# Reducing the Complexity of MLD by combining ZF in OFDM/SDM system

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## 1. Introduction

In recent years, Orthogonal Frequency Division Multiplexing (OFDM) has been widely studied and applied in many broadband wireless systems, because it simplifies the receiver and leads to inexpensive hardware implementation by exploiting Inverse Fast Fourier Transform (IFFT) and proper circle prefix (CP) [1][2]. On the other hand, theoretical studies [3] of Multiple-Input-Multiple-Output (MIMO) systems, here restricted to Space Division Multiplexing (SDM) systems, have shown that systems that employ multiple transmit and receive antennas are capable of providing much larger transmission capacity than Single-Input-Single-Output (SISO) systems occupying the same bandwidth. In order to meet the demand for higher data rates, and more spectrally efficient wireless communication, combining OFDM with SDM seems to be a promising solution.

Many spatial multiplexing signal detection schemes [4][5][6] have been developed. Among these studies, it has been demonstrated that Maximum Likelihood Detection (MLD) [4] has the best performance. However, it is necessary to exhaustively evaluate all possible symbol vectors, so its complexity increases exponentially against the transmit antenna number or/and constellation size. To reduce the complexity, Viterbo proposed the sphere-decoding algorithm, which reduces the search area but it needs pre-computation of the radius of the corresponding searching sphere [7]. Ashina attempted to reduce the number of possible symbol vectors according to the upper bits of the receive signal, but it suffers from another complexity of generating a lookup table [8].

In this paper, we propose a novel detection scheme that combines ZF and MLD. From the viewpoint of complexity, Zero-Forcing (ZF) detection [6] is an attractive approach since it separates the spatial multiplexing signals by just linearly combining the receive signals. The computation cost of ZF is far less than that of MLD, so it is more practical in systems with large numbers of transmit and receive antennas. This motivated us into developing an algorithm that possessed the advantages of both schemes while receiving a small performance penalty compared to MLD.
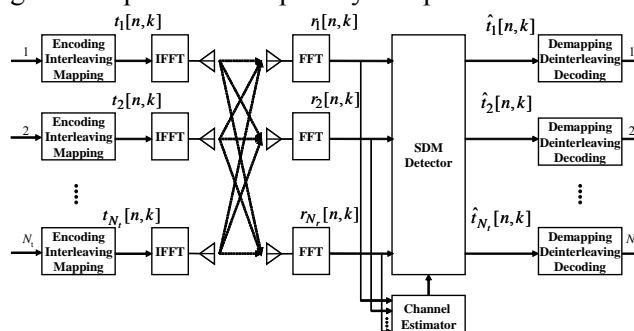


**Fig. 1** OFDM/SDM wireless link

## 2. OFDM/SDM system model

An OFDM/SDM wireless link using $N_t$ transmit $N_r$ receive antennas with $K$ subcarriers is presented in Fig. 1 We assume the channel is frequency-selective and time-invariant within at least one

OFDM block, which is reasonable for low mobility broadband wireless communication systems. The discrete frequency channel impulse response between transmit antenna $j$ (=1,...,$N_t$) and receive antenna $i$ (=1,...,$N_r$) on OFDM block $n$ (=-∞,...,0,...,+∞) over subcarrier $k$ (=0,...,$K$-1) is described as:

$$H[n,k] = \sum_{l=0}^{L-1} a[n,l] e^{-j\frac{2\pi}{K}kl} , \qquad (1)$$

$L$ stands for the FIR channel filter tap number, $a_{ij}[n,l]$ is the complex Gaussian channel gain between the $j$-th transmit antenna and the $i$-th receive antenna at the $l$-th path, with zero-mean and variance $\sigma_l^2$. In addition, the channel gains at the different paths are independent for each propagation channel. On the transmit side at time $n$, after $N_t$ parallel bit streams are encoded, interleaved, and constellation mapped (or subcarrier modulated), each bit stream is broken into OFDM blocks, for the $j$-th transmit antenna on $n$-th OFDM block at $k$-th subcarrier as denoted by $t_j[n,k]$. The time domain OFDM blocks produced by the IFFT circuit are then extended with the cyclic prefix (CP), which is assumed to be greater than or equal to $L$, and transmitted by the appropriate antennas. Note that the signal on each receive antenna is a mixture of the $N_t$ fading experienced transmitted signals, so we can express the input-output relation in vector form as:

$$\mathbf{r}[n,k] = \mathbf{H}[n,k]\mathbf{t}[n,k] + \mathbf{w}[n,k] , \qquad (2)$$

where $\mathbf{r}[n,k]=[r_1[n,k],...,r_{Nr}[n,k]]^T$, $\mathbf{t}[n,k]=[t_1[n,k],...,t_{Nt}[n,k]]^T$, $\mathbf{w}[n,k]=[w_1[n,k],...,w_{Nt}[n,k]]^T$,

$$\mathbf{H}[n,k] = \begin{bmatrix} H_{11}[n,k] & \cdots & H_{1N_t}[n,k] \\ \vdots & \ddots & \vdots \\ H_{N_r1}[n,k] & \cdots & H_{N_rN_t}[n,k] \end{bmatrix},$$

$(.)^T$ denotes transpose, and $\mathbf{w}[n,k]$ denotes the additive complex Gaussian noise vector, with each entry has zero-mean and common variance $\sigma_w^2$. The spatial multiplexing signals are separated by any detection scheme and the remaining processes performed on each substream are just the reverse versions of the processes done on the transmit side. We emphasize here that the $N_t$ different OFDM blocks are transmitted by different antennas in the same frequency band simultaneously. We also assume the following conditions hold true.

a) The average power of the complex symbol is normalized to $E[|t_j[n,k]|^2]=1$.

b) All $N_t \times N_r$ propagation channels are independent and satisfy the condition of $E[|H_{ij}[n,k]|^2]=1/N_t$, which ensures that the average SNR at each receive antenna equals $1/\sigma_w^2$, regardless of the number of antennas on the transmit side.

## 3. Proposed detection scheme - Combining ZF and MLD

The demerit of MLD is its complexity, as $N_t$ or/and constellation size $M$ becomes large, the size of the possible symbol vector set $\Omega$ increases exponentially. In order to avoid an exhaustive evaluation over $\Omega$, we attempt to extract a smaller subset from $\Omega$ according to the signal vector estimated by ZF. The criterion is explained easily using Fig. 2(a). For example, we use 16QAM in an OFDM/SDM system with $N_t \geq 2$, and illustrate the constellations corresponding to transmit antenna 1 and 2 on the $n$-th OFDM block at the $k$-th subcarrier. The double circles denote the signals estimated by ZF, $d$ stands for straight line distance between two adjacent complex symbols. The receive signals are separated according to the next three steps. **Step 1**, perform ZF and quantize the estimate $\hat{t}_j$ to complex symbol $\tilde{t}_j$, which denoted by the single circle. **Step 2**, take all $M_j$ points (complex symbols) lying inside the circle centered at $\tilde{t}_j$ with radius $R$ as candidates corresponding to the $j$-th antenna. In the example here $M_1$=9 and $M_2$=4. Next, form the symbol vector set $\Omega_R$ by the next equation:

$$\Omega_R = \Omega_1 \otimes \Omega_2 \otimes , ... , \otimes \Omega_{N_t} . \qquad (3)$$

Here $\otimes$ denotes the Cartesian product and $\Omega_j$ denotes the set including all $M_j$ points. **Step 3**, perform the Euclidean metric calculation over set $\Omega_R$ as follows:

$$\hat{\mathbf{t}} = \arg \min_{\mathbf{t}_u \in \Omega_R} \left\| \mathbf{r}[n,k] - \mathbf{H}[n,k]\mathbf{t}_u[n,k] \right\| \qquad u = 1,...,Q_R , \qquad (4)$$

where $Q_R = M_1 \times M_2 \times ,...,\times M_{Nt}$. By applying these three steps above, the number of possible candidates

use to be evaluated in the Maximum Likelihood sense, in other words, the number of metric calculations is reduced from $Q$ $(=M^{Nt})$ to $Q_R$ $(\leq M_R^{Nt})$, where $M_R$ denotes the maximum value of $M_j$. For a certain specific kind of constellation, $M_R$ depends only on radius $R$. In the case of Fig. 2(a), for a 16QAM constellation, if $R = d, \sqrt{2}\, d$ then $M_R = 5, 9$. Note that this proposal is a form of suboptimal detection because it is not capable of guaranteeing that the solution of equation (4) is identical to that of original MLD, except when $\Omega_R$ is identical to $\Omega$, the case in which $R$ is large enough to cover the entire transmit constellation.

The system diagram of the proposed SDM detector is given in Figure 2(b). On the receive side, after Fast Fourier Transform (FFT), **Step1:** the $N_r$ parallel received signals are separated by the ZF approach. **Step 2**: the Candidate Generation block then generates symbol vector set $\Omega_R$, and then **Step 3**: the metric calculation is performed over $\Omega_R$ to locate the estimate of transmitted signal vector. The remaining processes are identical to those of the conventional scheme.



(a) 16QAM Transmit constellation with $N_t \geq 2$      (b) System diagram
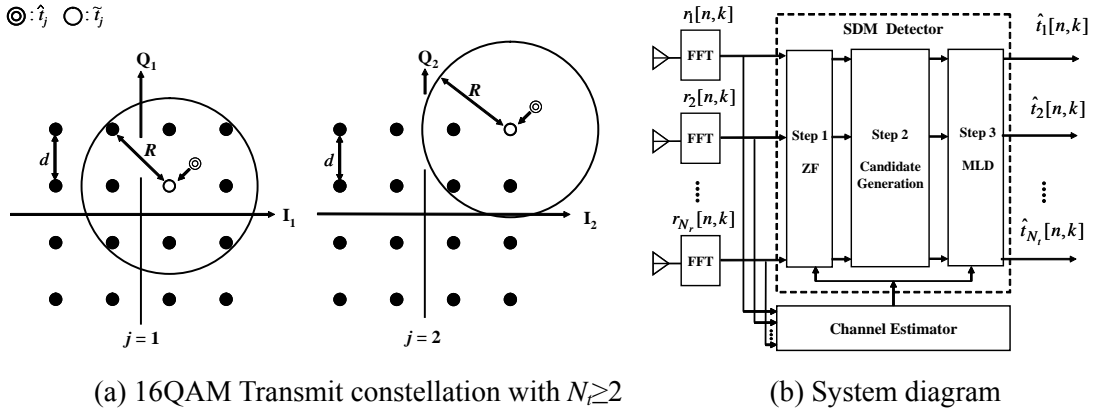
**Fig. 2** Concept and system structure of proposal

## 4. Simulation and Discussion

**Table1**. Simulation parameters

| Number of antennas | $N_t = Nr = 2, 3$ |
|---|---|
| Subcarrier modulation | 64QAM |
| Information bit rate | 108, 162Mbps |
| Channel encoding | Punctured Convolutional Code, Rate = ¾, Constraint length = 7 |
| Channel decoding | Viterbi algorithm, Hard Decision Decoding |
| Packet size | 64 byte for each Tx antenna |
| FFT point | 64 |
| Number of subcarriers | 48 |
| OFDM symbol length | 4micros |
| GI length | 800ns |
| Channel model | Exponentially decaying Rayleigh fading, RMS Delay Spread=100ns |

The packet error rate (PER) performance of an OFDM/SDM system with $N_t=N_r=2, 3$ using 64QAM subcarrier modulation was evaluated by computer simulations. OFDM-related parameters listed in Table 1 were taken from IEEE802.11a. Each propagation channel between transmit-receive antenna pair $(i,j)$ was assumed to be exponentially decaying Rayleigh fading with root-mean-square (RMS) delay spread of 100nsec, Doppler frequency shift was not considered. The channel estimation and timing synchronization were assumed to be ideal. In addition, we assume that the required $E_b/N_0$ was evaluated at the PER of $10^{-2}$.

The performance of the proposed scheme with $N_t=N_r=2, 3$ utilizing hard decision decoding (HDD) for various values of $M_R$ is illustrated in Figure 3. It should be noted that ZF is identical to the proposed scheme if $M_R=1$ and MLD is identical to proposed scheme if $M_R=64$. For $M_R=5$ with $N_t=N_r=2$ (3) the performance improvement at the required $E_b/N_0$ over ZF is about 2.0dB (2.0dB). When $M_R$ is increased from 5 to 9, the performance improvement over ZF is about 4.0dB (5.0dB), and

the performance degradation from MLD is suppressed to about 1.5dB (3.0dB). This implies that $M_R$, i.e., $R$ can be used as an index for assessing the trade-off between performance and complexity. This feature is thought to be useful in the system design stage. In the case evaluated here, for $M_R$=9, i.e., $R=\sqrt{2}\ d$, the proposed scheme reduces the number of metric calculations from $64^2$ ($64^3$) to less than $9^2$ ($9^3$) with $N_t$=2 (3).



(a) $N_t$=$N_r$=2                    (b) $N_t$=$N_r$=3
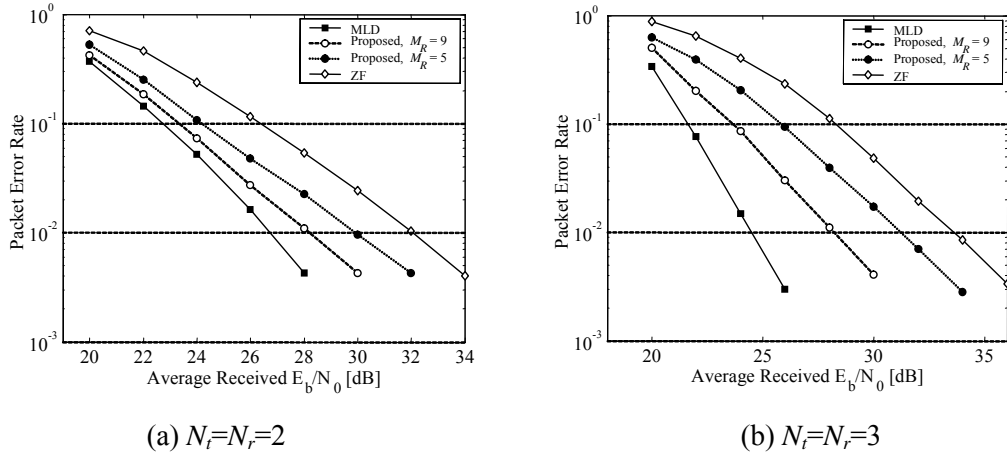
**Fig. 3** Performance versus $M_R$ (or $R$)

## 5. Conclusion

This paper has proposed a novel detection scheme for spatial multiplexing signals. The scheme reduces the complexity of MLD by a combination of ZF and MLD. The effect of the proposed scheme in an OFDM/SDM system was verified by computer simulations using OFDM-related parameters in IEEE802.11a. For $M_R$=9 with $N_t$=$N_r$=2 (3) the proposed scheme reduces the number of metric calculations from $64^2$ ($64^3$) to less than $9^2$ ($9^3$). Simulation results show that it achieves about 4.0dB (5.0dB) performance improvements over ZF and incurs a 1.5dB (3.0dB) performance penalty against MLD at the PER of $10^{-2}$.

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