Multi-Stream Coding for MIMO OFDM Systems with Space-Time-Frequency Spreading

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Abstract

We propose a coding strategy for MIMO OFDM systems that exploits full spatial (transmit and receive) diversity, as well as frequency and time diversity, through the combination of binary codes and linear precoding. The proposed scheme, based on streams (layers/threads), was designed to make iterative decoding with interference-cancellation relatively simple, while preserving most of the diversity available from the channel. It accommodates any number of transmit and receive antennas. Nevertheless, it is a full-rate scheme, since no redundancy is added by the linear precoding operations.

Keywords

Spatial multiplexing, MIMO, OFDM, frequency selective channel, linear precoding, diversity, space-time-frequency coding/spreading, coding gain, interence cancellation, iterative/turbo receiver.

INTRODUCTION

Multiple-antenna transmission/reception has been shown to be a way to deliver very high data rates over wireless channels [1]. Space-Time codes for Multiple-Input-Multiple-Output (MIMO) channels [2] have been proposed in order to mitigate the effects of channel fades. While analyses of the diversity achieved by such schemes are readily available, their decoding requires maximum-likelihood (ML) [3]. This is not an issue for systems with a small number of antennas (for example, the Alamouti scheme [4] is an elegant solution to this problem for the special case of 2 transmit antennas). Unfortunately, ML decoding is not practical for systems with a large number of TX antennas and/or large symbol alphabets. The use of multiple antennas on wideband channels provides an extra degree of diversity, namely frequency, which can be exploited through the use of Orthogonal Frequency Division Multiplexing (OFDM). Although the number of channel coefficients involved in the propagation of an OFDM symbol (i.e. the width of the Fast Fourier Transform used) is usually rather large, they are in general correlated, since they are all derived from a (much smaller) number of time-domain coefficients of the impulse response. Hence, the frequency diversity actually available in the channel is limited. We seek to exploit it through linear precoding.

We present here a stream-based coding scheme, along the lines of the threaded scheme proposed by El Gamal in [5]. In our approach, each stream is separately encoded by a binary code, mapped onto a complex constellation, and then spread over the OFDM tones and TX antennas so that it is not self-interfering. Iterative decoding with stream-based interference cancelation is proposed.

NOTATIONS

M and N denote respectively the number of transmit (Tx) and receive (Rx) antennas. P is the number of tones in the OFDM transmission. $\mathcal{I}\left(\,.\,;\,.\,\right)$ and $\mathcal{H}\left(\,.\,\right)$ are the mutual information and entropy functions, respectively. The Kronecker product is denoted by $\otimes$. The Hermitian transpose operator is denoted by ^{H}. If $\mathbf{v}$ is a vector, $\mathrm{diag}(\mathbf{v})$ denotes the diagonal or block-diagonal matrix with the elements of $\mathbf{v}$ on its diagonal. For a matrix $\mathbf{A}$, $\mathrm{diag}(\mathbf{A})$ is the matrix corresponding to the diagonal part of $\mathbf{A}$. $\mathcal{R}_{\mathbf{uv}}$ is the covariance matrix between vectors $\mathbf{u}$ and $\mathbf{v}$.

CODING SCHEME

Our approach relies upon parallel, separately coded, data streams. As stated in the introduction, an important characteristic is that streams should not be self-interfering. This eases the requirements on the decoder, since this independence enables iterative streams-based interference cancelation, while being assured that the estimated interference remains uncorrelated with the stream being decoded. This is achieved by a frequency (or tone) allocation scheme that maps different streams to all Tx antennas over each frequency subband. For this reason, the number of streams is chosen to be equal to the number of Tx antennas.

The available (space-, time-, and frequency-) diversity is exploited through spreading : constellation symbols are grouped into vectors that are linearly coded (spread) with a square matrix $\mathbf{Q}$, and interspersed over the available antennas and frequencies.

Channel model

Over each tone $p \in \{1,\ldots,P\}$, the channel at OFDM symbol period j is represented by a $N \times M$ matrix of complex

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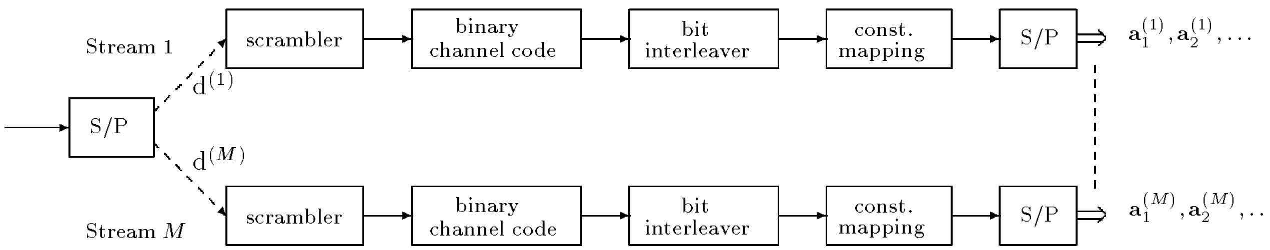


Figure 1. Initial stream separation

coefficients

$$\mathrm{H}_{j,p} = \begin{pmatrix} h_{j,1,1,p} & \dots & h_{j,1,M,p} \\ \vdots & \ddots & \vdots \\ h_{j,N,1,p} & \dots & h_{j,N,M,p} \end{pmatrix}$$

Since signals at different tones do not interfere with each other, these can be gathered in the block-diagonal matrices

$$\mathbf{H}_j = \begin{pmatrix} \mathrm{H}_{j,1} & & 0 \\ & \ddots & \\ 0 & & \mathrm{H}_{j,P} \end{pmatrix}, \ \underline{\mathbf{H}} = \begin{pmatrix} \mathbf{H}_0 & & 0 \\ & \ddots & \\ 0 & & \mathbf{H}_{J-1} \end{pmatrix}$$

$\underline{\mathbf{H}}$ representing J successive channel realizations. We assume that the channel coefficients are spatially and temporally uncorrelated (temporal decorrelation can always be achieved by interleaving with a sufficient delay, whereas spatial decorrelation is a function of antenna spacing). Channel coefficients can be correlated between frequencies for a single Tx-Rx antenna pair. These assumptions can be summarized as

$$\forall (p_1, p_2), \quad (t_1, m_1, n_1) \neq (t_2, m_2, n_2) : \\ \mathrm{E}\left[h_{t_1,n_1,m_1,p_1} h_{t_2,n_2,m_2,p_2}{}^*\right] = 0 . \quad (1)$$

We also assume that the channel frequency diversity is at least L. In other words, frequencies taken P/L tones (coherence bandwidth [6]) apart or more are assumed totally uncorrelated.

$$\forall (t, m, n), \quad |p_1 - p_2| \geq \frac{P}{L} : \\ \mathrm{E}\left[h_{t,n,m,p_1} h_{t,n,m,p_2}{}^*\right] = 0 . \quad (2)$$

Taking fewer tones leads to incomplete exploitation of frequency diversity, taking more tones leads to strongly correlated channel coefficients (and possibly no covariance matrix rank increase). For a given number L of tones used, they should be spaced out as widely and as evenly as possible to minimize mutual correlation. The noise samples are assumed to be white complex Gaussian, independent and identically distributed over all frequencies.

Stream Principle

The incoming data is split into M streams (equal to the number of TX antennas) that are treated separately. Each stream is encoded using a binary (convolutional or block) channel code, mapped onto complex symbols, and serial-to-parallel (S/P) converted to form $ML \times 1$ vectors $\mathbf{a}_t^{(k)}$, see Fig. 1. Inside each stream k, those are linearly precoded through multiplication by a square matrix $\mathbf{Q}$:

$$\mathbf{x}_t^{(k)} = \mathbf{Q}\mathbf{a}_t^{(k)} \quad (3)$$

Since $\mathbf{Q}$ is square, no redundancy is added at the linear precoding stage : this is a full-rate precoding.

Tone/TX Antenna Assignment

Every constellation symbol is spread by matrix $\mathbf{Q}$ over a set of ML values, that must be assigned onto particular tones and Tx antennas to form OFDM symbols. They are interspersed in time, space and frequency to ensure that the fading coefficients are as little correlated as possible. To this end, we used the following criteria :

- all antennas must be used evenly
- all frequencies must be used evenly

Due to the correlation between adjacent frequencies over the same antenna pair, the spectrum need only be sampled L times on each Tx antenna by every symbol. On the other hand, due to total spatial decorrelation, it makes sense for every symbol to use all M Tx antennas.

Furthermore, in order to facilitate the iterative process of co-antenna interference cancelation, we want to ensure that all the interference comes only from other streams, and thus is uncorrelated to the stream being considered. This adds the following constraint :

- no two symbols from the same stream can be transmitted over the same tone in an OFDM symbol period

This leads to the following design : let us consider J consecutive OFDM symbols, and denote by

$$\mathrm{S}_{j,m} = \begin{pmatrix} s_{j,1,m} \\ \vdots \\ s_{j,P,m} \end{pmatrix}$$

the frequency-domain representation of the OFDM symbol transmitted over antenna m at time j. We assume that P is a mutiple of ML : $P = GML$. Over this period of time, the output of the coders for all M streams is

$$\mathbf{x}_{jG+g}^{(k)}, \quad \begin{array}{l} k \in \{1, \dots, M\}, \\ g \in \{0, \dots, G-1\}, \\ j \in \{0, \dots, J-1\} \end{array}$$

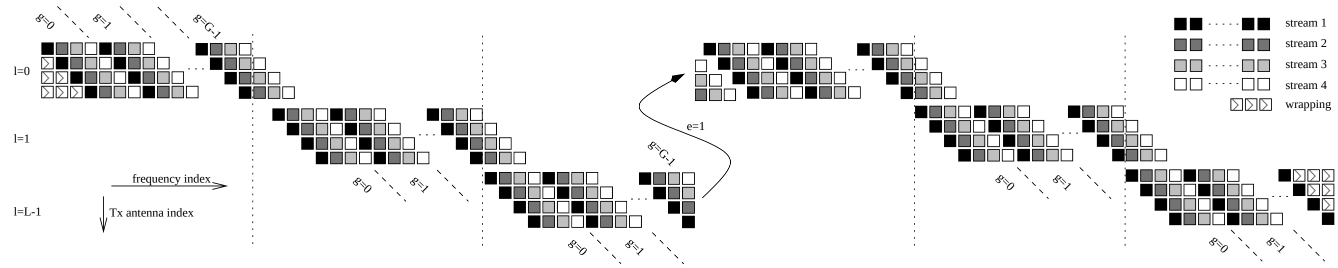


Figure 2. Stream tone/TX antenna assignment for $M = 4$, $J = 2$, $L = 3$.

Since each of those contains ML values, a total of $M^2JGL = JMP$ complex values are output. These values are mapped onto the OFDM symbols according to

$$\forall l \in \{0, \dots, L-1\}, i \in \{1, \dots, M\}, f \in \{1, \dots, P\},$$
$$i + k - 1 + lGM + gM = eP + f$$
$$\Rightarrow s_{j+e,f,i} = x^{(k)}_{jG+g,lM+i} \quad (4)$$

whith $e = 1$ iff $i + k - 1 + lGM + gM \geq P$ and $j \neq J - 1$ (when $\mathbf{x}^{(k)}_{jG+g}$ is split over two consecutive OFDM symbols), and $e = -(J-1)$ iff $i + k - 1 + lGM + gM \geq P$ and $j = J - 1$ (wrapping). An example of this tone allocation is shown in Fig. 2. Due to the splitting of tones in L groups, actually ML rows appear in the figure, to emphasize the spreading of symbols over ML (antenna,tone) pairs. The overall (antenna,tone) assignment operation is a $JMP \times JMP$ permutation, and can be described by a permutation matrix $\mathbf{W}$. Let us gather all the $\mathbf{a}^{(k)}_t$ transmitted over J OFDM symbol periods, and the corresponding received signals

$$\mathbf{A^{(k)}} = \begin{pmatrix} \mathbf{a}^{(k)}_1 \\ \vdots \\ \mathbf{a}^{(k)}_{JG} \end{pmatrix}, \quad \mathbf{A} = \begin{pmatrix} \mathbf{A^{(1)}} \\ \vdots \\ \mathbf{A^{(M)}} \end{pmatrix}$$

$$\mathrm{Y}_{j,p} = \begin{pmatrix} y_{j,p,1} \\ \vdots \\ y_{j,p,N} \end{pmatrix}, \mathbf{Y}_j = \begin{pmatrix} \mathrm{Y}_{j,1} \\ \vdots \\ \mathrm{Y}_{j,P} \end{pmatrix}, \mathbf{Y} = \begin{pmatrix} \mathbf{Y_0} \\ \vdots \\ \mathbf{Y_{J-1}} \end{pmatrix}$$

Using these notations, and with

$$\underline{\mathbf{Q}} = \mathbf{I}_{MGJ} \otimes \mathbf{Q} \quad (5)$$

the MIMO channel effects and the iid white Gaussian noise $\mathbf{V}$ yields the received vector (over J consecutive OFDM symbol periods)

$$\mathbf{Y} = \underline{\mathbf{H}}\mathbf{W}\underline{\mathbf{Q}}\mathbf{A} + \mathbf{V} \quad (6)$$

Choice of the Linear Precoding Matrix $\mathbf{Q}$

Mutual Information Criterion

We use an information-theoretic criterion to restrict our search to information-lossless precoding matrices. Using equation (6), and assuming that the noise samples in $\mathbf{V}$ and the symbols in $\mathbf{A}$ are independent circularly symmetric, Gaussian random variables, the mutual information between $\mathbf{A}$ and $\mathbf{Y}$ is [7] :

$$\mathcal{I}(\mathbf{A};\mathbf{Y}) = \log\det\left(\pi e\, \mathcal{R}_{\mathbf{YY}}\right) - \log\det\left(\pi e\, \mathcal{R}_{\mathbf{VV}}\right)$$
$$= \log\det\left(\mathbf{I}_{JNP} + \rho\, \underline{\mathbf{H}}\mathbf{W}\left(\mathbf{I}_{MGJ} \otimes \mathbf{Q}\mathbf{Q}^H\right)\mathbf{W}^H\underline{\mathbf{H}}^H\right) \quad (7)$$

where $\rho = \frac{\sigma_a^2}{\sigma_v^2}$. Since the channel capacity is known to be

$$\mathrm{C} = \log\det\left(\mathbf{I}_{JNP} + \rho\, \underline{\mathbf{H}}\underline{\mathbf{H}}^H\right) \quad (8)$$

and since $\mathbf{W}\mathbf{W}^H = \mathbf{I_{JMP}}$ by definition of a permutation, we can infer that choosing $\mathbf{Q}$ among unitary matrices ($\mathbf{Q}\mathbf{Q}^H = \mathbf{I_{LM}}$) ensures that there is no capacity loss at the linear precoding stage ($\mathcal{I}(\mathbf{A};\mathbf{Y}) = \mathrm{C}$).

Pairwise Error Probability Criterion : Diversity and Coding Gain Analysis

To analyze the coding gain of the linear precoding by itself, let us consider a single error event, focusing on one particular symbol ($\mathbf{a}^{(k)}_{jG+g}$, with j, k and g fixed) : the decoded vector $\mathbf{a}'^{(k)}_{jG+g}$ differs from the transmitted one $\mathbf{a}^{(k)}_{jG+g}$ only in the u-th coefficient, $u \in \{1, \dots, ML\}$

$$\mathbf{E} = \mathbf{a}^{(k)}_{jG+g} - \mathbf{a}'^{(k)}_{jG+g} = (0, \dots 0, e_u, 0, \dots 0)^T \quad (9)$$

where $e_u \neq 0$ is any difference between two complex symbols of the constellation. With j, k and g fixed, equation (4) shows that the values in $\mathbf{x}^{(k)}_{jG+g}$ are affected to ML different tones, on one and only one Tx antenna for every tone involved. Denoting by $\{p_1 \dots p_{ML}\}$, $\{i_1 \dots i_{ML}\}$ and $\{j_1 \dots j_{ML}\}$ respectively the tone number, antenna number and OFDM symbol number corresponding to $\mathbf{x}^{(k)}_{jG+g}$, we can form a vector that gathers the LMN channel coefficients actually involved in its transmission :

$$\mathbf{C}_{j,k,g} = (h_{j_1,1,i_1,p_1} \dots h_{j_1,N,i_1,p_1}, \dots,$$
$$h_{j_{LM},1,i_{LM},p_{LM}}, \dots, h_{j_{LM},N,i_{LM},p_{LM}})^T$$

Similarly, at the receiver, $\mathbf{u}^{(k)}_{jG+g} =$

$$(y_{j,p_1,1} \dots y_{j,p_1,N}, \dots, y_{j,p_{LM},1} \dots y_{j,p_{LM},N})^T$$

The received signals corresponding to the transmission of $\mathbf{a}^{(k)}_{jG+g}$ can be written

$$\mathbf{u}^{(k)}_{jG+g} = \left(\text{diag}(\mathbf{Q}\mathbf{a}^{(k)}_{jG+g}) \otimes \mathbf{I}_N\right) \mathbf{C}_{j,k,g} \qquad (10)$$

therefore,

$$\mathbf{u}^{(k)}_{jG+g} - \mathbf{u}'^{(k)}_{jG+g} = \left(\text{diag}\left(\mathbf{Q}(\mathbf{a}^{(k)}_{jG+g} - \mathbf{a}'^{(k)}_{jG+g})\right) \otimes \mathbf{I}_N\right) \mathbf{C}_{j,k,g} \qquad (11)$$

which lets us write the squared Euclidean distance between the received samples as

$$\begin{aligned} &\left\|\mathbf{u}^{(k)}_{jG+g} - \mathbf{u}'^{(k)}_{jG+g}\right\|^2 \\ =\ & \mathbf{C}^H_{j,k,g}\left(\text{diag}(\mathbf{QE})^H \otimes \mathbf{I}_N\right)\left(\text{diag}(\mathbf{QE}) \otimes \mathbf{I}_N\right)\mathbf{C}_{j,k,g} \\ =\ & \mathbf{C}^H_{j,k,g}\Big(\left(\text{diag}(\mathbf{QE})^H \text{diag}(\mathbf{QE})\right) \otimes \mathbf{I}_N\Big)\mathbf{C}_{j,k,g}\ . \end{aligned} \qquad (12)$$

As shown in [2], assuming that the channel coefficients in $\mathbf{C}_{j,k,g}$ are complex, zero-mean, independent Gaussian random variables, the rank of $\left(\left(\text{diag}(\mathbf{QE})^H \text{diag}(\mathbf{QE})\right) \otimes \mathbf{I}_N\right)$ determines the diversity advantage of this coding scheme. It is obvious from equation (12) that if $\mathbf{Q}$ contains no zero, our scheme achieves diversity MNL.In the sequel, we will assume that this condition is fulfilled.

In order to maximize the coding gain at high SNR [2], $\mathbf{Q}$ should maximize

$$\begin{aligned} &\min_{n,e_n} \det\Big(\left(\text{diag}(\mathbf{QE})^H \text{diag}(\mathbf{QE})\right) \otimes \mathbf{I}_N\Big) \\ &= \min_{n,e_n} \det\left(\text{diag}(\mathbf{QE})^\dagger \text{diag}(\mathbf{QE})\right)^N \end{aligned} \qquad (13)$$

Which means that we have to maximize

$$\min_n \left(\prod_{i=1}^{ML} q_{i,n} q^*_{i,n}\right)^M \qquad (14)$$

under the energy constraint $\sum_{i=1}^{ML} q_{i,n} q^*_{i,n} = 1\ ,\ \forall n$. The Lagrange multiplier method yields

$$|q_{i,n}| = \frac{1}{\sqrt{ML}}\ ,\ \forall n, \forall i\ . \qquad (15)$$

A satisfactory solution w.r.t. both the mutual information and the pairwise error criteria, is the Vandermonde matrix

$$\mathbf{Q} = \frac{1}{\sqrt{ML}} \begin{pmatrix} 1 & \theta_1 & \dots & \theta_1^{ML-1} \\ 1 & \theta_2 & \dots & \theta_2^{ML-1} \\ \vdots & \vdots & & \vdots \\ 1 & \theta_{ML} & \dots & \theta_N^{ML-1} \end{pmatrix} \qquad (16)$$

$$\text{where} \quad \theta_k = e^{j\frac{\pi}{ML}(1+2k)},\ k = 1\dots ML$$

This choice ensures that we both achieve a MNL diversity order, and that the coding gain is maximized. A deeper analysis for QAM symbol constellations shows that the pairwise error is indeed worst for single error events (when no wrapping is performed) and that in that case the spreading matrix in (16) maximizes the coding gain when ML is a power of two [8]. Also, for ML decoding, the Matched Filter Bound is proportional to $\|\mathbf{C}_{j,k,g}\|^2$ which is a sum of squares of MNL Rayleigh variables and hence shows again the diversity order MNL.

DECODING

The decoding process is done in two steps, similarly to the encoding operations : the linear precoding and spreading operations are undone using linear minimum mean-square error (LMMSE) estimation

$$\hat{\mathbf{A}}_{LMMSE} = \mathcal{R}_{\mathbf{AY}} \mathcal{R}^{-1}_{\mathbf{YY}} \mathbf{Y} \qquad (17)$$

or its unbiased version

$$\hat{\mathbf{A}}_{ULMMSE} = \text{diag}\left(\mathcal{R}_{\mathbf{AY}} \mathcal{R}^{-1}_{\mathbf{YY}} \mathcal{R}_{\mathbf{YA}}\right)^{-1} \mathcal{R}_{\mathbf{AY}} \mathcal{R}^{-1}_{\mathbf{YY}} \mathbf{Y}. \qquad (18)$$

Since $\underline{\mathbf{H}}$ is block-diagonal, $\mathcal{R}_{\mathbf{AY}} \mathcal{R}^{-1}_{\mathbf{YY}}$ simplifies to tone-wise $M \times N$ LMMSE followed by depermutation ($\mathbf{W}^H$) and despreading ($\mathbf{Q}^H$). The estimates $\hat{a}^{(k)}_{jG+g,lM+i}$ are then demapped, deinterleaved, and fed into M decoders, which estimate the original uncoded data bits $\hat{\mathrm{d}}^{(k,0)}$ separately for each stream k. Several options are available, including hard-decision decoders (such as the Viterbi algorithm) or soft-decision decoders (like the BCJR algorithm [9]).

Iterative interference cancelation

Thanks to the constraints on the tone assignment, for one given tone p, the received signal is a superposition of signals from M different streams. This makes iterative interference cancelation straightforward: during the decoding of stream k at iteration c, the estimated interference is

$$\mathbf{Y}^{(\mathrm{k,c})}_{\mathrm{IC}} = \underline{\mathbf{H}}\mathbf{W}\underline{\mathbf{Q}}\hat{\mathbf{A}}^{(\mathrm{k,c})}_{\mathrm{IC}} \qquad (19)$$

where

$$\hat{\mathbf{A}}^{(\mathrm{k,c})}_{\mathrm{IC}} = \left(\hat{\mathbf{A}}^{(\mathbf{1,c})}, \dots, \hat{\mathbf{A}}^{(\mathbf{k-1,c})}, 0, \hat{\mathbf{A}}^{(\mathbf{k+1,c})}, \dots \hat{\mathbf{A}}^{(\mathbf{M,c})}\right)^T$$

is the expectation of the interference, obtained using re-encoded versions of $\hat{\mathrm{d}}^{(1,c-1)}, \dots, \hat{\mathrm{d}}^{(k-1,c-1)}, \hat{\mathrm{d}}^{(k+1,c-1)}, \dots, \hat{\mathrm{d}}^{(M,c-1)}$. For the decoding of stream k, at the c^{th} ($c \geq 1$) iteration, the LMMSE decoder is fed

$$\mathbf{Y} - \mathbf{Y}^{(\mathrm{k,c})}_{\mathrm{IC}} = \underline{\mathbf{H}}\mathbf{W}\underline{\mathbf{Q}}\hat{\mathbf{A}}^{(\mathrm{k,c-1})} + \mathbf{V} \qquad (20)$$

plus detection errors in the interference. The process is initialized with

$$\hat{\mathbf{A}}^{(\mathrm{k,0})}_{\mathrm{IC}} = \mathbf{0}\ . \qquad (21)$$

SIMULATIONS RESULTS

Simulations were carried out using a rate $\frac{1}{2}$ convolutional code, and QPSK mapping, with various flavours of decoders. In a first channel model, the frequency diversity was introduced by randomly generating L $N \times M$ channel coefficients for the L tone segments and replicating the channel values

Figure 3. Channel with 3 frequency segments

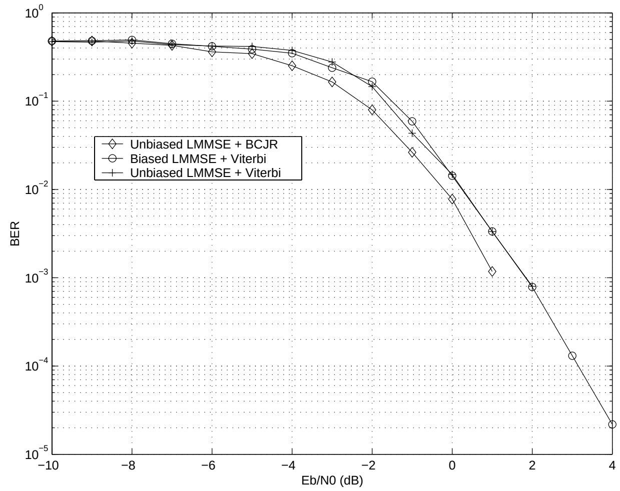
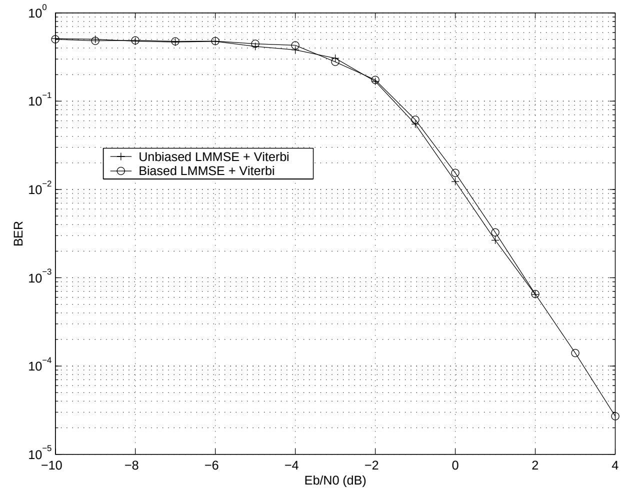


Figure 4. BRAN A Channel



over the P/L tones within a segment, similarly to what is often done in the time domain for a block-fading channnel. Simulation results for a 4×4 system with $L = 3$, and $J = 4$ over a 3 frequency segments channel with independent realizations are presented in Fig. 3. Notice that the soft-output BCJR algorithm outperforms the hard output Viterbi algorithm. Notice also that the suboptimal biased Linear MMSE estimator is virtually as efficient as its unbiased counterpart. Simulation results for a 4×4 system with $L = 3$, and $J = 16$ over a BRAN A [10] channel with independent realizations are presented in Fig. 4. Here again, the biased version of the LMMSE estimator is as good as the unbiased version.

ADDING TEMPORAL DIVERSITY

If the channel changes over the frame of J OFDM symbol periods, then similarly to the grouping of tones, one can group the OFDM symbol periodss in K groups of J/K where J/K OFDM symbol periods would correspond to the channel coherence time. In that case symbols get blocked into vectors $\mathbf{a}$ of size KML, spread by $\mathbf{Q}$ of that size and assigned to M different antennas in L different tone segments and K different time segments.

CONCLUSIONS

We proposed a hybrid coding scheme for MIMO OFDM systems based on separate streams and careful tone assignment ensuring that the streams are not self-interfering. This enables straightforward interference cancellation using a combination of a binary code, and a full-rate full-diversity linear spreading operation. Unlike many previous schemes such as the one proposed by Alamouti, it can accomodate any number of Tx antennas.

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