An Interference Suppression Scheme for MIMO-OSTBC System

JIE WANG[1], ZHONGHAO ZHANG[2], YUEHUAN GONG[1], Tao Liu[1]
[1]School of Electronic Engineering
University of Electronic Science and Technology of China
Chengdu, 610054
CHINA
wehco@hotmail.com
[2]Department of Electronic Engineering
City University of Hong Kong
HongKong
CHINA

Abstract: - Multiple Input Multiple Output (MIMO) technique is considered as a cost-effective and promising technique for the next generation wireless system. Orthogonal Space-Time Block Codes (OSTBC) scheme is one of the most important coding schemes for MIMO system. The performance of an MIMO system using OSTBC (MIMO-OSTBC), however, degrades significantly in presence of strong interferences. In this paper, an interference suppression scheme for MIMO-OSTBC system in strong interference scenario is presented. The proposed scheme is based on simplified Maximum Likelihood (ML) decoding and noise subspace projection. The scheme has the following advantages: realizing ML decoding of the OSTBC and greatly reducing searching calculations by simple linear processing; avoiding matrix-inversion computation needed in ordinary suppression schemes through noise subspace projection. The paper also discusses the method of obtaining the efficient channel state information (CSI) needed in the proposed scheme in strong interference scenario. Simulation results are given to verify the interference suppression ability of the proposed scheme and our theoretical analysis. Compared with other suppression scheme, the proposed scheme has better SER performance and is more flexible in practical applications.

Key-Words: - MIMO, OSTBC, ML Decoding, Interference Suppression, Noise Subspace Projection,

1 Introduction

The next generation communication system should transmit information reliably with high data rate, and try to attain as high communication capacity as possible. However, the wireless channel is usually unreliable, and it suffers fading caused by multi-path propagation, which significantly affect communication performance.

In recent years, Multiple Input Multiple output (MIMO) technique [1-3] using multiple antennas at both transmitter and receiver has attracted much attention in wireless communication field. MIMO technique can significantly improve communication capacity by making use of the multi-path propagation which was usually considered to be harmful to the system performance. By exploiting some diversity processing techniques, for example, an MIMO system using Orthogonal Space-Time Block Codes (OSTBC) can significantly improve spectral efficiency and link reliability in rich scattering environment [4-6]. Compared with other diversity processing techniques, OSTBC do not require Channel State Information (CSI) at transmitter and has low decoding complexity at receiver. Therefore, MIMO technique using OSTBC is widely considered as a promising technique in the next generation communication system and has been adopted in proposals in 3GPP evolution [7] and IEEE 802.16m [8].

In a practical scenario, unfortunately, MIMO-OSTBC system may face many kinds of interferences, such as multi-user interferences, directional interferences and so on. These interferences may cause MIMO system suffer significant performance loss and even make MIMO system fail to transmit useful information. Therefore, it is very important and necessary to build an MIMO system with a high anti-interference ability.

The suppression of multi-user interferences for MIMO system has been concerned in many references [9-10]. Directional interference exists widely. However, literatures focusing on suppression of directional interference in MIMO system are limited.

For MIMO system with Alamouti coding scheme [6], an adaptive Digital Beam Forming (DBF) reception scheme based on Minimum Mean Square Error (MMSE) criterion to suppress Gaussian distributed directional interferences was presented in [11]. And in [12], an interference cancellation scheme based on a simplified Maximum Likelihood (ML) decoding is presented for MIMO system with Alamouti coding scheme. An extended version of the MMSE-DBF reception scheme [11] is presented in [13], which can support any form of OSTBC. The simulation results of these interference suppression schemes show that they are quite effective for interference suppression. However, there are still some problems remain unsolved, for example, how to avoid the matrix-inversion computation needed in the interference cancellation scheme presented in [12], how to obtain the CSI needed in decoding at the receiver in strong interference scenario in practical applications, and so on.

In this paper, an MIMO-OSTBC system in strong directional Gaussian distributed interference scenario is discussed, and an interference suppression scheme based on simplified ML decoding and noise subspace projection is presented. Compared with conventional ML decoding scheme, the proposed scheme can achieve ML decoding with much less searching calculations. The proposed scheme achieves detection symbol by symbol through simple linear processing unlike conventional ML decoding scheme, where the symbols are detected jointly. And through noise subspace projection, the proposed scheme avoids matrix-inversion computation compared with the scheme in [11]. Moreover, the method of obtaining the efficient CSI needed in the proposed scheme in strong interference scenario is also discussed in this paper. The simulation results show that the proposed scheme is effective on interference suppression and well in accordance with our theoretical analysis. Compared with the scheme in [11, 13], the performance of the proposed scheme is better and it is less sensitive to the length of pilot symbols. Furthermore, since the proposed scheme supports any form of OSTBC, it is more flexible in practical applications compared with the schemes in [11, 12].

This paper is organized as follows: In section 2, the system model for MIMO-OSTBC system in interference case is discussed. In section 3, conventional ML decoding scheme for interference suppression in interference case is reviewed, by simple linear processing, the simplified ML decoding scheme which supports any form of OSTBC for interference suppression is derived in section 4. And then in section 5, the interference suppression scheme based on simplified ML decoding and noise subspace projection is introduced and how to obtain the channel information required in the scheme at the receiver is presented. The simulation results are given in section 6. We conclude the paper in section 7.

2 System Overview

An MIMO-OSTBC system, using N_T transmit antennas (Tx), N_R receive antenna (Rx), with N_J ($N_J < N_R$) interferences is illustrated in Fig.1. At the transmitter, a block of L symbols $s_1, s_2, \ldots, s_L$ taken from a constellation $\mathcal{S}$ is fed into an OSTBC encoder which produces an $N_T \times T$ space-time code $\mathbf{C}$. Denoting $\mathbf{s} = [s_1, s_2, \cdots s_L]^T$, the output of OSTBC encoder can be written as $\mathbf{C}(\mathbf{s})$. Note that for any OSTBC codeword, Eq.(1) should be satisfied [4]

$$\mathbf{C}(\mathbf{s})\mathbf{C}(\mathbf{s})^H = \alpha \sum_{l=1}^{L} |s_l|^2 \mathbf{I}_{N_T} \ , \tag{1}$$

where α is a constant and depends on code rate. For power constraint reasons, we assume that $E(|s_l|^2) = 1$.

The MIMO channel between the transmitter and receiver can be described by a matrix $\mathbf{H}$ of dimension $N_R \times N_T$. We restrict the channels between the transmitter and receiver to quasi stable frequency-flat Rayleigh fading channels. The element of $\mathbf{H}$, $[\mathbf{H}]_{n_R n_T}$ represents the channel coefficient between the $n_T th$ transmit antenna and the $n_R th$ receive antenna ($n_T = 1, 2, \ldots, N_T$ and $n_R = 1, 2, \ldots, N_R$). We also assume that the elements of $\mathbf{H}$ have the independent and identically distributed (i.i.d) complex Gaussian distribution with zero mean and unit variance, and are spatially uncorrelated to each other, i.e., $[\mathbf{H}]_{n_R n_T} \sim CN(0, 1)$, for $n_T = 1, 2, \ldots, N_T$ and $n_R = 1, 2, \ldots, N_R$.

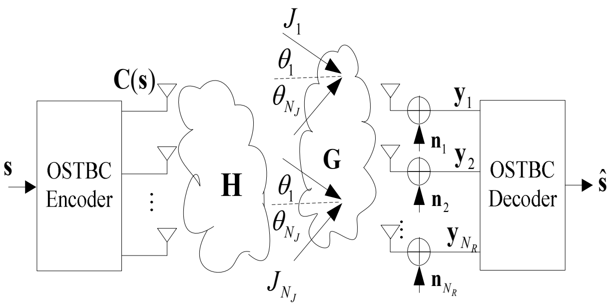

Fig.1 MIMO-OSTBC system with N_J interferences

In interference-free case, we assume that the transmission power are allocated equally across all the transmit antennas, and that pulse shaping, match-filtering, sampling are all optimal. Therefore, the received signal $\mathbf{Y}=[\mathbf{y}_1,\mathbf{y}_2,\cdots,\mathbf{y}_{N_R}]^T$ can be expressed as

$$\mathbf{Y}=\sqrt{\frac{\rho}{N_T}}\mathbf{HC(s)}+\mathbf{N} , \qquad (2)$$

where $\mathbf{N}=[\mathbf{n}_1,\mathbf{n}_2,\cdots\mathbf{n}_{N_R}]^T$ is the noise matrix of dimension $N_R\times T$ whose entries are independent to each other and distributed according to $CN(0,\ 1)$, and ρ is the signal-to-noise ratio (SNR).

Considering the case that there are several Gaussian distributed interference resources and each uses single antenna to transmit interference signal. They may suffer scattering since there are often rich scatterers around the receiver. Assume that the main power component of the n_J th interference arrives at the receiver directly from the direction of θ_{n_J} ($n_J=1,2,\cdots N_J$), which is called direct component $P_{D_{n_J}}$, and the rest component suffers Rayleigh fading and arrives at the receiver in scattered paths, which is called scattering component $P_{S_{n_J}}$. And we define a factor K_{n_J} as the ratio of power in direct component $P_{D_{n_J}}$ to the power in scattering component $P_{S_{n_J}}$, i.e.,

$$K_{n_J}=P_{D_{n_J}}/P_{S_{n_J}} .$$

Define the vector $\mathbf{g}_{n_J}$ of dimension $N_R\times 1$ as the channel vector between the n_J th interference source and the receiver. According to the above assumptions and definitions, $\mathbf{g}_{n_J}$ can be modeled as

$$\mathbf{g}_{n_J}=\sqrt{\frac{K_{n_J}}{K_{n_J}+1}}\mathbf{a}\left(\theta_{n_J}\right)+\sqrt{\frac{1}{K_{n_J}+1}}\tilde{\mathbf{g}}_{n_J}\ n_J=1,2,\cdots N_J ,$$

where $\mathbf{a}\left(\theta_{n_J}\right)$ is the steer vector of the n_J th interference, and the elements of $\tilde{\mathbf{g}}_{n_J}$ are circularly symmetric zero mean complex Gaussian random variables with unit variance, which indicates the channel information of the scattered paths.

Therefore, the channel between interferences and the receiver can be described by the matrix $\mathbf{G}=[\mathbf{g}_1,\mathbf{g}_2,\cdots\mathbf{g}_{N_J}]$ of dimension $N_R\times N_J$. Define the n_J th interference-to-noise ratio (INR) as η_{n_J} . And $\mathbf{\Phi}$ is a diagonal matrix of dimension $N_J\times N_J$, where $\left[\mathbf{\Phi}\right]_{n_J,n_J}=\sqrt{\eta_{n_J}}$, and η_{n_J} is the n_J *th* interference-to-noise ratio (INR).

The interference signal can be described by a matrix $\mathbf{J}$ of dimension $N_J\times T$. The elements of $\mathbf{J}$, i.e., $\left[\mathbf{J}\right]_{n_J,t},n_J=1,2,\cdots,N_J,t=1,2,\cdots T$, denotes the interference signal transmitted from n_J th interference source at time periods *t*. The elements of $\mathbf{J}$ are assumed to be independent to each other and Gaussian distributed according to $CN(0,\ 1)$.

Similarly, with the assumptions that the transmission power are allocated equally across all the transmit antennas, and that pulse shaping, match-filtering, and sampling are all optimal, the received signal can be expressed as

$$\mathbf{Y}=\sqrt{\frac{\rho}{N_T}}\mathbf{HC(s)}+\mathbf{G\Phi J}+\mathbf{N} . \qquad (3)$$

3 Conventional ML Decoding Scheme

In this section, conventional ML decoding scheme for MIMO-OSTBC system in both interference-free case and interference case are reviewed respectively.

In interference-free case, assume the receiver knows the channel matrix $\mathbf{H}$, conventional ML decoding scheme for Eq.(2) is given by the following expression

$$\hat{\mathbf{s}}_{MLD}=\arg\min_{\mathbf{s}}\left\|\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}\left(\mathbf{s}\right)\right\|^2 , \qquad (4)$$

where $\left\|\bullet\right\|^2$ denotes the Euclidean norm of a vector or the Frobenius form of a matrix.

In interference case, decoding directly using Eq.(4) will not work correctly. The impact of interferences on the ML decoding scheme should be considered. Assume the receiver knows the channel matrix $\mathbf{H}$, since the interference and noise are both

Gaussian, according to [15], the conventional ML decoding scheme for Eq.(3) is given by

$$\hat{\mathbf{s}}_{MLD} = \arg\min_{\mathbf{s}} Tr\left[\left(\mathbf{Y} - \sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)^H \left(\mathbf{R}_{JN}\right)^{-1}\left(\mathbf{Y} - \sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)\right] \quad (5)$$

or equivalently

$$\hat{\mathbf{s}}_{MLD} = \arg\min_{\mathbf{s}} \left\|\left(\mathbf{R}_{JN}\right)^{-1/2}\left(\mathbf{Y} - \sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)\right\|^2, \quad (6)$$

where the matrix $\mathbf{R}_{JN}$ of dimension $N_R \times N_R$ is the covariance matrix for interferences plus noise, which is defined as:

$$\mathbf{R}_{JN} = E\left\{\left(\mathbf{G\Phi J} + \mathbf{N}\right)\left(\mathbf{G\Phi J} + \mathbf{N}\right)^H / T\right\}. \quad (7)$$

It is well known that conventional ML decoding scheme has optimum SER performance. However, from Eq.(4)~(6), it can be seen that conventional ML decoding scheme has high complexity of searching calculations since the symbols are detected jointly. The number of searching calculations is up to C^L, where C is the number of points in constellation $\mathcal{S}$, and L is the number of symbols in a block needs to be detected.

The searching calculations for some set of C and L are listed in Table 1. From Table 1, we can see, for large values of C and L, the searching calculations of conventional ML decoding scheme is much high. Therefore, simplified methods are needed.

Table 1 Searching Calculations of Conventional ML Decoding Scheme

Searching Calculations		L		
		2	4	8
C	4	4^2	4^4	4^8
	8	8^2	8^4	8^8
	16	16^2	16^4	16^8

4 Simplified ML Decoding Scheme

In this section, by simple linear processing, a simplified ML decoding scheme which supports any form of OSTBC for interference suppression is derived. The simplified ML decoding scheme for interference cancellation presented in [12] which only supports MIMO system with Alamouti coding scheme can be viewed as a special case of the simplified MLD scheme derived in this section.

Let's begin with the OSTBC coding matrix $\mathbf{C}(\mathbf{s})$. Generally, $\mathbf{C}(\mathbf{s})$ has the form of

$$\mathbf{C}(\mathbf{s}) = \sum_{l=1}^{L}\left(\mathbf{E}_l s_l + \mathbf{F}_l s_l^*\right) \quad (8)$$

for complex-valued constellation $\mathcal{S}$ or

$$\mathbf{C} = \sum_{l=1}^{L}\mathbf{E}_l s_l \quad (9)$$

for real-valued constellation $\mathcal{S}$, where $\mathbf{E}_l$ and $\mathbf{F}_l$ are $N_T \times T$ matrices, $l = 1,2,\ldots,L$.

Take Alamouti coding scheme [6] as an example. When the symbols s_1, s_2 are taken from a complex-valued constellation $\mathcal{S}$, the coding matrix $\mathbf{C}(\mathbf{s})$ is given by

$$\mathbf{C}(\mathbf{s}) = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix}.$$

Then, the corresponding matrices $\mathbf{E}_l$ and $\mathbf{F}_l$ are given respectively as follows:

$$\mathbf{E}_1 = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix} \mathbf{E}_2 = \begin{bmatrix} 0 & 0 \\ 1 & 0 \end{bmatrix}$$
$$\mathbf{F}_1 = \begin{bmatrix} 0 & 0 \\ 0 & 1 \end{bmatrix} \mathbf{F}_2 = \begin{bmatrix} 0 & -1 \\ 0 & 0 \end{bmatrix}.$$

When the symbols s_1, s_2 are taken from a real-valued constellation $\mathcal{S}$, the coding matrix $\mathbf{C}(\mathbf{s})$ is given by

$$\mathbf{C}(\mathbf{s}) = \begin{bmatrix} s_1 & -s_2 \\ s_2 & s_1 \end{bmatrix}.$$

Then, the corresponding matrices $\mathbf{E}_l$ are given by

$$\mathbf{E}_1 = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \mathbf{E}_2 = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}.$$

Note any OSTBC codeword satisfies the orthogonality described by Eq.(1) [4], therefore, substituting Eq.(8) and (9) into Eq.(1), respectively, we have

$$\mathbf{E}_l\mathbf{E}_l^H + \mathbf{F}_l\mathbf{F}_l^H = \alpha\mathbf{I}_{N_T} \qquad l = 1,2,\ldots,L$$
$$\begin{cases} \mathbf{E}_k\mathbf{E}_l^H = \mathbf{O} \\ \mathbf{F}_k\mathbf{F}_l^H = \mathbf{O} \end{cases} \qquad l,k = 1,2,\ldots,L \quad l \neq k$$
$$\mathbf{E}_k\mathbf{F}_l^H = \mathbf{O} \qquad l,k = 1,2,\ldots,L \quad (10)$$

for complex-valued constellation $\mathcal{S}$ or

$$\begin{cases} \mathbf{E}_l\mathbf{E}_l^H = \alpha\mathbf{I}_{N_T} & l = 1,2,\ldots,L \\ \mathbf{E}_k\mathbf{E}_l^H = \mathbf{O} & l,k = 1,2,\ldots,L \quad l \neq k \end{cases} \quad (11)$$

for real-valued constellation $\mathcal{S}$.

Define a L-dimensional vector $\tilde{\mathbf{s}}$ as follows:

$$\tilde{\mathbf{s}} = [\tilde{s}_1, \tilde{s}_2, \cdots, \tilde{s}_L]^T. \quad (12)$$

Using the orthogonal property Eq.(1) and Eq.(10) ~(12), the l th elements of $\tilde{\mathbf{s}}$ is obtained via

$$\tilde{s}_l = \frac{Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{Y}\mathbf{E}_l^H + \mathbf{H}^T\left(\left(\mathbf{R}_{JN}\right)^{-1}\right)^*\mathbf{Y}^*\mathbf{F}_l^T\right)}{\left(\alpha\sqrt{\frac{\rho}{N_T}}\cdot Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)\right)} \quad (13)$$

for complex-valued constellation $\mathcal{S}$ or

$$\tilde{s}_l = \frac{Tr\left(\mathrm{Re}\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{Y}\mathbf{E}_l^H\right)\right)}{\left(\alpha\sqrt{\frac{\rho}{N_T}}\cdot Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)\right)} \quad (14)$$

for real-valued constellation $\mathcal{S}$, where $Tr(\bullet)$ denotes the trace of a matrix, $(\bullet)^*$ denotes the conjugation operation.

Through derivation, we can prove that the conventional ML decoding scheme Eq.(5) in interference case is equivalent to the following expression

$$\hat{\mathbf{s}}_{MLD} = \arg\min_{\mathbf{s}\in S^L}\left\|\mathbf{s}-\tilde{\mathbf{s}}\right\|^2 . \quad (15)$$

Eq.(15) can be rewritten as:

$$\hat{s}_{MLD,l} = \arg\min_{s_l\in S}\left|s_l-\tilde{s}_l\right|^2 \qquad l=1,2,\ldots,L . \quad (16)$$

Let's begin with the likelihood function. When the transmitted symbols are selected from a real-valued constellation $\mathcal{S}$, according to the conventional ML decoding scheme Eq.(5) in interference case, the likelihood function can be rewritten as :

$$\begin{aligned}\xi(\mathbf{s}) &= Tr\left[\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)^H\left(\mathbf{R}_{JN}\right)^{-1}\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)\right]\\
&= Tr\left[\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)^H\left(\mathbf{R}_{JN}\right)^{-1}\right]\\
&= Tr\left(\mathbf{Y}\mathbf{Y}^H\left(\mathbf{R}_{JN}\right)^{-1}\right) - Tr\left(\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\mathbf{Y}^H\left(\mathbf{R}_{JN}\right)^{-1}\right)\\
&\quad - Tr\left(\sqrt{\frac{\rho}{N_T}}\mathbf{YC}(\mathbf{s})^H\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\right)\\
&\quad + Tr\left(\frac{\rho}{N_T}\mathbf{HC}(\mathbf{s})\mathbf{C}(\mathbf{s})^H\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\right).\end{aligned} \quad (17)$$

Substituting Eq.(9) and (1) into Eq.(14), and utilizing Eq.(11), Eq.(17) can be rewritten as:

$$\begin{aligned}\xi(\mathbf{s}) = \alpha\frac{\rho}{N_T}Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)\\
\bullet\sum_{l=1}^{L}\left(s_l - \frac{Tr\left(\mathrm{Re}\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{Y}\mathbf{E}_l^H\right)\right)}{\alpha\sqrt{\frac{\rho}{N_T}}Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)}\right)^2\\
+Tr\left(\mathbf{Y}\mathbf{Y}^H\left(\mathbf{R}_{JN}\right)^{-1}\right)\\
-\frac{\sum_{l=1}^{L}\left(Tr\left(\mathrm{Re}\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{Y}\mathbf{E}_l^H\right)\right)\right)^2}{\alpha Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)} .\end{aligned} \quad (18)$$

Therefore, substituting Eq.(18) and (13) into Eq.(5), we have

$$\begin{aligned}\hat{\mathbf{s}}_{MLD} &= \arg\min_{\mathbf{s}\in S^L}\xi(\mathbf{s})\\
&= \arg\min_{\mathbf{s}\in S^L} Tr\left(\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)^H\left(\mathbf{R}_{JN}\right)^{-1}\left(\mathbf{Y}-\sqrt{\frac{\rho}{N_T}}\mathbf{HC}(\mathbf{s})\right)\right)\\
&= \arg\min_{\mathbf{s}\in S^L}\sum_{l=1}^{L}\left(s_l - \frac{Tr\left(\mathrm{Re}\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{Y}\mathbf{E}_l^H\right)\right)}{\alpha\sqrt{\frac{\rho}{N_T}}Tr\left(\mathbf{H}^H\left(\mathbf{R}_{JN}\right)^{-1}\mathbf{H}\right)}\right)^2\\
&= \arg\min_{\mathbf{s}\in S^L}\sum_{l=1}^{L}\left(s_l-\tilde{s}_l\right)^2\\
&= \arg\min_{\mathbf{s}\in S^L}\left\|\mathbf{s}-\tilde{\mathbf{s}}\right\|^2 .\end{aligned} \quad (19)$$

Similarly, when the transmitted symbols are selected from a complex-valued constellation $\mathcal{S}$, we have

$$\begin{aligned}\hat{\mathbf{s}}_{MLD} &= \arg\min_{\mathbf{s}\in S^L}\xi(\mathbf{s})\\
&= \arg\min_{\mathbf{s}\in S^L}\sum_{l=1}^{L}\left|s_l-\tilde{s}_l\right|^2 .\\
&= \arg\min_{\mathbf{s}\in S^L}\left\|\mathbf{s}-\tilde{\mathbf{s}}\right\|^2\end{aligned}$$

Through the above derivation, Eq.(12) has been proved to be equivalent to Eq.(5).

According to the simplified ML decoding scheme Eq.(13)~(16), once $\tilde{\mathbf{s}}$ is obtained via linear processing, the elements of $\mathbf{s}$ can be detected respectively, and the corresponding searching calculations is reduced to $C\times L$. Therefore, for large values of C and L, the searching calculations of the simplified ML decoding scheme Eq.(13)~(16) is much more less than that of the conventional ML decoding scheme Eq.(5)~(6). Table 2 lists the searching calculations for some set of C and L.

Table 2 Searching Calculations of Simplified ML Decoding Scheme

Searching	L

Calculations		2	4	8
C	4	8	16	32
	8	16	32	64
	16	32	64	128

5 Proposed Interference Suppression Scheme

The channel matrix $\mathbf{H}$ is assumed to be known at the receiver in both the conventional ML decoding scheme Eq.(5)~(6) and the simplified ML decoding scheme Eq.(13)~(16). In practical applications, channel estimation techniques should be used to help the receiver obtain the estimation of channel matrix $\mathbf{H}$ and recover the transmitted symbols. However, the conventional channel estimation techniques [14-15] are assumed to work in the interference-free case and will suffer a great performance loss in the strong interference case. Moreover, matrix-inversion computation is necessary in order to obtain $\left(\mathbf{R}_{JN}\right)^{-1}$ in the simplified ML decoding scheme Eq.(13)~(16). Therefore, the simplified ML decoding scheme Eq.(13)~(16) needs to be further improved

In this section, for strong interference, we present an interference suppression scheme based on simplified ML decoding and noise subspace projection. And how to obtain the channel information required and how to avoid computing$\left(\mathbf{R}_{JN}\right)^{-1}$ in the scheme at the receiver are presented.

5.1 Interference Suppression Scheme

In order to avoid computing$\left(\mathbf{R}_{JN}\right)^{-1}$, we start with the eigen decomposition of $\mathbf{R}_{JN}$. Note that the $\mathbf{R}_{JN}$ can be decomposed as:

$$\mathbf{R}_{JN} = \left(\mathbf{U}_J, \mathbf{U}_N\right) \times diag(\delta_1, \delta_2, \cdots, \delta_{N_J}, 1, \cdots, 1) \begin{pmatrix} \mathbf{U}_J^H \\ \mathbf{U}_N^H \end{pmatrix}$$

$$= \mathbf{U}_J diag(\delta_1, \delta_2, \cdots, \delta_{N_J}) \mathbf{U}_J^H + \mathbf{U}_N \mathbf{U}_N^H, \quad (20)$$

where $\delta_1 \geq \delta_2 \cdots \geq \delta_{N_J} \geq 1$ are the eigen values of interferences subspace, the columns of $\mathbf{U}_J$ are basis vectors of interferences subspace and the columns of $\mathbf{U}_N$ are basis vectors of noise subspace. Therefore, from Eq.(20), it is not difficult to obtain the form of $\left(\mathbf{R}_{JN}\right)^{-1}$

$$\left(\mathbf{R}_{JN}\right)^{-1} = \mathbf{U}_J diag(\delta_1^{-1}, \delta_2^{-1}, \cdots, \delta_{N_J}^{-1}) \mathbf{U}_J^H + \mathbf{U}_N \mathbf{U}_N^H . \quad (21)$$

In strong interference case, the power of interference is much stronger than that of the noise, i.e. $\delta_{n_J} >> 1$, for $n_J = 1, 2, \cdots N_J$. Thus, the value of $\left(\mathbf{R}_{JN}\right)^{-1}$ is close to

$$\left(\mathbf{R}_{JN}\right)^{-1} = \mathbf{U}_N \mathbf{U}_N^H , \quad (22)$$

then Eq.(5) can be rewritten as:

$$\hat{\mathbf{s}}_{MLD} = \arg\min_{\mathbf{s}} \left\| \left(\mathbf{U}_N^H \mathbf{Y}\right) - \sqrt{\frac{\rho}{N_T}} \left(\mathbf{U}_N^H \mathbf{H}\right) \mathbf{C}(\mathbf{s}) \right\|^2 . \quad (23)$$

Comparing Eq.(23) with Eq. (4), we obtain Eq.(23) from Eq.(4) by replacing $\mathbf{Y}$ and $\mathbf{H}$ with $\mathbf{U}_N^H \mathbf{Y}$ and $\mathbf{U}_N^H \mathbf{H}$ respectively. As the columns of $\mathbf{U}_N$ are the basis vectors of noise subspace, Eq.(23) can be viewed as a noise-subspace-projected version of Eq.(4).

In the following of the paper, we call

$$\mathbf{Y}_{PRO} = \mathbf{U}_N^H \mathbf{Y} \quad (24)$$

noise-subspace-projected receive signal matrix, and call

$$\mathbf{H}_{PRO} = \mathbf{U}_N^H \mathbf{H} \quad (25)$$

noise-subspace-projected channel matrix. Therefore, Eq.(23) can be rewritten as

$$\hat{\mathbf{s}}_{MLD} = \arg\min_{\mathbf{s}} \left\| \mathbf{Y}_{PRO} - \sqrt{\frac{\rho}{N_T}} \mathbf{H}_{PRO} \mathbf{C}(\mathbf{s}) \right\|^2 . \quad (26)$$

Utilizing Eq.(22), (24) and (25), the ML decoding scheme Eq.(26) can be simplified as

$$\hat{\mathbf{s}}_{PRO} = \arg\min_{\mathbf{s} \in S^L} \left\| \mathbf{s} - \mathbf{s}' \right\|^2 \quad (27)$$

or equivalently

$$\hat{s}_{PRO,l} = \arg\min_{s_l \in S} \left| s_l - s_l' \right|^2 \qquad l = 1, 2, \cdots, L \quad (28)$$

where $\mathbf{s}' = [s_1', s_2', \cdots, s_L']^T$, and the l th elements of $\mathbf{s}'$ is obtained via

$$\mathbf{s}_l' = \frac{Tr\left[\left(\mathbf{H}_{PRO}{}^H \mathbf{Y}_{PRO}\right) \mathbf{E}_l^H + \left(\mathbf{H}_{PRO}{}^H \mathbf{Y}_{PRO}\right)^* \mathbf{F}_l^T\right]}{\alpha \sqrt{\frac{\rho}{N_T}} \left\| \mathbf{H}_{PRO} \right\|^2} \quad (29)$$

for complex-valued constellation S or

$$\mathbf{s}_l' = \frac{Tr\left[\operatorname{Re}\left\{\left(\mathbf{H}_{PRO}{}^H \mathbf{Y}_{PRO}\right) \mathbf{E}_l^H\right\}\right]}{\alpha \sqrt{\frac{\rho}{N_T}} \left\| \mathbf{H}_{PRO} \right\|^2} \quad (30)$$

for real-valued constellation S.

We call the scheme Eq.(27)~(30) interference suppression scheme based on simplified ML decoding and noise subspace projection.

5.2 Performance Analysis

Here, we discuss the performance of the interference suppression scheme Eq.(27)~(30) based on simplified ML decoding and noise subspace projection in strong interference case.

We assume that the MIMO channels between the transmitter and receiver are i.i.d Rayleigh channels. Also we assume that the number of receive antennas is larger than that of interferences, i.e., $N_R > N_J$, and that the channel matrix $\mathbf{G}$ between interferences and the receiver is full column rank.

Obviously, the elements of noise-subspace-projected channel matrix $\mathbf{H}_{PRO}$ have the i.i.d complex Gaussian distribution with zero mean and unit variance, i.e., $\left[\mathbf{H}_{PRO}\right]_{nn_T} \sim CN(0,\ 1)$, for $n_T = 1,\ 2,...,N_T$ and $n = 1,\ 2,\ ...,\ N_R - N_J$.

And according to Eq.(24), we have

$$\mathbf{Y}_{PRO} = \sqrt{\frac{\rho}{N_T}}\mathbf{H}_{PRO}\mathbf{C}(\mathbf{s}) + \mathbf{N}_{PRO}\ , \qquad (31)$$

where $\mathbf{N}_{PRO} = \mathbf{U}_{\mathbf{N}}^{H}\mathbf{N}$, and the elements of matrix $\mathbf{N}_{PRO}$ have the i.i.d complex Gaussian distribution with zero mean and unit variance.

Thus, Eq.(31) can be also used to describe a MIMO system with N_T Tx and $\left(N_R - N_J\right)$ Rx in interference-free case, where the i.i.d Rayleigh channels is described by a matrix $\mathbf{H}_{PRO}$. And the receiver of the N_T-by-$\left(N_R - N_J\right)$ MIMO system performs the ML decoding according to Eq.(26).

Here, two MIMO systems System Ω_1 and System Ω_2 are considered. System Ω_1 is an N_T-by-N_R MIMO system with N_J strong interferences, the receiver perform decoding according to the interference suppression scheme based on simplified ML decoding and noise subspace projection Eq.(27)~(30) ; and System Ω_2 is an N_T-by-$\left(N_R - N_J\right)$ MIMO system in interference-free case and perform ML decoding according to Eq.(26). Since the decoding scheme Eq.(26) is equivalent to the decoding scheme Eq.(27)~(30), With the assumption given firstly in this subsection, we have the following conclusion that System Ω_1 and System Ω_2 have the same SER performance.

5.3 Schemes Comparison

Compared with the simplified ML decoding scheme Eq.(13)~(16), the proposed interference suppression scheme based on simplified ML decoding and noise subspace projection scheme Eq.(27)~(30) has two advantages.

First, the proposed scheme Eq.(27)~(30) avoid matrix-inversion computation by using noise subspace projection, since in strong interference case, the value of $\left(\mathbf{R}_{JN}\right)^{-1}$ is close to that of $\mathbf{U}_N\mathbf{U}_N^H$.

Second, in this scheme, from the similarity between Eq.(26) and Eq.(4), it is not difficult to find that the value of the noise-subspace-projected channel matrix $\mathbf{H}_{PRO}$ can be estimated with conventional channel estimation techniques using noise-subspace-projected receive signal matrix $\tilde{\mathbf{Y}}$ and pilot symbols.

It is obviously that in strong interference case, these two schemes have the same SER performance. When the interference is not strong enough to satisfy Eq.(22), the SER performance of the proposed scheme Eq.(27)~(30) is a little worse than that of the scheme Eq.(13)~(16).

5.4 The steps of interference suppression scheme

Finally, we conclude the steps of our interference suppression scheme based on simplified ML decoding and noise subspace projection scheme:

(1)Transmit several zero symbols at the transmitter, i.e., $\mathbf{C} = \mathbf{O}$;

(2)By using eigen decomposition or Projection Approximation Subspace Tracking (PAST) algorithm [16], obtain $\mathbf{U}_N$ from received signals including noise and interference only;

(3)Using conventional channel estimation techniques, estimate the noise-subspace-projected channel $\mathbf{H}_{PRO}$ from the designed pilot symbols and noise-subspace-projected receive signal matrix $\mathbf{Y}_{PRO}$;

(4) Obtain the ML decoding results according to the scheme (27) ~ (30).

6 Simulation Results

In this section, the simulation results of the proposed interference suppression scheme based on simplified ML decoding and noise subspace projection are presented.

We carry out four experiments, Ex1, Ex2, Ex3, and Ex4. For Ex1, Ex2, Ex3, and Ex4, uniform linear array (ULA) with three antennas is employed at the receiver, the antenna space at the receiver is half the wavelength, the symbols are taken from 16QAM constellation set, and the MIMO channels are supposed to be Rayleigh channels and to be invariant during each frame of 300 symbols but vary between different frames. For Ex1, Ex2 and Ex3, we assume that the channel matrix $\mathbf{H}$ and the matrix $\left(\mathbf{R}_{JN}\right)^{-1}$ are known at the receiver, and the space-time code used is as follows:

$$\mathbf{C}(\mathbf{s})=\begin{bmatrix} s_1 & -s_2^* & \beta s_3^* & \beta s_3^* \\ s_2 & s_1^* & \beta s_3^* & -\beta s_3^* \\ \beta s_3 & \beta s_3 & \gamma(-s_1-s_1^*+s_2-s_2^*) & \gamma(s_1-s_1^*+s_2+s_2^*) \end{bmatrix}$$

$$(\beta=\sqrt{1/2},\gamma=1/2).$$

The purpose of Ex1 is to verify the interference suppression ability of the proposed scheme. There are three different cases in Ex1. The first case is that of non-interference. The second case is that there is only one interference from direction of $\theta_1=-20°$, INR=30dB, K=6dB. And the third case is that there are two interferences from $\theta_1=-20°$ and $\theta_2=40°$, the corresponding interference to noise ratios and K factors are INR1=INR2=30dB, $K$1=$K$2=6dB respectively. The SER curves in these cases without interference suppression and with the proposed interference suppression scheme are shown in Fig. 2.

From Fig.2, compared with the case of non-interference, it can be easily seen that the SER performance of the system without interference suppression degrades significantly for one interference and two interferences cases. And with the proposed interference suppression scheme, the SER performance is greatly improved. It is indicated that the proposed scheme is valid to suppress interference for MIMO-OSTBC system although with some performance loss compared with the ideal case of non-interference.

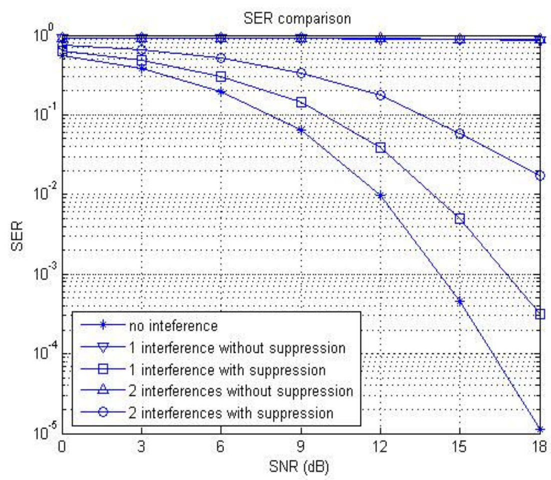

Fig.2 SER comparison
(Without interference suppression and with interference suppression)

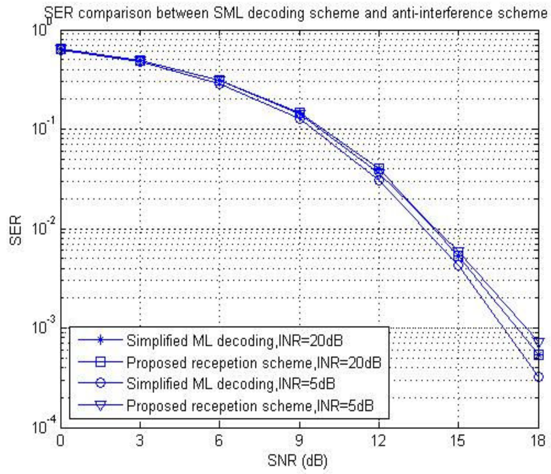

Fig.3 SER comparison of different Schemes
(Different INR)

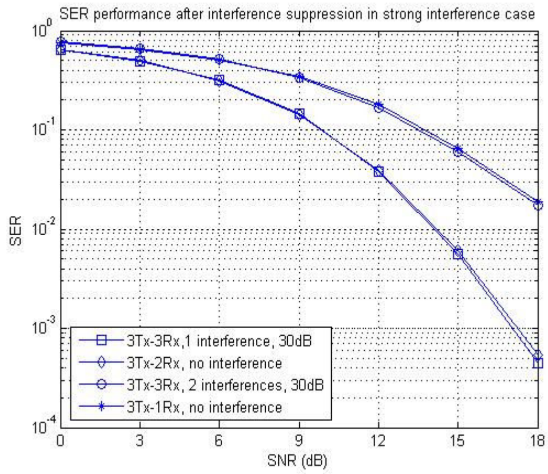

Fig.4 SER performance with interference suppression
(In strong interference case)

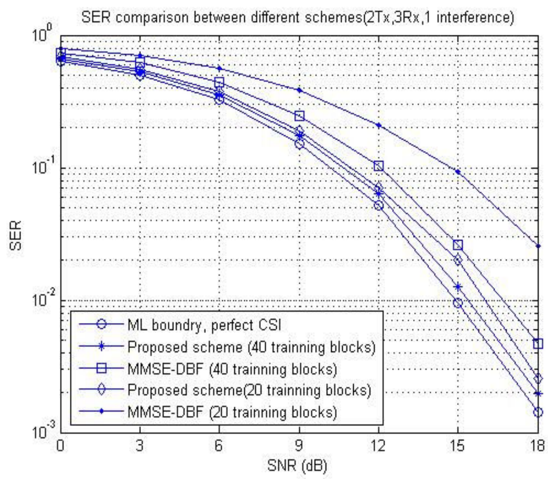

Fig.5 SER comparison of different Schemes
(Different training sequence length)

In Ex2, we compare the SER performances of the proposed interference suppression scheme and the simplified ML decoding scheme. There is only one interference from direction of $\theta_1 = -20°$, the INR is set as 20dB and 5dB, respectively, the K factor is fixed at K=6dB. The corresponding result is shown in Fig.3.

As shown in Fig.3, it is obviously that when INR=20dB, i.e., the interference is strong enough, the proposed interference suppression scheme and the simplified ML decoding scheme have the same SER performance; when INR=5dB, i.e., the interference is not strong enough, the SER performance of the proposed scheme is a little worse than that of the simplified ML decoding scheme. It accord with the result of scheme comparison in section 5.3.

In Ex3, we focus on the performance of MIMO-OSTBC system in strong interference case with the proposed interference suppression scheme. Three kinds of MIMO systems are simulated, which are 3-by-3 MIMO system with interferences, 3-by-2 MIMO system without interference and 3-by-1 MIMO system without interference, respectively. For 3-by-3 MIMO system with interference, the proposed scheme is used. There are two cases. The first case is that there is only one interference from direction of $\theta_1 = -20°$, INR=30dB, K=6dB. And the second case is that there are two interferences from $\theta_1 = -20°$ and $\theta_2 = 40°$, INR1=INR2=30dB, $K$1=$K$2=6dB. For the 3-by-2 and 3-by-1 MIMO systems without interference, the receiver performs ML decoding. The SER curves of these three kinds of systems are shown in Fig.4.

From Fig.4, we can see that in the case of one strong interference, the SER curve of the 3-by-3 MIMO system with the proposed scheme almost coincides with that of the 3-by-2 MIMO system without interference using ML decoding scheme. And similarly, in the case of two strong interferences, the SER curve of the 3-by-3 MIMO system with the proposed scheme almost coincides with that of the 3-by-1 MIMO system without interference using ML decoding scheme. The simulation results of Ex3 is consistent with the conclusion drawn in section 5.2 that an N_T-by-N_R MIMO system with N_J strong interferences performing decoding according to the proposed scheme and an N_T-by-$(N_R - N_J)$ MIMO system in interference-free case performing ML decoding have the same SER performance.

In Ex4, we compare the SER performance of proposed interference suppression scheme with that of the MMSE-DBF reception scheme proposed in [11] with Alamouti coding scheme [6]. In the simulation, we assume the channel information is not known at the receiver. Thus, for the proposed interference suppression scheme, we need to estimate the noise-subspace-projected channel $\mathbf{H}_{PRO}$ using pilot symbols. And for the MMSE-DBF reception scheme proposed in [11], pilot symbols are also needed to estimate the auto correlation matrix of the received signal and the cross correlation vector between the received signal and the desired signal to calculate the weight vector. There is only one interference from direction of $\theta_1 = -20°$, INR=30dB, K=6dB. The pilot symbols are also encoded into blocks of codewords. For the proposed scheme, the pilot symbols include the zero symbols and designed pilot symbols used in step 1 and 2 respectively, we assume the length of these two parts are equal. There are two kinds of total length of pilot symbols, 20 blocks and 40 blocks. The simulation results are shown in Fig. 5.

For comparison, the curve of ML decoding with perfect CSI is shown in Fig.5, which means the receiver knows the CSI. From Fig.5, we can see that the SER performance of the proposed interference suppression scheme is much better than that of the MMSE-DBF reception scheme in [11] with the same length of pilot symbols, and is more close to the curve of the ML decoding. When the length of pilot symbols is decreasing, the SER performances of both two schemes degrades, but the proposed scheme is less sensitive to the length of pilot symbols, which is more flexible in practical applications.

7 Conclusion

In this paper, an interference suppression scheme for MIMO-OSTBC system in strong Gaussian distributed interference scenario is presented. The proposed interference suppression scheme is based on simplified ML decoding and noise subspace projection. And the method using conventional channel estimation techniques to obtain the efficient estimation of the noise-subspace-projected channel $\mathbf{H}_{PRO}$ needed in the proposed scheme in strong interference scenario is also discussed. The theoretical analysis and the simulation show that the proposed scheme is valid to suppress interference for MIMO-OSTBC system. The proposed scheme can achieve ML decoding with simple linear processing, and can avoid matrix-inversion computation through noise subspace projection. The theoretical analysis and the simulation results are in good agreement. Compared with the MMSE-DBF reception scheme in [11], the performance of the proposed scheme is better and it is less sensitive to the length of pilot symbols. And in practical applications, the proposed reception scheme is more flexible which supports not only Alamouti coding scheme [6], but also other OSTBC. It represents a promising candidate for interference suppression for MIMO-OSTBC system.

References:

[1] I. E. Telatar, "Capacity of multi-antenna Gaussian channels", European Trans. On Telecomm, Vol.10: 585-595, Nov. /Dec. 1999.

[2] G.J. Foschini, M.J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas", Wireless Personal Communications. Vol. 6: 311-335, 1998.

[3] D. Gesbert, M. Shafi, D. S. Shiu, P. Smith, and A. Naguib, "From theory to practice: an overview of MIMO space-time coded wireless systems", IEEE J. on Selct Areas in Commun., Vol.21, No.3, pp. 281-302, Apr.2003.

[4] V. Tarokh, H. Jafarkhani, A. R. Calderbank. "Space-time block codes from orthogonal designs", IEEE Trans. Inf. Theory, vol. 45, pp. 1456-1467, 1999.

[5] V. Tarokh, N. Seshadri, A.R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction", IEEE Trans. on Inform. Theory, 44(2): 744-765, 1998.

[6] S. M. Alamouti, "A simple transmit diversity technique for wireless communications", IEEE Journal. Select. Areas Commun., Vol 16, pp. 1451-1458, 1998.

[7] 3GPP TR 25.869 (v1.2.0) "Tx Diversity Solutions for Multiple Antennas", August 2003, Rel6. http://www.3gpp.org

[8] WiMAX Forum. White paper: Fixed, nomadic, portable and mobile applications for 802.16-2004 and 802.16e WiMAX networks, http://www.wimaxforum.org

[9] A. Naguib, N. Seshadri and A. Calderbank, "Applications of space-time block codes and interference suppression for high capacity and high data rate wireless systems", Thirty-Second A silomar Conference on Signals, Systems & Computers,Vol.2, pp. 1803 - 1810, Nov. 1998.

[10] Xuan Nam Tran, T. Taniguchi and Y. Karasawa, "Adaptive beam forming for multiuser space-time block coded systems", 2004 IEEE Eighth International Symposium on Spread Spectrum Techniques and Applications, Sydney, Austrialia, pp. 745-749, 30Aug.-2 Sept. 2004.

[11] Xiaoxu Chen, Yu Gong and Yaohuan Gong, "Suppression of directional interference for STBC MIMO system based on beam-forming", ICCCAS2006, Guilin, China, pp. 983-987, 2006.

[12] Zhonghao Zhang, Yu Gong, Yuehuan Gong, Yalan Ye and Chen Jin, "A Simplified ML Approach in Suppression of Directional Interference for STBC MIMO System", WiCom2007, Shanghai, China, pp. 346-349, Sept.21-25, 2007

[13] Jie Wang, Zhonghao Zhang, Yuehuan Gong, and Xun Wang, "An MMSE-DBF interference suppression reception scheme for MIMO-OSTBCs system", WiCOM2008, Dalian, China, p4678011, Oct12-14, 2008

[14] S. M. Kay, "Fundamentals of Statistical Signal Processing: Estimation Theory", Prentice-Hall, Inc, 1993

[15] H. Arslan and G. E. Bottomley, "Channel estimation in narrowband wireless communication", Wireless Communications and Mobile Computing, 1:201–219, 2001.

[16] B. Yang, "Projection Approximation Subspace Tracking," IEEE Trans. on Signal Proc., vol. 44, no. 1, Jan 1995.