An Improved Detection Technique For Receiver Oriented MIMO-OFDM Systems

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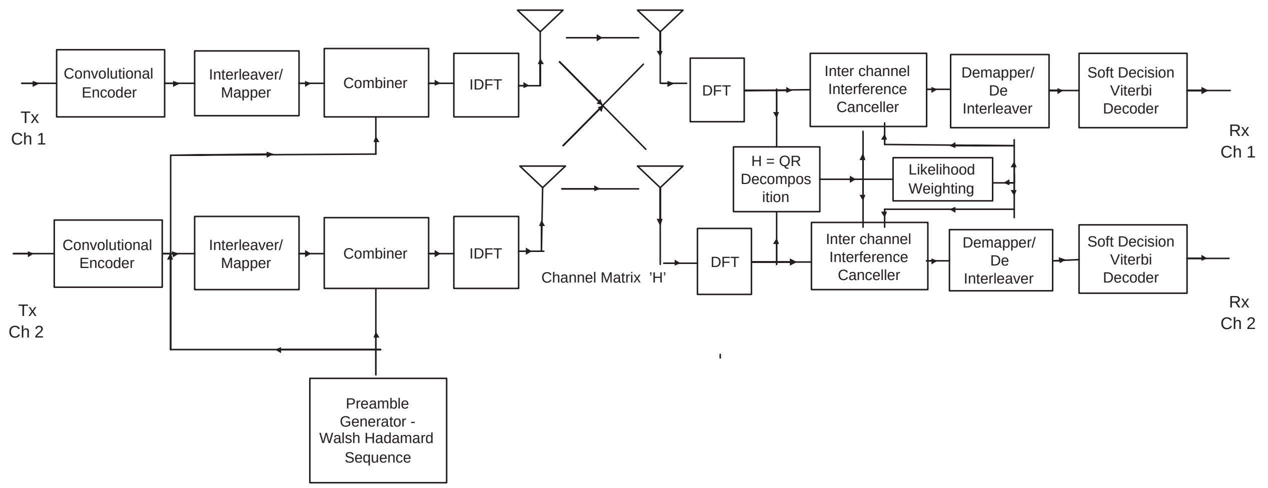
***Abstract*— In rich scattering environments, layered Space Time architectures like the BLAST (Bell Labs Layered Space Time) system may exploit the capacity advantage of multiple antenna systems. In case of a broadband wireless transmission, the channel undergoes frequency selective fading which necessitates the receiver to have a multi tap equalizer. This can be avoided by employing an OFDM (Orthogonal Frequency Division Multiplexing) scheme which would also increase the spectral efficiency of the system. In this paper, we present an efficient algorithm for detecting layered architectures as applied to an OFDM system. It utilizes a QR decomposition of the channel matrix for each subcarrier and a successive interference cancellation (SIC) structure. The performance of the above method is further improved by likelihood weighting of the detected symbols based on the estimated channel coefficients and soft decision Viterbi decoding.**

I. Introduction

The increasing need for fast, reliable and high data rate wireless communication links has opened the discussion about two different kinds of systems. Firstly, MIMO (Multiple Input Multiple Output) systems with multiple antennas located at both the transmitter (Tx) and receiver (Rx) have been shown to increase the capacity and performance [1]. Secondly, the tremendous growth in the multimedia applications makes it necessary to have a broadband wireless channel for high data rate communications. The fading experienced by such a broadband channel would be frequency selective accompanied by a large group delay.

In this work, we look upon how alternate modulation schemes such as OFDM (Orthogonal Frequency Division Multiplexing) could be used in conjunction with multiple antennas (MIMO) in a broadband wireless system. We study the case of a rich scattering environment normally encountered in an indoor communication system and concentrate on the receiver algorithms that could be employed to achieve a good system performance. A simple zero forcing scheme for a MIMO-OFDM receiver was proposed in [2]. It estimates the propagation matrix for each OFDM subcarrier and subsequently multiplies the received symbols by the inverse of that matrix to remove inter carrier interference (ICI) between the data streams from different transmit antennas. The main drawback of such a scheme is the complexity arising because of repeated computations of a matrix inverse, which increases with the number of antennas. In this paper, we concentrate on the alternate technique of a QR decomposition of the channel matrix $\mathbf{H}$ instead of the inverse computation. The algorithm iteratively estimates the transmitted symbols and allows the usage of successive interference cancellation (SIC) in between. These detected symbols are multiplied by a weighting factor based on the $\mathbf{H}$ matrix and fed into a soft decision Viterbi decoder which could further reduce the required signal-to-noise ratio (SNR) in a frequency selective fading channel. We then analyze an improvement in the QR decomposition method proposed by Wübben at al. [3] for frequency flat channels, wherein the channel matrix $\mathbf{H}$ is first sorted based on its column norms and then made to undergo QR decomposition. Such a scheme, known as sorted QR decomposition (SQRD) ensures detection of the strongest signal first, thereby reducing error propagation within the SIC. We will present performance evaluation of this method applied to the frequency flat subcarriers of a MIMO-OFDM system based on Rayleigh channel assumptions.

The organization of this paper is as follows. Section II describes the MIMO-OFDM system in general and the receiver scheme using the QR decomposition of $\mathbf{H}$, the improvement in performance that could be



Block Diagram of a 2 * 2 MIMO - OFDM System

Fig. 1. Schematic of a 2×2 MIMO-OFDM System with independent transmitter chains and interference cancellation

further obtained by sorting the channel matrix $\mathbf{H}$ before doing the QR decomposition and advantages of calculating the likelihood weighting factor based on the coefficients of channel matrix $\mathbf{H}$ before decoding the weighted signal by a soft decision Viterbi decoder. Section III gives the computer simulation results obtained, evaluating the performance of the proposed scheme and comparing it with that proposed in [2]. Section IV looks at the computational complexity when implemented on a DSP. Concluding remarks are found in section V.

II. System Model and Receiver Architecture

A. General system model

Figure 1 shows a block diagram of the MIMO-OFDM scheme with $N = 2$ data streams, $N_T = 2$ transmit and $N_R = 2$ receive antennas. Each data stream has its own convolutional encoder, interleaver and constellation mapper. Each encoded data stream is then distributed to I OFDM subcarriers. The preamble symbols, which are used for the estimation of the propagation matrix, are prefixed to the data symbols of each subcarrier. After an inverse Fourier transform (IFFT) operation the OFDM symbols of each data stream are transmitted through the N_T antennas. For each stream separately, the OFDM symbol is expanded with a cyclic prefix (CP) before transmission, which makes the channel a circulant one and avoids inter symbol interference (ISI) if the length of the prefix is greater than the maximum access delay of the channel. The transmitted symbols from different antennas superimpose each other at the N_R receiver antennas. After the removing of the cyclic prefix the remaining signal undergos a Fourier transform (FFT) operation to yield samples in the frequency domain. The relationship between the transmitted and the received data symbols of the i^{th} subcarrier may be expressed as follows

$$\mathbf{r}^i = \mathbf{H}^i \cdot \mathbf{s}^i + \mathbf{n}^i\,, \tag{1}$$

where $\mathbf{r}^i$ and $\mathbf{s}^i$ are the received and transmitted symbol vector respectively and $\mathbf{n}^i$ represents the zero mean, complex Additive White Gaussian Noise (AWGN) present at the receive antennas transformed into the frequency domain. The matrix $\mathbf{H}^i$ is the channel matrix per subcarrier and can be written as

$$\mathbf{H}^i = \begin{bmatrix} h^i_{1,1} & h^i_{1,2} & \cdots & h^i_{1,N_T} \\ h^i_{2,1} & h^i_{2,2} & \cdots & h^i_{2,N_T} \\ \vdots & & \ddots & \vdots \\ h^i_{N_R,1} & h^i_{N_R,2} & \cdots & h^i_{N_R,N_T} \end{bmatrix}. \tag{2}$$

It consists of the propagation coefficients $h_{p,q}$ of all possible $N_T \cdot N_R$ antenna connections for the i^{th} subcarrier ($1 \le p \le N_R, 1 \le q \le N_T$). The per subcarrier channel matrix can be estimated by calculating the propagation coefficient between each transmitted and received preamble symbol. If we assume noise free transmission of the preamble symbols, the relationship among the transmitted preamble symbol $\mathbf{C}^i$ and the received preamble symbol $\mathbf{B}^i$ can be described by

$$\mathbf{B}^i = \mathbf{H}^i \cdot \mathbf{C}^i\,. \tag{3}$$

The preamble symbols in $\mathbf{C}^i$ must be orthogonal to be able to dissolve the channels at the receiver. This can be reached by using a Space-Time coded preamble. The simplest orthogonal structure for $\mathbf{C}^i$ that could be used is a Walsh-Hadamard matrix multiplied by an arbitrary complex symbol c^i[4]. Then the channel could be estimated in the frequency domain as

$$\mathbf{H}^i = \mathbf{B}^i \cdot (\mathbf{C}^i)^{-1}. \tag{4}$$

B. QR decomposition of H

In the zero forcing receiver case the received symbol vector $\mathbf{r}$ would be multiplied by the pseudo inverse channel matrix $\mathbf{H}^+$ to yield the estimates for the transmitted symbols[1]. The drawback of this receiver is not only the possible noise amplification, but the symbol from one transmit antenna is decided with full interference from all others. We will use a QR decomposition instead of the pseudo inverse. The estimated channel matrix $\mathbf{H}$ is decomposed into the unitary $N_R \times N_T$ dimensional matrix $\mathbf{Q}$ ($\mathbf{Q}^H\mathbf{Q} = \mathbf{I}$) and the $N_T \times N_T$ dimensional upper triangular matrix $\mathbf{R}$:

$$\mathbf{H} = \mathbf{QR}. \tag{5}$$

In a preprocessing step the vector of received symbols $\mathbf{r}$ is multiplied with the Hermitian of the matrix $\mathbf{Q}$

$$\begin{aligned} \mathbf{y} = \mathbf{Q}^H\mathbf{r} &= \mathbf{Q}^H(\mathbf{Hs}+\mathbf{n}) \\ &= \mathbf{Q}^H(\mathbf{QRs}+\mathbf{n}) \\ &= \mathbf{Rs}+\mathbf{Q}^H\mathbf{n}. \end{aligned} \tag{6}$$

Note that, $\mathbf{Q}$ being a unitary matrix, the variance of noise term remains unaffected. $\mathbf{R}$ being an upper triangular structure, the transmitted symbols $\mathbf{s}$ could be detected from the vector $\mathbf{y}$ by employing a Gauss elimination algorithm. The symbol from the last transmit antenna is detected first and is then used to cancel the interference onto the other transmit symbols. For e.g., the k^{th} element of $\mathbf{y}$ is given by

$$y_k = r_{k,k} \cdot s_k + \sum_{i=k+1}^{N_R} (r_{k,i} \cdot s_i) + n_k, \tag{7}$$

and is free of interference from layers $1, 2, \ldots, k-1$. So, by starting with the last element y_{N_R} and applying a successive interference cancellation (SIC) scheme, all transmitted symbols can be detected by

$$\tilde{s}_k = Q\left[\frac{y_k - \sum\limits_{i=k+1}^{N_R} r_{k,i}\tilde{s}_i}{r_{k,k}}\right]. \tag{8}$$

[1]Subsequently the index i is dropped for convenience, but the equations still apply to the i^{th} subcarrier.

No. of antennas	2×2 and 4×4
No. of FFT points	64
No. of data subcarriers	48
Modulation	QPSK
Interleaving	Depth of 16 bits
Forward Error Correction	Rate 1/2 Convolutional encoding (Constraint length = 7), 6 bit soft decision Viterbi decoding
Channel model	18 ray Rayleigh fading (HiperLAN/2 model A: 50ns RMS delay spread, 50Hz max. Doppler frequency)

TABLE I
SIMULATION PARAMETERS

C. Sorted QR Decomposition (SQRD)

The receiver described in the previous paragraph considers an arbitrary $\mathbf{H} = \mathbf{QR}$ decomposition and then employs the SIC scheme. This SIC procedure has the drawback that any error in detecting the transmitted symbol is propagated further into subsequent symbols due to interference subtraction. To minimize this error propagation probability the signal transmitted over the strongest channel should be detected first with a high reliability. After SIC the second strongest signal would follow. In this way, the signal detected last would be that transmitted across the weakest channel with the smallest SNR and the highest detection error probability, but would not contribute to the SIC anymore. The optimum sorting method would ensure that the values on the main diagonal of $\mathbf{R}$ are in increasing order and are the ones with the maximum value over all possible permutations of the channel matrix $\mathbf{H}$. However, a suboptimal but computational less complex sorting can be incorporated into the Gram-Schmidt orthogonalization procedure [3]. Now, $r_{1,1}$ is simply the norm of the column vector h_1, so the first optimization in the SQRD algorithm consists merely of permuting the column of $\mathbf{H}$ with minimum norm to this position. During the following orthogonalization of the vectors h_2,...,h_{N_T} with respect to the normalized vector h_1, the first row of $\mathbf{R}$ is obtained. Next $r_{2,2}$ is calculated in a similar manner from the remaining $N_T - 1$ orthogonalized vectors and so on. Thereby, the channel matrix $\mathbf{H}$ is successively transformed into the matrix $\mathbf{Q}$ associated with the desired ordering, while the corresponding $\mathbf{R}$ is calculated row by row. A point to be emphasized is that the column norms have to be calculated only once at the start of the algorithm and then can be easily updated afterwards. Hence the computational overload due to sorting is negligible.

D. Likelihood Weighting

Comparable to the linear zero-forcing receiver the detected symbols after the SQRD-SIC algorithm do not contain information about channel quality anymore. From (8) it can be seen that prior to the symbol decision of $\tilde{s}_k$ the information vanishes due to the division by $r_{k,k}$. Instead, the channel attenuation information could be effectively utilized such that the detected symbols are multiplied by a weighting vector $\mathbf{w}$, which is calculated as the square root of the average SNR from the norms of the columns of the per subcarrier channel matrix $\mathbf{H}$:

$$\mathbf{v} = \mathbf{w} \cdot \tilde{\mathbf{s}}\,, \tag{9}$$

where $\tilde{\mathbf{s}}$ is the vector of quantized symbols. The vector of likelihood weighted data symbols $\mathbf{v}$ is then decoded using a soft input Viterbi decoder. This $\mathbf{H}$ matrix based likelihood weighting influences the computation of path metrics in soft Viterbi and results in an improvement in the error correction performance as seen in the performance analysis curves.

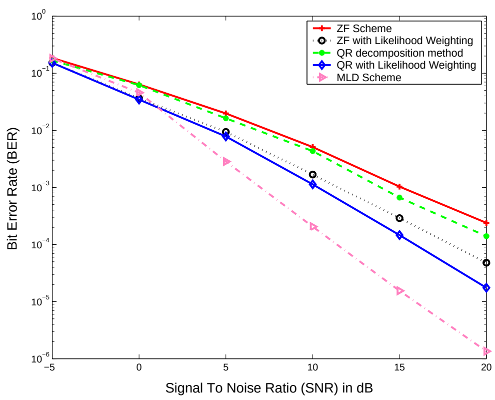


Fig. 2. Comparison between ZF and QR decomposition method for a 2×2 system with and without likelihood weighting

III. Performance Analysis

In this section, we investigate the bit error rate performance (BER) for MIMO-OFDM systems with $N = N_T = N_R = 2$ and $N = N_T = N_R = 4$ transmit and receive antennas. The simulation parameters are

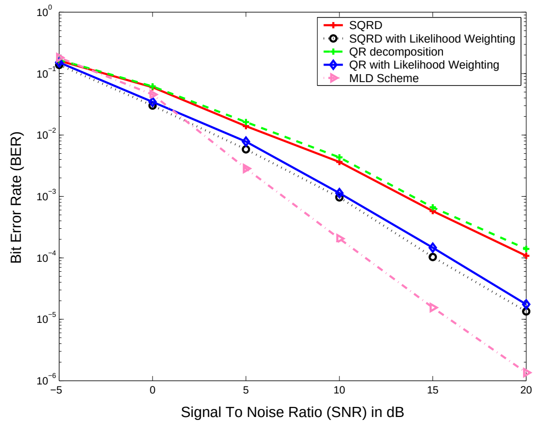


Fig. 3. Comparison between QR and Sorted QR decomposition method for a 2×2 system with and without likelihood weighting

listed in table I. The QR decomposition method offers a distinct improvement in performance over zero forcing receivers (ZF) by inverse channel detector (ICD) method, both in case with likelihood weighting and without likelihood weighting. This can be seen in figure 2. The reasons for that can be attributed to two facts:

1) The QR method employs a serial detection scheme wherein the interference from other transmitted symbols are cancelled before detection.
2) Unlike ZF using ICD, wherein small eigenvalues of $\mathbf{HH}^H$ will lead to large errors due to noise amplification, there is no noise amplification in the QR decomposition method.

The SQRD method, which sorts the channel matrix $\mathbf{H}$ before doing the QR decomposition gives a marginal improvement in performance for the 2×2 system (see figure 3). This is because in a 2×2 system, the number of permutations of the channel matrix $\mathbf{H}$ and therefore the number of possible QR decompositions is only two. The improvement due to sorting before the QR decomposition is more prominent for systems with higher number of antennas, e.g. 4×4 as in figure 4.

IV. Computational Complexity

The computational efforts required for the QR method have been computed on the basis of MAC (Multiply Add Compare) operations when implemented on a *SHARC* DSP of *Analog Devices* and compared with the required operations for a maximum likelihood detection (MLD) scheme. The QR

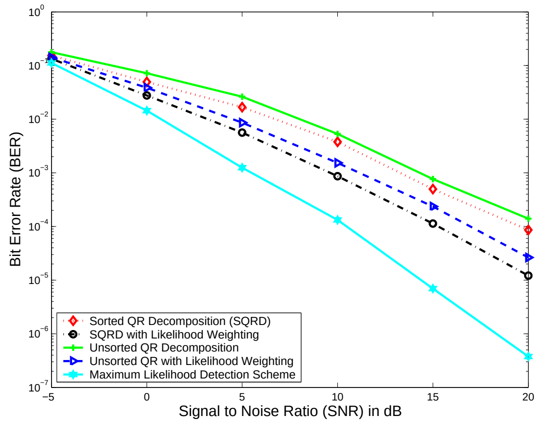


Fig. 4. Comparison between QR and SQRD method for a 4×4 system with and without likelihood weighting

decomposition method requires

$$\begin{aligned}&(N_T(N_R N_T + N_T + 1) + 1) \times \\ &((N_T - 1) + (N_T - 2) + \ldots + 1) \\ &+ 4N_R N_T + 6N_T\end{aligned}$$

MAC operations. Figure 5 compares the number of operations for the MLD scheme with the QR decomposition method. The complexity of the QR method is a polynomial function of the number of transmit and receive antennas whereas for the MLD scheme, the number of operations grows exponentially.

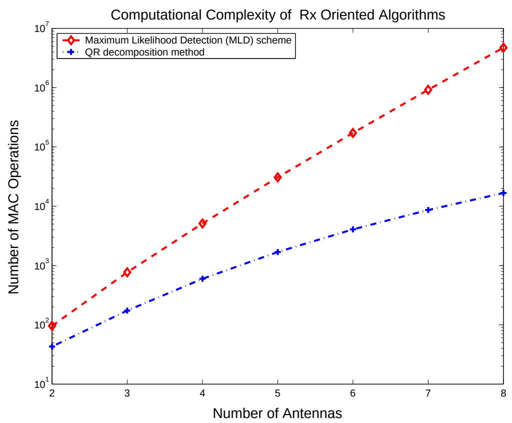


Fig. 5. Comparison of computational complexity for the QR decomposition mehtod and the MLD scheme

V. Conclusions

We have proposed a new detection algorithm for a MIMO-OFDM system based on modified Gram-Schmidt procedure. The algorithm carries out a QR decomposition of the channel matrix $\mathbf{H}^i$ for each OFDM subcarrier. The sorted QR method permutes the channel matrix based on its norms before doing the QR decomposition, thereby resulting in detection of stronger signals first. This minimizes the probability of error propagation in the successive interference canceller.

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