received signal vector may be expressed as

$$\begin{aligned} \underline{x}[n] &= [\mathbb{H}_s, \mathbb{H}_f[n]] \begin{bmatrix} \underline{b}_s[n] \\ \underline{b}_f[n] \end{bmatrix} + \underline{n}[n] \\ &= \mathbb{H}_s \underline{b}_s[n] + \mathbb{V}_f \mathbb{U}_f[n] \underline{b}_f[n] + \underline{n}[n] \end{aligned} \tag{33}$$

where the substitution $\mathbb{H}_f[n] = \mathbb{V}_f \mathbb{U}_f[n]$ is made for the fast-varying channel matrix, analogous to the step taken in (20). Defining the matrices

$$\begin{aligned} \mathbb{L} &= [\mathbb{H}_s, \mathbb{V}_f] \\ \mathbb{T}[n] &= \begin{bmatrix} \mathbb{I}_{3M_s} & \mathbb{O}_1 \\ \mathbb{O}_2 & \mathbb{U}_f[n] \end{bmatrix} \end{aligned} \tag{34}$$

where $\mathbb{O}_1$ is a $3M_s \times 3M_f$ matrix of zeros, $\mathbb{O}_2$ is a $3Q \times 3M_s$ matrix of zeros and $Q = M_f N N_{sc} D$, then the received signal vector becomes

$$\underline{x}[n] = \mathbb{L}\mathbb{T}[n] \underline{b}[n] + \underline{n}[n]. \tag{35}$$

Substituting for

$$\underline{g}[n] = \mathbb{T}[n] \underline{b}[n] = \begin{bmatrix} \underline{b}_s[n] \\ \mathbb{U}_f[n] \underline{b}_f[n] \end{bmatrix} \tag{36}$$

finally produces

$$\underline{x}[n] = \mathbb{L} \underline{g}[n] + \underline{n}[n]. \tag{37}$$

For this equation in the general linear form, the following hybrid MMSE weight matrix is proposed:

$$\mathbb{W} = \mathbb{L} \left(\mathbb{L}^{H} \mathbb{L} + \sigma^2 \mathbb{R}_{gg}^{-1} \right)^{-1} \in \mathcal{C}^{2NL \times 3(M_s + Q)} \tag{38}$$

which is constant for the duration of the block of B symbols.

Noting that the transmitted symbols and channel coefficients for different users are uncorrelated, the covariance matrix of

$\underline{g}[n]$ needed to formulate the hybrid MMSE weight matrix is calculated from

$$\mathbb{R}_{gg} = \mathcal{E}\left\{\underline{g}[n]\underline{g}[n]^H\right\} = \mathcal{E}\left\{\mathbb{T}[n]\mathbb{T}[n]^H\right\} = \begin{bmatrix} \mathbb{I}_{3M_s} & \mathbb{O}_2^T \\ \mathbb{O}_2 & \mathbb{R}_{\cup_f \cup_f} \end{bmatrix} \tag{39}$$

with the covariance matrix $\mathbb{R}_{\cup_f \cup_f}$ containing the average powers of the path coefficients for the fast moving users.

Application of the hybrid MMSE weight matrix produces the signal

$$\underline{z}[n] = \mathbb{W}^H \underline{x}[n] \in \mathcal{C}^{3(M_s+Q)\times 1}. \tag{40}$$

This directly generates the decision variables for the slow-moving users, but for the fast moving users there are multiple output signals which need to be diversity combined. The vector composed of the decision variables for all users is therefore generated by

$$\underline{d}[n] = \mathbb{T}[n]^H \underline{z}[n] \in \mathcal{C}^{3M\times 1}. \tag{41}$$

The advantage of the hybrid MMSE receiver is that interference from slow users is reduced using the minimum number of degrees of freedom, whereas it is only necessary to evaluate the weight matrix once per block of symbols, even when some users are fast moving.

Channel estimation for the fastest users is performed as per Section IV because subspace tracking is necessary to follow their rapidly time-varying contributions to the signal subspace. For the slow users, composite channel vector accuracy is improved by generating an estimate of the signal subspace from an eigendecomposition of the estimated covariance matrix $\widehat{\mathbb{R}}_{xx}$. However, in the formation of $\widehat{\mathbb{R}}_{xx}$, averaging over the channel fading has occurred. This has the effect that there is no longer a distinct separation of signal and noise space eigenvalues. Fig. 3 shows an example plot of the normalized $\widehat{\mathbb{R}}_{xx}$ eigenvalue magnitudes evaluated over a 2-ms data block for a system with two antennas, two subcarriers, seven chips per symbol, and five users, all moving with the same speed. At 0 km/h, it is clear that there are 15 eigenvalues associated with the signal subspace, and then, a sharp reduction in magnitude denotes the start of the eigenvalues associated with the noise space. When the five users are in motion, there is not such a sharp change in eigenvalue magnitude. Consequently, a basis for a quasisignal subspace is defined by matrix $\widehat{\mathbb{Q}}_s$, which contains the most significant eigenvectors of $\widehat{\mathbb{R}}_{xx}$. In general, $\widehat{\mathbb{Q}}_s$ should contain more than $3M$ vectors so that most of the signal energy is present within the quasisignal subspace. The projector into the quasisignal subspace is then generated prior to channel estimation by alternating projection.

A block diagram of the blind hybrid MMSE receiver is in Fig. 4. It shows the two different channel estimation procedures which are selected between depending on the user speed. A simple procedure to decide whether a user is "fast" or "slow" is to initially assume all users are fast moving and require tracking channel estimation. By monitoring the rate of change of each user's composite channel vector, a decision on the mobility category can then be made.

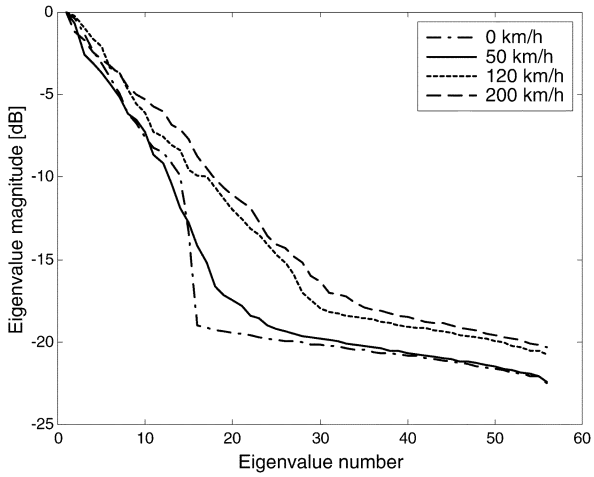


Fig. 3. Magnitudes of the eigenvalues of $\widehat{\mathbb{R}}_{xx}$ at an E_b/N_0 of 12 dB.

VI. Simulation Studies

To evaluate the adaptive channel estimation procedure and the overall performance of the proposed MMSE receivers, computer simulation results were produced. In general, the arrays simulated were of a uniform linear arrangement with half wavelength antenna spacing. Differential quaternary phase shift keying (QPSK) modulation was used throughout, which suffers an approximate performance disadvantage of 2.3 dB at large $\mathrm{SNR}_{\mathrm{in}}$ (input signal-to-noise ratio) when compared to coherent QPSK but means that absolute carrier phase information is not required. The data rate was set at 1 Mbit/s, and the carrier frequency was 2 GHz. Gold codes were used for all PN-sequences.

With regard to the simulation channel model, all paths were incident upon the base station array within a 120° sector. The paths for each user were clustered around a nominal user direction generated from a uniform probability density function (pdf) within this sector. The directions of the paths for a particular user were then generated from a Gaussian pdf with a 10° standard deviation and a mean equal to the nominal user direction. Path delays were set by a uniform pdf in the range of 0 to T_{cs} sec. Scatterers were assumed to be uniformly distributed around the mobile transmitters so that the distribution of Doppler frequencies was provided by Clarke's Doppler spectrum with its characteristic U-shape [20]. Finally, the individual path coefficients were generated with random starting phases drawn from a uniform pdf.

Initially considering composite channel vector estimation, simulations were run for a four-antenna system with eight active users, all transmitting with equal average power. The PN-codes were seven chips long and transmission occurred over two subcarriers. Three times oversampling was used and AWGN added to produce an E_b/N_0 of 7 dB. In the channel model three temporally resolvable paths per user were generated, each path being made up of five temporally unresolvable paths.

The plots in Fig. 5 show the magnitude of the correlation coefficient between the estimated and actual composite channel vector for user-1 over a 1000-symbol time period when all active users have a speed of 50, 120, or 200 km/h. At 2 GHz, these speeds correspond to maximum Doppler frequencies of 93, 222,

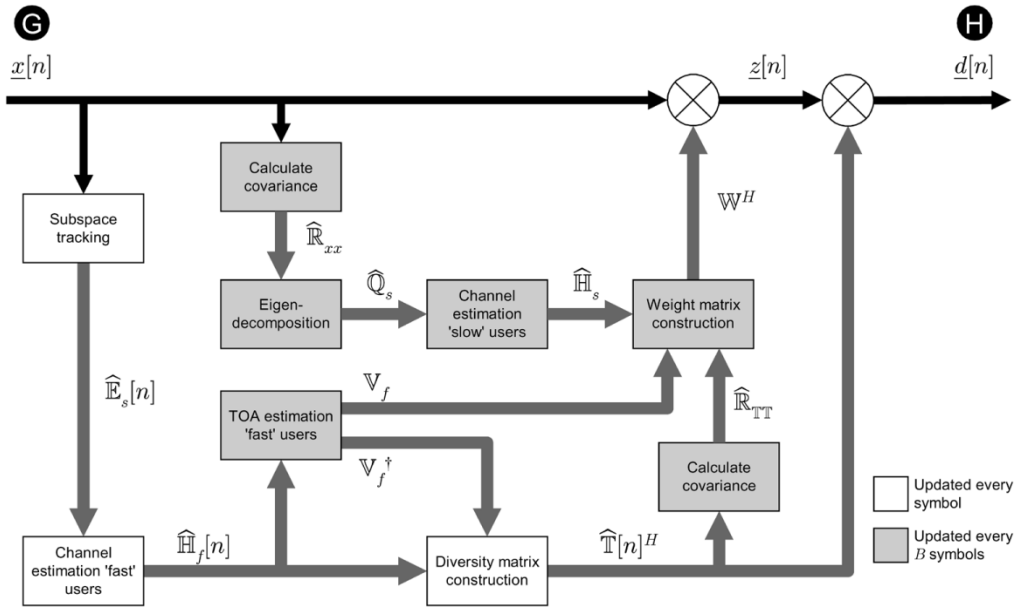


Fig. 4. Structure of the blind hybrid MMSE receiver.

and 370 Hz, respectively.[1] With the symbol duration equal to 2 μs, this does not represent a fast fading situation, but over the block duration of 1000 symbols, the radio channel will experience significant time-variation.

For blockwise estimation in Fig. 5(a), a quasistationary assumption has been made, which is obviously invalid for the 120- and 200-km/h cases, hence the poor results. At 50 km/h, the channel estimate is more accurate with a correlation coefficient above 0.95 for the whole data block. To produce Fig. 5(b), the tracking channel estimator has been applied, and it can be seen that after the initial convergence period,[2] the correlation coefficient magnitude is 0.97 or better at 50 km/h, above 0.93 at 120 km/h, and better than 0.91 at 200 km/h. Note that for these simulations the received signal was generated by applying (13), which includes the Doppler shift within the symbol period. The fact that the tracking composite channel vector estimator operates successfully vindicates the assumption which was used to derive the estimator—namely that the Doppler shift within a single symbol interval is negligible.

For the 120-km/h data, the tracking composite channel vector estimates were also used for TOA estimation, again at an E_b/N_0 of 7 dB. Fig. 6 shows the TOA statistics for user-1 when the three temporally resolvable paths were at $[0.190,\ 0.405,\ 0.738]T_{cs}$. The positions of the three dominant peaks in the plot are coincident with these TOAs, which demonstrates that reliable TOA estimates are produced by building up statistics over a block of symbols.

Receiver performance has been investigated for a system with two antennas at the base station and five active users. Each user's channel was composed of ten paths in total, arriving at two resolvable time delays. The other system parameters remained

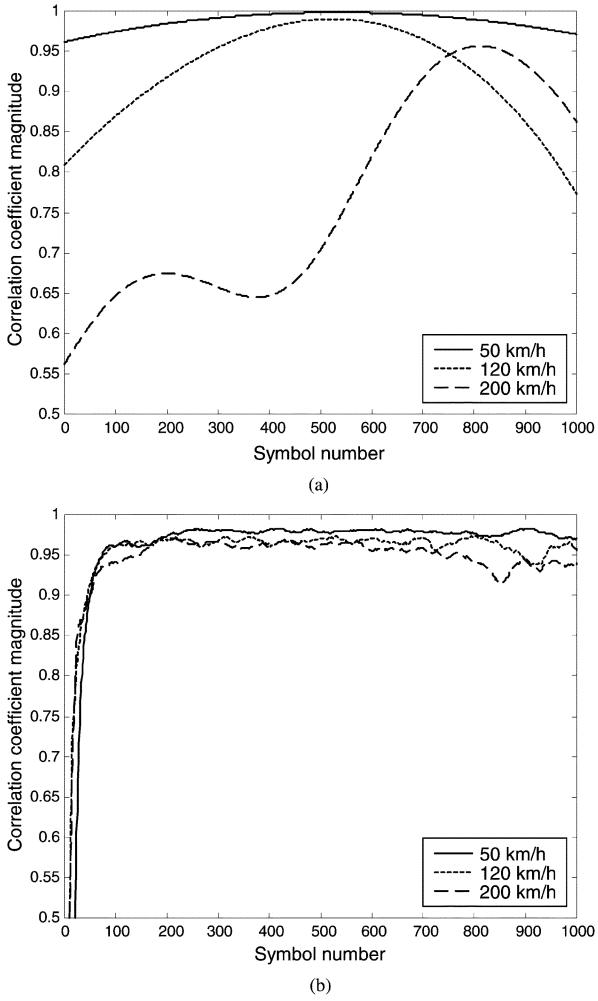


Fig. 5. Composite channel vector accuracy when using a four-element array. (a) Blockwise estimation. (b) Tracking estimation.

the same as for the previous simulations with each data frame containing $B = 1000$ symbols after the transient convergence

[1]It is noted that even greater Doppler frequencies are expected with future standards which operate at higher carrier frequencies.

[2]In the RO-FST subspace tracking algorithm, the forgetting factor was set to 0.99 so that the effective window duration was 100 symbols; hence, steady state performance is achieved after approximately 100 symbol periods.

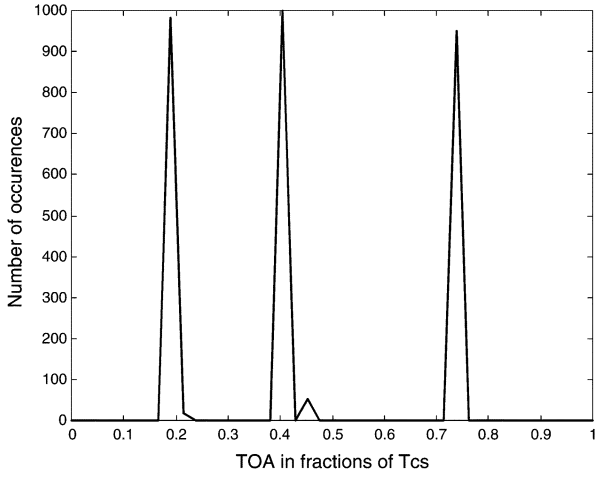


Fig. 6. TOA statistics when there are three temporally resolvable paths and user speed is 120 km/h.

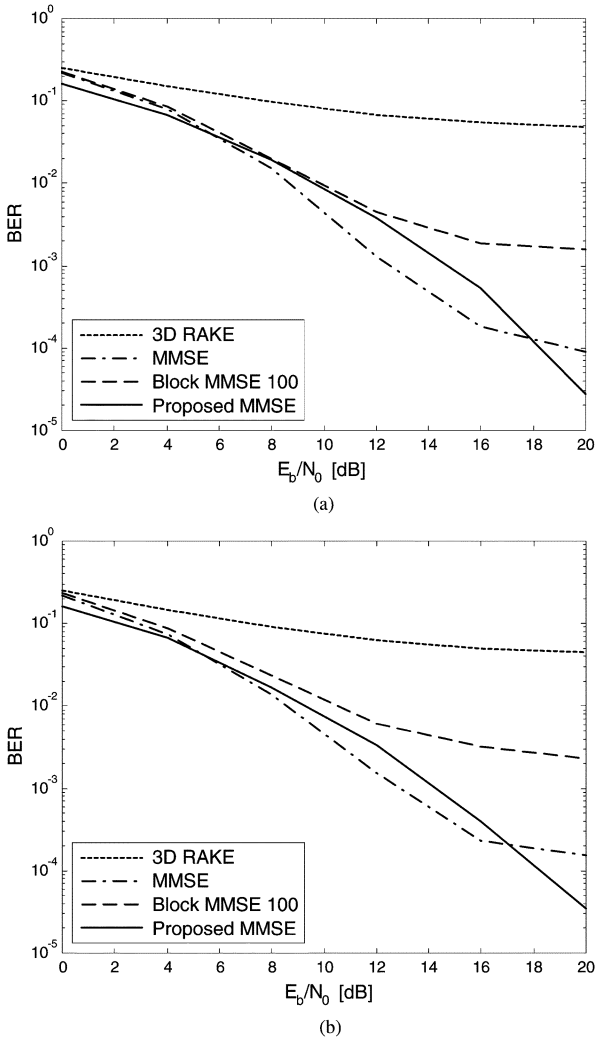


Fig. 7. BER performance for a two-antenna system at (a) 120 km/h. (b) 200 km/h.

period. Fig. 7(a) and (b) shows BER plots for a number of receivers, all blindly implemented, when every user has a velocity of 120 or 200 km/h, respectively. Steady-state BER has been calculated, i.e., after 100 symbols when the subspace tracking

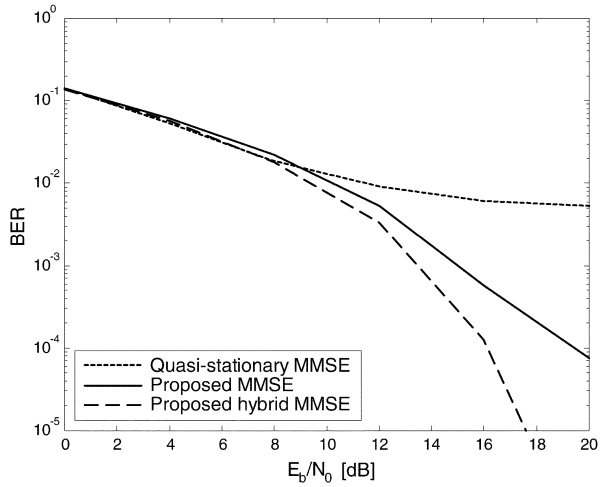


Fig. 8. BER performance for user-1 when there are seven users with different speeds.

algorithm has converged. The different receivers simulated are as follows:

1) 3-D RAKE: Performs maximal ratio combining in space, time, and frequency.
2) MMSE: Defined by (15) and evaluated every symbol period.
3) Block MMSE 100: Defined by (15) and evaluated every 100 symbol periods.
4) Proposed MMSE: Defined by (23) and evaluated once per frame.

Considering the 120-km/h results, it is clear that receiver 1 produces poor performance due to ineffective MAI reduction. Receivers 2 and 3 have higher performance, but they both suffer an error floor at approximately 15 dB E_b/N_0 due to the presence of residual MAI. The error floors appear because of small imperfections in the estimated composite channel vectors, which are directly applied within the receiver [see (15)]. In the case of receiver 3, the performance is further limited due to the implicit quasistationary assumption over the 100-symbol block processing period. The proposed receiver 4 has similar performance to receiver 3 at low and medium SNR levels but does not suffer an error floor in the high SNR region; therefore, it has significantly better performance above 12 dB E_b/N_0. It is to be expected that receiver 2 can outperform the proposed receiver (at least prior to the onset of the error floor) because interference cancellation after diversity combining is known to produce a lower BER than precombining interference cancellation [21]. However, it is stressed that receiver 2 is not recommended for practical implementation due to its excessive complexity compared to the other receivers. In formulating receiver 4, the composite channel vectors are used to estimate more reliable long-term information (e.g., path TOAs), and this channel information is then applied indirectly within (23), with the result that this receiver has no error floor—in contrast to the receivers which directly use the estimated vectors.

At 200 km/h, similar conclusions can be made, but the increased rate of channel variation has the greatest impact on receivers 2 and 3, with comparatively less of a BER reduction for receiver 4. The reason is again because receivers 2 and 3 perform

interference reduction by directly using composite channel vectors which become harder to track as the user speed increases.

A final set of BER results is shown in Fig. 8 to demonstrate the performance of the proposed hybrid MMSE receiver. Seven users were simulated with speeds of $[120,\ 180,\ 100,\ 75,\ 5,\ 15,\ 30]$ km/h, and the threshold between "fast" and "slow" users was set at 50 km/h. The quasistationary MMSE receiver uses the channel estimation procedure of [9] followed by application of (15), the proposed MMSE uses (23), and the proposed hybrid MMSE receiver uses (38). For this mixed-speed scenario, assuming a quasistationary environment for all users significantly reduces performance, and assuming all users channels are rapidly time-varying impairs performance to a lesser extent. The hybrid MMSE receiver therefore succeeds in applying the most appropriate processing for each mobile user.

VII. Conclusion

In this paper array MC-DS-CDMA has been considered for application in mobile communication systems. The mobile radio channel is characterized by multipath propagation and time-variation, which produces frequency-selective fading. In this environment, standard multiuser receivers, which reduce the effect of MAI, become time-varying. However, it is impractical to calculate a new weight matrix every symbol period for complexity reasons. Instead, it has been demonstrated that an alternative approach is possible, which takes into account the different rates of time variation of the channel parameters by separating interference suppression from diversity combining, thus producing a modified MMSE receiver. This modified design only needs to be updated once per block of symbols.

A blind implementation of the modified MMSE receiver has been proposed which applies subspace methods to estimate the composite channel vectors for each user. These estimates are applied to identify the TOAs of the paths arriving from each user; hence, the modified MMSE weight matrix can be calculated. Finally, a diversity-combining matrix is derived and applied to coherently combine the different signal contributions.

Simulation results for time-varying channels indicate the effectiveness of the channel estimation procedure, even in the presence of significant MAI. Furthermore, the overall receiver concept is shown to outperform a block-by-block implementation of the standard MMSE receiver, particularly in the high SNR region.

An alternative blind hybrid MMSE receiver has also been proposed, which is effective when the transmitters have a variety of speeds. This receiver can outperform both the modified MMSE receiver and the quasistationary MMSE receiver by altering the method of interference reduction to be appropriate to the time-varying nature of each user's channel.

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