systems, in this paper, we consider the NGMN to limit its dimensioning process to the core network. A dimensioning framework is proposed that derives a bounded initial estimate of core network elements (the number of BSs/APs per AR, the number of ARs per MAP, and the number of MAPs per GW). The proposed algorithm facilitates NGMN deployment by enabling the providers/operators to select the core network estimate in accordance with their preferred cell loading bounds, future traffic projection, and data rate variability. The cell loading bounds are known a priori, whereas the upper limit on variable data rates assumes pedestrian and vehicular terminal mobility. To accommodate the increased number of users and the bandwidth-intensive applications of the future, fiber optic links with high data transmission rate are considered in the core network.

REFERENCES

MAP-Based Channel Estimation for MIMO–OFDM Over Fast Rayleigh Fading Channels

Jin-Goog Kim and Jong-Tae Lim

Abstract—This paper presents a channel estimation scheme for multiple-input multiple-output with orthogonal frequency-division multiplexing (MIMO-OFDM) in fast varying Rayleigh channels. To handle rapid variation of channels within a transmission block, we propose a novel maximum a posteriori probability-based channel estimation scheme using pilot symbols. With the estimate of the channel matrix for the current symbol interval, a zero-forcing (ZF) receiver is applied to detect the spatially multiplexed data. In simulation results, the effectiveness of the proposed method is shown, as compared with the polynomial and the perfect channel estimates.

Index Terms—Fast varying Rayleigh channels, maximum a posteriori probability (MAP)-based channel estimation, multiple-input multiple-output with orthogonal frequency-division multiplexing (MIMO–OFDM),

I. INTRODUCTION

In recent years, multiple-input–multiple-output (MIMO) antennas combined with orthogonal frequency-division multiplexing (OFDM) have been widely studied in wireless communications because they can provide high data rates and are robust to multipath delay. Channel parameters are needed to coherently decode the transmitted signal and to combine the diversity branches. Channel estimation has been extensively studied for single-antenna systems [3], [4], [7], [8]. For MIMO–OFDM systems, most of the channel estimation schemes have focused on pilot-assisted approaches [3], [8] based on a quasi-static fading model that allows the channel to be constant for a block of OFDM symbols. However, in fast fading channels, the time variation of a fading channel over an OFDM symbol period results in a loss of subchannel orthogonality, which leads to intercarrier interference (ICI) [12]–[14]. To support high-speed mobile channels, the time variation of a fading channel over an OFDM block must be considered. There are more channel parameters in fast fading channels than in quasi-static fading channels. Hence, if we use a channel estimation method that only estimates a few coefficients corresponding to different multipath delays, then we can improve the estimation performance. In [1] and [4], a channel estimation algorithm based on a Qth-order polynomial function has been proposed, but the Qth-order polynomial approximation does not reflect the channel characteristic.

In this paper, we derive a novel pilot-symbol-aided maximum a posteriori probability (MAP)-based channel estimation scheme for a MIMO–OFDM system over fast Rayleigh fading channels. Since the proposed method uses the covariance matrix of the channel, the channel characteristic is considered by the estimation scheme. Then, a zero-forcing (ZF) receiver is applied to detect the spatially multiplexed data transmitted in the current symbol interval.
II. SYSTEM MODEL

Consider a MIMO–OFDM system with \( N_T \) transmit antennas, \( N_R \) receive antennas, and \( N \) subcarriers, which employs quadrature amplitude modulation (QAM). The transmitted symbols are denoted by \( \mathbf{X} = [ \mathbf{x}(0)^T, \mathbf{x}(1)^T, \ldots, \mathbf{x}(N-1)^T]^T \) with \( \mathbf{x}(k) = [x_1(k), x_2(k), \ldots, x_N(k)]^T \), where \( x_i(k) \) is the transmitted signal by the \( i \)th transmit antenna on subcarrier \( k \); and the received symbols are denoted by \( \mathbf{Y} = [ \mathbf{y}(0)^T, \mathbf{y}(1)^T, \ldots, \mathbf{y}(N-1)^T]^T \) with \( \mathbf{y}(k) = [y_1(k), y_2(k), \ldots, y_N(k)]^T \), where \( y_i(k) \) is the received signal by the \( j \)th receive antenna on subcarrier \( k \). The transmitted symbols are fed to an inverse discrete Fourier transform (IDFT) to produce the OFDM signal, and a guard interval is inserted, which is the case of slowly fading channels. Consequently, \( \mathbf{G} = \mathbf{F}_{N_T} \mathbf{H} \mathbf{F}_{N_R}^H \) is no longer a block diagonal matrix. This shows that time-selective fading causes ICI, which is represented by the off-diagonal blocks of \( \mathbf{G} \). Note that (1) can be rewritten as

\[
\mathbf{Y}(p) = \mathbf{G}(p,p)\mathbf{X}(p) + \sum_{q=0}^{N-1} \sum_{q \neq p} \mathbf{G}(p,q)\mathbf{X}(q) + \mathbf{Z}(p). \tag{6}
\]

In a time-selective channel, the first term in the right-hand side of (6) is the desired term without ICI in the frequency domain, which denotes the contribution from the same symbol, and the second term in the right-hand side of (6) is the ICI term in the frequency domain, which denotes the contribution from other symbols.

The fading channel process \( h(t) \) is modeled as a normalized zero-mean complex wide-sense stationary Gaussian process with a correlation function \( r_h(t) = E \{ h(t) h^*(t+\triangle t) \} \) [6], where \( E \{ \cdot \} \) denotes the statistical mean, and \( (\cdot)^* \) represents the complex conjugate of \( (\cdot) \). As shown in [5] and [15], the space–time correlation function of the channel in a typical mobile communication environment can be modeled as

\[
r_h(\triangle t) = J_0(2\pi f_d \triangle t) \tag{7}
\]

where \( f_d \) is the maximum Doppler shift, and \( J_0(\cdot) \) is the zeroth-order Bessel function of the first kind.

III. CHANNEL ESTIMATION

In this section, we consider the problem of channel estimation in a MIMO–OFDM system over time-selective channels. To estimate parameters, we change the equation \( \mathbf{Y} = \mathbf{G} \mathbf{X} + \mathbf{Z} \) to the form of

\[
\mathbf{Y} = \mathbf{W} \mathbf{h} + \mathbf{Z},
\]

where \( \mathbf{h} = [h_T(0), h_T(1), \ldots, h_T(N-1)]^T \) and \( \mathbf{h}(n) = [\mathbf{h}(n,0)^T, \ldots, \mathbf{h}(n, L-1)^T]^T \), where \( \mathbf{h} \) is the vector obtained by stacking the columns of \( \mathbf{x} \) on top of each other. Here, \( \mathbf{h}(n, l) \) is an \( N_R \times N_T \) matrix, and the entry of \( \mathbf{h}(n, l) \) is defined by

\[
h_{n,l} = \begin{bmatrix} h_{1,1}(n,l) & \cdots & h_{1,N_R}(n,l) \\ \vdots & \ddots & \vdots \\ h_{N_R,1}(n,l) & \cdots & h_{N_R,N_T}(n,l) \end{bmatrix}, \quad 0 \leq n \leq N-1, \quad 0 \leq l \leq L-1. \tag{8}
\]

The \( p \)th received symbol at the \( r \)th receive antenna, i.e., \( y_{r,p}(p) \), is expressed as follows:

\[
y_{r,p}(p) = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} \sum_{q=0}^{N_T-1} h_{r,l}(n,l) x_{l,q}(p) D(p,q,n,l) + z_{r,p}(p)
\]

where \( W(p,q,n,l) = \frac{1}{N} \sum_{q=0}^{N_R-1} x_{r,q}(p) D(p,q,n,l) \), and \( D(p,q,n,l) = e^{-j(2\pi/N)np} e^{-j(2\pi/q)/N} \).

Letting a set of \( N_P \) pilot tones be \( \mathcal{T} = \{ P(1), \ldots, P(N_P) \} \), \( W(p,n,l,t) \) can be split into two matrices as follows:

\[
W(p,n,l,t) = \frac{1}{N} \sum_{q \in \mathcal{P}} x_{r,q}(p) D(p,q,n,l) + \frac{1}{N} \sum_{q \notin \mathcal{P}} x_{r,q}(p) D(p,q,n,l)
= W_{P}^{p}(p,n,l,t) + W_{N}^{p}(p,n,l,t). \tag{9}
\]
Thus, we obtain
\[ Y(p) = W_p^D h + W_p^D h + Z(p) \]  
(10)

with \( W_p^D = \left[ W_p^D(p,0,0,1) \otimes I_{N_R} \right], ..., \left[ W_p^D(p,N-1,N-1,N_p) \otimes I_{N_R} \right] \)
and \( W_p^D = \left[ W_p^D_p(P,0,0,1) \otimes I_{N_R} \right], ..., \left[ W_p^D_p(P,N-1,N-1,N_p) \otimes I_{N_R} \right] \).
Since we do not know \( P \), \( W_p^D \) has unknown elements.

A. Channel Estimation Scheme

When we consider a system where ICI is not negligible, \( W_p^D \) cannot be simply removed. Hence, we should account for the ICI in the channel estimation process, as shown in [16] and [17]. For the pilot-based channel estimation scheme over fast fading channels, the unknown \( W_p^D \) contributes to the approximation error. By defining \( e(p) := W_p^D h + Z(p) \), (10) can be rewritten as
\[ Y(p) = W_p^D h + e(p) \]

where \( e(p) \) is composed of the ICI contribution from nonpilot tones and the noise contribution. Let \( \hat{Y} = [Y(P(1)), ..., Y(P(N_P))]^T \) and \( \hat{W} = [W_p^D(P(1))^T, ..., W_p^D(P(N_P))^T]^T \). Then, we can form the \( N_P N_R \times N_L N_T \) system of linear equations as
\[ \hat{Y} = \hat{W} h + \hat{e} \]
(11)

where \( \hat{Y} \), \( \hat{e} \), and \( \hat{W} \) are defined as shown at the bottom of the page.

However, \( W \) is an \( N_P N_R \times N_L N_T \) matrix, and \( N_P \leq N_L T \) for each receive antenna. To obtain \( h \) as the least square solution, the number of rows of \( \hat{W} \) will be larger than the number of columns of \( \hat{W} \). Thus, we need to reduce the number of channel parameters. To reduce the number of parameters needed for channel estimation, we use the sliding-window approach in which \( h_v, t \) is derived from the previous channel parameters \( h_v, (n - 1) \), \( m = n - M, ..., n - 1 \), where \( M \) is the window size, and \( M \leq N_P / L_N T \). That is, we construct \( h_v, t \) as follows:
\[ h_v, (n - 1) = a_1 h_v, (n - M - 1) + \cdots + a_M h_v, (n - 1, l) \]
\[ = a^T [h_v, (n - M - 1, l), ..., h_v, (n - 1, l)]^T . \]
(12)

Note that we have used the same \( M \times 1 \) tap weight vector \( a \) for all of the \( L \) channel taps of \( h_v, (n) \).

Let us define an \( M \times 1 \) vector \( u_v, t \) and \( (M + 1) \times 1 \) vector \( v_v, t \) as
\[ u_v, t = [h_v, (n - M, l), ..., h_v, (n - 1, l)]^T \]
(13)
\[ v_v, t = [u_v, t, h_v, (n, l)]^T . \]
(14)

The covariance matrix of the zero-mean vector \( v_v, t \) can be written as
\[ R_h = E \{ v_v, t (n, l)v_v, t^H (n, l) \} = \begin{bmatrix} R_{h11} & r_{h12} \\ r_{h12}^* & r_{h22} \end{bmatrix} \]
(15)

where
\[ R_{h11} = E \{ u_v, t (n, l)u_v, t^H (n, l) \} \]
(16)
\[ r_{h12} = E \{ h_v, t (n, l)h_v, t^H (n, l) \} = r_{h12} \]
(17)
\[ r_{h22} = E \{ h_v, t (n, l)h_v, t^H (n, l) \} . \]
(19)

If we know \( u_v, t \), the estimate of \( h_v, t \) can be obtained by maximizing its conditional probability density function \( p[h_v, t | u_v, t] \) as
\[ \hat{h}_v, t := \max_{h_v, t | u_v, t} p[h_v, t | u_v, t] . \]
(20)

For the Rayleigh channels being considered, the \( (M + 1) \times 1 \) vector \( v_v, t \) is a complex Gaussian. As shown in [10], we write the conditional probability density function as
\[ p[h_v, t | u_v, t] = \frac{1}{\sqrt{|A|}} \exp \left\{ -v_v, t^H R_h^{-1} v_v, t \right\} \]
(21)

where \( A \) denotes the determinant of matrix \( A \). By using the matrix inversion lemma, \( R_h^{-1} \) is expressed as
\[ R_h^{-1} = \begin{bmatrix} R_{h11} & r_{h12} \\ r_{h12}^* & r_{h22} \end{bmatrix}^{-1} \]
\[ = \begin{bmatrix} R_{h11} - r_{h12}r_{h21}^* & -r_{h11}r_{h12} \big) \] 
\[ \big( r_{h22} - r_{h21}r_{h12}^{-1} \big) \] 
(22)

\[ = \begin{bmatrix} R_{h11} - r_{h12}r_{h21}^* & -r_{h11}r_{h12} \big) \] 
\[ \big( r_{h22} - r_{h21}r_{h12}^{-1} \big) \] 
(22)

where \( R_{h11} = \big( R_{h11} - r_{h12}r_{h21}^* \big) \), and \( r_{h22} = r_{h22} - r_{h21}r_{h12}^{-1} \).

Maximizing the conditional probability density function is equivalent to minimizing the following objective function [9]:
\[ f = v_v, t^H R_h^{-1} v_v, t - u_v, t^H R_h^{-1} u_v, t . \]
(23)

By letting the conjugate derivative of \( f \) with respect to \( h_v, t \) be equal to zero, we obtain the estimate of \( h_v, t \) as
\[ \hat{h}_v, t = a^T u_v, t . \]
(24)
where \( a^T = r_{121}, R_{121}^{-1} \) is the \( M \times 1 \) tap weight vector, and \( r_{121} \) and \( R_{121} \) are obtained using (7). Since all elements of \( H \) are identically distributed, tap weight vector \( a \) is common for all coefficients [14]. Then, we can reduce the \( NLN_T N_R \) channel parameters to the \( MLN_T N_R \) channel parameters by using the tap weight vector.

Consider a linear system of \( N \) equations using (12) as follows:

\[
\begin{align*}
    h_{r,t}(0,l) &= a_1 h_{r,t}(s_1,l) + \cdots + a_M h_{r,t}(s_M,l) \\
    h_{r,t}(1,l) &= a_1 h_{r,t}(s_2,l) + \cdots + a_M h_{r,t}(s_M,l) \\
    & \vdots \\
    h_{r,t}(N-2,l) &= a_1 h_{r,t}(N-2-M,l) + \cdots + a_M h_{r,t}(N-3,l) \\
    h_{r,t}(N-1,l) &= a_1 h_{r,t}(N-1-M,l) + \cdots + a_M h_{r,t}(N-2,l).
\end{align*}
\]

We reconstruct the linear system as follows:

\[
\begin{bmatrix}
    h_{r,t}(0,l) \\
    h_{r,t}(1,l) \\
    \vdots \\
    h_{r,t}(N-1,l)
\end{bmatrix} = Q
\begin{bmatrix}
    h_{r,t}(s_1,l) \\
    h_{r,t}(s_2,l) \\
    \vdots \\
    h_{r,t}(s_M,l)
\end{bmatrix}
\]

where

\[
Q = \begin{bmatrix}
1 & 0 & \cdots & 0 \\
-a_M & 1 & \cdots & 0 \\
-a_M & -a_M & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & \cdots & -a_M & 1
\end{bmatrix}^{-1}
\begin{bmatrix}
a_1 & a_2 & \cdots & a_M \\
0 & a_1 & \cdots & a_{M-1} \\
\vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & a_1 \\
0 & \cdots & 0 & 0
\end{bmatrix}
\]

Note that matrix \( A \) is defined as \( A = Q \otimes I_d \otimes I_{N_T} \otimes I_{N_R} \), and vector \( h \) is defined as \( h = \text{vec}(h(0), \ldots, h(L-1))^T \), where \( h(l) = [h_1(l), \ldots, h_{N_T}(l), h_{N_T}(l)] \). Using our proposed method, we conclude that \( h = \hat{A}h \), and we need to estimate \( MLN_T N_R \) parameters. Then, we rewrite the system model in (11) as follows:

\[
\hat{Y} = \tilde{W}h + \hat{e} = \tilde{W}A\hat{h} + \hat{e}
\]

where \( \tilde{W}A \) has a full column rank. The estimated channel vector is \( \hat{h} = \hat{A}h = A(\tilde{W}A)^{+}\hat{Y} \), where \( (.)^{+} \) denotes the pseudoinverse of \( (.) \).

### B. Pilot Tone Selection and Detection

Pilot tone placement is very important for the performance of channel estimation. For a time-invariant frequency-selective channel, pilot tones should minimize the effects of frequency selectivity and be equispaced on the fast Fourier transform (FFT) grid. However, for a time-varying frequency-selective channel, both frequency and time selectivity should be considered. For a time-selective channel, pilot tones should be placed as close as possible. From the aforementioned results, in a time-invariant frequency-selective channel and a time-varying channel, we choose the grouped and equispaced pilot tones on the FFT grid.

The estimation error by the least square is

\[
E_e = \| I - (\tilde{W}A)(\tilde{W}A)^{+} \|^2 \| \hat{Y} \|^2.
\]

Given the channel parameters, the transmitted symbols can be detected using several algorithms, such as the maximum-likelihood detection, the minimum mean square error detection, the ZF detection, and the Bell Laboratories layered space–time architecture scheme. For the lowest computational complexity, the ZF scheme will be adopted in this paper for symbol detection.

In the ZF scheme [11], the transmitted symbol vector is written as

\[
\hat{X} = \hat{G}^{\dagger}Y.
\]

We will further discuss the pilot placement problem in the next section.

### IV. Simulation

In the simulation, we consider a system with a \( 2 \times 2 \) antenna structure. To construct an OFDM signal, assume that the entire channel bandwidth, i.e., 800 kHz, is divided into 128 tones. The symbol duration is 160 \( \mu \)s, and an additional 40-\( \mu \)s guard interval is used to avoid ISI due to channel multipath delay spread. Thus, the total block length \( T = 200 \) kHz, and the subchannel symbol rate \( r_s = 5 \) kbd. For its simplicity with coherent demodulation, quadrature phase-shift keying modulation is chosen to demonstrate the performance of channel estimation scheme, and bit error rates (BERs) are computed over the symbol duration. To gain the average behavior of the channel estimator, we have averaged the performance over 1000 OFDM blocks. The pilot symbols are inserted in the system with \( N_P = 16 \). Fading channels among all transmit and receive antennas are assumed to be independent. Then, we consider channels with \( f_d = 100 \) Hz \((f_dT = 2 \times 10^{-2})\) and \( f_d = 200 \) Hz \((f_dT = 4 \times 10^{-2})\), where \( f_d \) is the Doppler frequency, and the channels corresponding to different receivers have the same statistics. In Figs. 1–5, \( Q \) denotes the polynomial order, and \( M \) denotes the window size.

Figs. 1 and 2 show the BER performance of channel estimation schemes using pilot set 4 of Table I. For comparison purposes, the error rate curve with the perfect channel estimate is also shown. In Figs. 1 and 2, we show that the proposed method with \( M = 1, 2 \) has better BER performance than the polynomial approximation with \( Q = 0, 1 \) at low \( E_b/N_0 \), respectively. In both schemes, the error performance significantly improves when the window size \( M = 3 \) and the polynomial order \( Q = 2 \); whereas the computational complexity increases as \( O(n^3) \), where \( n = (Q + 1)LN_T N_R \) for the polynomial approximation and \( n = MLN_T N_R \) for the proposed method. The computational complexity is critical for the high transmission system, and the high computational burden is undesirable, as shown in [1].
TABLE I
PILOT SETS

<table>
<thead>
<tr>
<th>Pilot Set</th>
<th>Pilot Place</th>
</tr>
</thead>
<tbody>
<tr>
<td>Set 1</td>
<td>(26,31,36,41,46,51,56,61,66,71,76,81,86,91,96,101)</td>
</tr>
<tr>
<td>Set 2</td>
<td>(56,57,58,59,60,61,62,63,64,65,66,67,68,69,70,71)</td>
</tr>
<tr>
<td>Set 3</td>
<td>(14,15),(30,31),(46,47),(62,63),(78,79),(94,95),(110,111),(126,127)</td>
</tr>
<tr>
<td>Set 4</td>
<td>(28,29,30,31),(60,61,62,63),(92,93,94,95),(124,125,126,127)</td>
</tr>
</tbody>
</table>

[4], [16], and [17]. Hence, $Q = 2$ and $M = 3$ might not be practical choices due to the intensive computational burden. Then, the proposed method shows better performance than the polynomial approximation at low $E_b/N_0$ under the constraint of low computational complexity. In Fig. 3, we show the BER performance of estimation schemes with various Doppler frequencies using pilot set 4 of Table I. Since it uses the result of the least square estimation, it is sensitive to the Doppler frequency. However, for all Doppler frequency ranges, the proposed scheme has the lower error rate.

In Figs. 4 and 5, we show the performance for the pilot placement schemes. We consider four pilot sets. The first pilot set is the equispaced pilot tones, and the second pilot set is the grouped pilot tones. The third and fourth pilot sets are the grouped and equispaced pilot tones. For a time-invariant frequency-selective channel, the pilot tones should be equispaced on the FFT grid. However, for a time-varying channel, the pilot tones should be placed as close as possible. Hence, it seems that the grouped and equispaced pilot tones should be chosen. These pilot sets are described in Table I. As expected, the grouped and equispaced pilot sets result in good performance. Between pilot set 3 and pilot set 4, pilot set 4 shows better performance, which results from the ability of estimating the time selectivity of the channel.

V. CONCLUSION

In this paper, we have proposed MAP-based channel estimation using pilot symbols over fast Rayleigh fading channels in a MIMO–OFDM system. We used the sliding-window approach to reduce the number of channel parameters needed for channel estimation within an OFDM block. Considering the computational burden, the proposed estimation scheme showed better performance than the polynomial approximation. Through simulations, we showed that our proposed scheme has a lower error rate at low $E_b/N_0$ than the scheme based on the polynomial approximation, under the constraint of low computational complexity.
On the Transmitter-Based Preprocessing for 2-D OFDM-CDMA Forward-Link Systems Over Time-Varying Rayleigh Fading Channels

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Abstract—Transmitter-based preprocessing is investigated for 2-D orthogonal frequency-division multiplexing code-division multiple-access (OFDM-CDMA) forward-link systems for improving performance and shifting signal processing complexity from a mobile unit to a base station. Preprocessing schemes that are based on zero forcing (ZF) with power normalization, minimum mean square error (MMSE), and ZF with multi-user water filling (ZF-MWF) criteria are jointly investigated with 2-D spreading pattern optimization and multi-user scheduling from an information-theoretic viewpoint. Numerical results show that 1) the performance of preprocessing is quite sensitive to the 2-D spreading pattern for SNRs of interest, for example, 20% degradation on the sum data rate is observed for MMSE preprocessing if the spreading pattern is not properly selected; 2) ZF-MWF may substantially outperform the other two criteria depending on the SNRs; and 3) multi-user scheduling provides a significant performance improvement on the system sum data rate.

Index Terms—Forward-link systems, sum data rate, transmitter preprocessing, two-dimensional (2-D) orthogonal frequency-division multiplexing code-division multiple access (OFDM-CDMA).

I. INTRODUCTION

Orthogonal frequency-division multiplexing code-division multiple access (OFDM-CDMA) is a promising radio access technology for the next-generation mobile communication systems thanks to its ability to overcome intersymbol interference that is incurred in high-data-rate transmission, its ability to provide universal frequency reuse in a multicell environment, and its ability to achieve high-order diversity gain by spreading data over frequency and time domains [1]–[5]. Traditionally, OFDM and CDMA are combined in a 1-D fashion, that is, a data symbol is spread either in frequency or time domain (see [1] and references therein). Recently, 2-D OFDM-CDMA, where data are spread over time and frequency domains, has been proposed to improve the performance of the 1-D one by simultaneously exploiting the temporal and spectral characteristics of the fading channels [2]–[5].

Transmitter-based preprocessing, on the other hand, has been proposed for improving performance and shifting signal processing complexity from a mobile unit to a base station in mobile communication systems [6]–[13]. Zero forcing (ZF) [6], [7] and pre- or [8], [9] preprocessing methods with a different degree of receiver complexity were proposed for direct-sequence CDMA (DS-CDMA) systems. In addition, transmitter preprocessing schemes that are based on the MMSE [11], [13] or ZF [10]–[12] were investigated for broadcast multiple-input–multiple-output (MIMO) systems.

This paper aims to design 2-D OFDM-CDMA forward-link systems with transmitter preprocessing to improve system performance. Three preprocessing methods, including ZF with power normalization (ZF-PN), the MMSE, and ZF with multi-user water filling (ZF-MWF),...