Pilot Symbol Assisted Hybrid Detection for OFDM-Based Spatial Multiplexing Systems

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In this paper, we provide a new detection scheme for a pilot symbol assisted interference nulling and cancellation operation to reduce unexpected effects owing to parallel transmission in orthogonal frequency division multiplexing (OFDM)-based spatial multiplexing systems. We have shown that the investigated OFDM vertical Bell laboratories layered space time (VBLAST) detection based on hybrid processing performs better than ordinary OFDM-VBLAST detections based on serial processing and parallel processing, respectively.

Keywords: OFDM, VBLAST, nulling, cancellation, parallel transmission.

I. Introduction

The goal of future wireless communication is to provide high quality wireless multimedia services; thus, high data rate communication must be transmitted reliably. Recent theoretic research has shown that multiple transmitting and receiving antenna systems can achieve enormous spectral efficiency. Various multiple-input multiple-output (MIMO) systems have been proposed to attain this available capacity [1], [2].

Orthogonal frequency division multiplexing (OFDM) is one of the most attractive multicarrier modulation schemes for high bandwidth efficiency and strong immunity to multi-path fading. Thus, multiple transmitting and receiving antennas can be used with OFDM to improve the communication capacity and quality of mobile wireless systems [3]-[6]. Among these algorithms, the OFDM-based Bell laboratories layered space-time (BLAST), with a number of transmitting and receiving antennas, increases the transmission rate efficiently using low multiplication operations [6]. Such algorithms could provide very high data-rate communication over wireless channels without increasing the total transmission power and bandwidth. Based on its knowledge of the matrix of propagation coefficients, the vertical-BLAST (VBLAST) detection scheme is often referred to as sequential nulling and cancellation. Recently, various detection methods for improving VBLAST systems have been proposed [7]-[9].

In this paper, we develop an efficient hybrid detection algorithm for the nulling vector and cancellation in VBLAST systems. The simulation results describe the effect of the pilot symbol assisted interference cancellation and hybrid detection in OFDM-based VBLAST systems, and the performance of the hybrid VBLAST detection algorithm is superior to that of

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ordinary VBLAST detection algorithms based on serial detection [2] and parallel detection [7].

This paper is organized as follows. In section II, the system description is presented, section III and IV explain pilot preamble structure and pilot symbol assisted channel separation, respectively. In sections V and VI, hybrid VBLAST detection and a modified nulling of the hybrid VBLAST detection algorithm are derived briefly. Finally, in section VII, simulation results and discussions are described.

II. System Description

We consider OFDM-based VBLAST systems with N_t transmitting and N_r receiving antennas. The OFDM data sequence of the n-th transmitting antenna is represented as $\mathbf{X}_n = [X_n(0)\, X_n(1) \cdots X_n(K-1)]^T$, where $X_n(l), n = 1, 2, \cdots, N_t$ is the complex baseband signal of the l-th subcarrier, and K is the length of the OFDM data sequence. Because the symbols are transmitted from N_t transmitting antennas in parallel, the $N_t \times K$ data sequence matrix is $\mathbf{X} = [\mathbf{X}_1 \mathbf{X}_2 \cdots \mathbf{X}_{N_t}]^T$.

A flat fading channel on each subcarrier and independent identically distributed fading among different subcarriers are assumed in our analysis. Also, we assume that the time variation of fading is negligible over a frame, which is estimated using the pilot preamble. The time delay and phase offset of each antenna are assumed to be known, i.e., tracked accurately. Therefore, the overall channel, **H,** can be represented as an $N_r \times N_t$ complex matrix, and the baseband received signal at the j-th receiving antenna is

$$r_j(l) = \sum_{i=1}^{N_t} h_{ji} X_i(l) + v_j(l), \tag{1}$$

where h_{ji} is the channel element with the i-th transmitting antenna and j-th receiving antenna and $v_j(l)$ is zero-mean Gaussian noise with variance σ_v^2. The overall received signals can be represented as

$$\mathbf{R} = \mathbf{HX} + \mathbf{V}. \tag{2}$$

where **R** and **V** are the $N_t \times K$ matrices, respectively.

III. Preamble Structure Using Pilot Symbol

We consider an OFDM system with K subcarriers that uses N_t transmitting and N_r receiving antennas. If we denote $X_i(k)$ for frequency-domain index $0 \le k \le K$ as a preamble pattern for the i-th transmitting antenna for $1 \le i \le N_t$, $x_1(n)$ is then the inverse fast Fourier transform output of the preamble

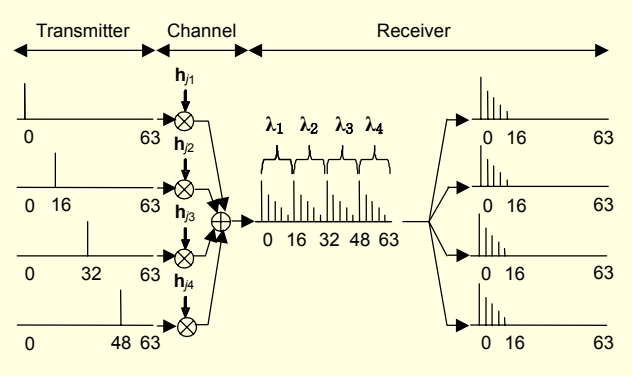

Fig. 1. Process of pilot symbol assisted channel separation in the case of $K = 64$ and $L = 16$.

pattern for the first transmitting antenna denoted as $X_1(k)$. The relationship is derived as the following rule:

$$X_i(k) = \mathrm{FFT}[x_1(n - n_i)], \tag{3}$$

where n is the time-domain index $(0, 1, \cdots, K-1)$ of an OFDM symbol and n_i is any integer for time shifting in the i-th preamble pattern. If we denote $x_1(n-n_i)$ as $x_i(n)$, the vector of preamble pattern $\mathbf{x}_i$ for the i-th transmitting antenna is presented as follows:

$$\begin{aligned} \mathbf{x}_i &= [x_i(0)\; x_i(1) \cdots x_i(K-1)]^T \\ &= [\underbrace{0 \cdots 0}_{n_i} 1 0 \cdots 0]^T \end{aligned}. \tag{4}$$

For a given number of substantial subcarriers used in data modulation denoted by K and a channel response with a length of L, the maximum number of transmitting antennas N_t can be given by

$$N_t = \lfloor K / L \rfloor, \tag{5}$$

where $\lfloor a \rfloor$ gives the largest integer not exceeding a, and L is the maximum number of channel paths between each pair of transmitting and receiving antennas. Considering (5), the time shift index for each N_t multiple transmitting antennas can be designed as

$$n_i = (i-1)\lfloor K / N_t \rfloor \quad \text{for} \quad 1 \le i \le N_t. \tag{6}$$

When a design criterion of $\lfloor K / N_t \rfloor \ge L$ is satisfied, orthogonality is still maintained even if encountered with a channel impulse response with a length of L, which is illustrated in Fig. 1. In this example, the number of transmitting antennas, $N_t = 4$ from (5), and $\mathbf{x}_i$'s for $1 \le i \le N_t$ are shown in Fig. 1. In the case of four transmitting antennas, from (3), the

preamble pattern in the frequency domain can be expressed as

$$\begin{aligned}\mathbf{X}_1 &= [1 \quad 1 \quad 1 \quad 1 \cdots 1 \quad 1 \quad 1 \quad 1]^T,\\ \mathbf{X}_2 &= [1 \quad -j \quad -1 \quad j \quad \cdots \quad 1 \quad -j \quad -1 \quad j]^T,\\ \mathbf{X}_3 &= [1 \quad -1 \quad 1 \quad -1 \quad \cdots \quad 1 \quad -1 \quad 1 \quad -1]^T,\\ \mathbf{X}_4 &= [1 \quad j \quad -1 \quad -j \quad \cdots \quad 1 \quad j \quad -1 \quad -j]^T. \end{aligned} \tag{7}$$

IV. Preamble Structure Using Pilot Symbol

For each transmitting antenna *i*, we assume that a transmitted OFDM symbol goes through a multi-path channel before reaching the *j*-th receiving antenna with the channel impulse response modeled by

$$\mathbf{h}_{ji} = [h_{ji}(0)\ h_{ji}(1) \cdots h_{ji}(L-1)]^T. \tag{8}$$

Then, the received time-domain samples of an OFDM symbol from receiver *j* can be expressed as

$$\mathbf{r}_j = \sum_{i=1}^{N_t} \mathbf{x}_i \otimes \mathbf{h}_{ji} + \mathbf{v}_j = \sum_{i=1}^{N_t} \mathbf{g}_{ji} + \mathbf{v}_j, \tag{9}$$

where $\otimes$ represents *K*-point circular convolution, $\mathbf{v}_j = [v_j(0)v_j(1)\cdots v_j(K-1)]^T$ are independent identically distributed additive white Gaussian noise (AWGN) samples with zero mean and a variance of σ_t^2, and $\mathbf{g}_{ji}$ is the n_i-th cyclic shifted version of $\mathbf{h}_{ji}$, which is given by

$$\mathbf{g}_{ji} = [\underbrace{0 \cdots 0}_{n_i}\ h_{ji}(0)\ h_{ji}(1) \cdots h_{ji}(L-1)\ 0 \cdots 0]^T. \tag{10}$$

As shown in Fig. 1, multiplying the received signal $\mathbf{r}_j$ at the *j*-th receiving antenna with a rectangular window vector, which is defined as $\boldsymbol{\lambda}$, yields

$$\mathbf{y}_j = \boldsymbol{\lambda}\, \mathbf{r}_j = \mathbf{g}_j + \hat{\mathbf{v}}_j, \tag{11}$$

where $\boldsymbol{\lambda} = [\boldsymbol{\lambda}_1\, \boldsymbol{\lambda}_2 \cdots \boldsymbol{\lambda}_{N_t}]^T$ with each element of a $K \times K$ diagonal matrix is

$$\boldsymbol{\lambda}_i = diag[\underbrace{0 \cdots 0}_{n_i} \underbrace{1 1 \cdots 1}_{\lfloor K/N_t \rfloor} 0 \cdots 0], \tag{12}$$

$\mathbf{g}_j = [\mathbf{g}_{j1}\mathbf{g}_{j2} \cdots \mathbf{g}_{jN_t}]$, and $\hat{\mathbf{v}}_j = [\mathbf{v}_{j1}\mathbf{v}_{j2} \cdots \mathbf{v}_{jN_t}]$ with each component expressed as

$$\mathbf{v}_{ji} = [\underbrace{0 \cdots 0}_{n_i}\ \underbrace{v_{ji}(n_i) \cdots v_{ji}(n_i + \lfloor K/N_t \rfloor - 1)}_{\lfloor K/N_t \rfloor}\ 0 \cdots 0]^T. \tag{13}$$

Then, as shown in Fig. 1, it removes the effect of $\mathbf{x}_i$ by shifting $\mathbf{g}_{ji}$ by $\mathbf{n}_i$ on the left, and $\mathbf{y}_j$ can be rewritten as

$$\mathbf{y}'_j = \mathbf{g}'_j + \hat{\mathbf{v}}'_j, \tag{14}$$

where $\mathbf{g}'_j = [\mathbf{g}'_{j1}\mathbf{g}'_{j2} \cdots \mathbf{g}'_{jN_t}]^T$ with each component of *K*-dimensional vector $\mathbf{g}'_{ji} = [h_{ji}(0)h_{ji}(1) \cdots h_{ji}(L-1)0 \cdots 0]^T$ and $\hat{\mathbf{v}}'_j = [\mathbf{v}'_{j1}\mathbf{v}'_{j2} \cdots \mathbf{v}'_{jN_t}]^T$, with each component rewritten as $\mathbf{v}'_{ji} = [v_{ji}(n_i) \cdots v_{ji}(n_i + \lfloor K/N_t \rfloor - 1)0 \cdots 0]^T$.

If we choose a normalization factor of the transmission power, the FFT output of the frequency-domain subcarriers can be formulated as

$$\mathbf{Y}'_j = \mathbf{F}\mathbf{y}'_j = \mathbf{H}_j + \hat{\mathbf{V}}_j, \tag{15}$$

where $\mathbf{H}_j$ and $\hat{\mathbf{V}}_j$ are the FFT output of $\mathbf{g}'_j$ and $\hat{\mathbf{v}}'_j$, respectively, and $\mathbf{F}$ is an $K \times K$ FFT matrix given by

$$\mathbf{F} = \begin{pmatrix} W_K^{00} & W_K^{01} & \cdots & W_K^{0(K-1)} \\ W_K^{10} & W_K^{11} & \cdots & W_K^{1(K-1)} \\ \vdots & \vdots & \ddots & \vdots \\ W_K^{(K-1)0} & W_K^{(K-1)1} & \cdots & W_K^{(K-1)(K-1)} \end{pmatrix}, \tag{16}$$

with $W_K^{nk} = \exp\{-j2\pi \frac{nk}{K}\}/\sqrt{K}$. Therefore, channel $\mathbf{H}_j$ can be estimated using a simple least square (LS) method, and the LS estimation of the channel impulse response is

$$\hat{\mathbf{H}}_{LS} = \mathbf{Y}'_j = \hat{\mathbf{H}}_j. \tag{17}$$

Furthermore, this can be easily extended to the linear minimum mean square error (LMMSE) estimator. In order to have the minimum amount of information necessary, the estimator assumes a priori knowledge of noise variance and channel covariance and is expressed as

$$\hat{\mathbf{H}}_{LMMSE} = \mathbf{R}_H \left(\mathbf{R}_H + \frac{1}{\text{SNR}} \mathbf{I} \right)^{-1} \hat{\mathbf{H}}_{LS}, \tag{18}$$

where $\mathbf{R}_H = E[\mathbf{H}_j \mathbf{H}_j^H]$ is the auto-covariance matrix of $\mathbf{H}_j$, the superscript $(\cdot)^H$ denotes the Hermitian transpose, and $\mathbf{I}$ is the identity matrix.

Both estimators (17) and (18) have their merits and demerits. The complexity of the LMMSE estimator is larger than that of the LS estimator. For a high signal-to-noise ratio (SNR), the LS estimator is simple and adequate. The LMMSE channel estimator is better than the LS channel estimator because of its

higher ability for noise and interference suppression. However, when the SNR is high, and when co-channel interference is not present, the difference is not obvious.

V. Hybrid VBLAST Detection

In this section, we will present techniques for signal detection, including hybrid detection for a nulling vector and cancellation. The detail operation of the hybrid VBLAST detection scheme is illustrated in Fig. 2. Initially, the antennas are ranked in decreasing order of the irrespective received signal power using the power ranking information extracted from the power ranking scheme.

The regenerated signals from the first antenna to the *j*-th antenna at the *j*-th stage at time *l* are subtracted from the delayed version of received signal $r^0(l)$ as follows:

$$r^j(l) = r^0(l) - \sum_{k=1}^{j} \hat{r}_k^{j-1}(l), \tag{19}$$

where $\hat{r}_k^{j-1}(l)$ is the regenerated signal of the *k*-th antenna from the (*j*-1)th stage. This replica is regenerated by using the estimated channel based on the pilot preamble.

Equation (19) is the interference-cancelled signal for the input of a nulling and regeneration unit of all the antennas at the *j*-th stage. In Fig. 3, the regenerated signal of the *i*-th antenna at the previous stage, denoted as $\hat{r}_k^{j-1}(l)$, is added to the interference cancelled signal, denoted as $\hat{r}^j(l)$, in order to obtain the corresponding composite signal as follows:

$$\hat{r}_i^j(l) = r^0(l) - \sum_{\substack{k=1 \\ k\neq i}}^{j} \hat{r}_k^{j-1}(l) = r^j(l) + \hat{r}_i^{j-1}(l), \tag{20}$$

which is nulled by the corresponding antenna's nulling vector and is forwarded to the decision block. Therefore, the nulling and regeneration unit in the hybrid scheme performs only the projection of the received signal of the corresponding antenna. Finally, the decision variable z_i^j is forwarded to the next stage for a cleaner regeneration when $j < N_t$. On the other hand, when $j = N_t$, the decision variable is forwarded to the final bit decision device.

As the delayed detection proceeds, the delayed version of the received signal is subtracted by all of the other antennas' signals except for the desired antenna's signal. This hybrid process is repeated until the cancellation stage reaches the maximum stage. In the original sequential VBLAST detection scheme, the received signal $r^j(l)$ is subtracted by only one regenerated signal of the *j*-th antenna at every inter-stage. In the hybrid VBLAST detection scheme, however, it is subtracted by all regenerated signals from the first to the *j*-th antenna at every inter-stage.

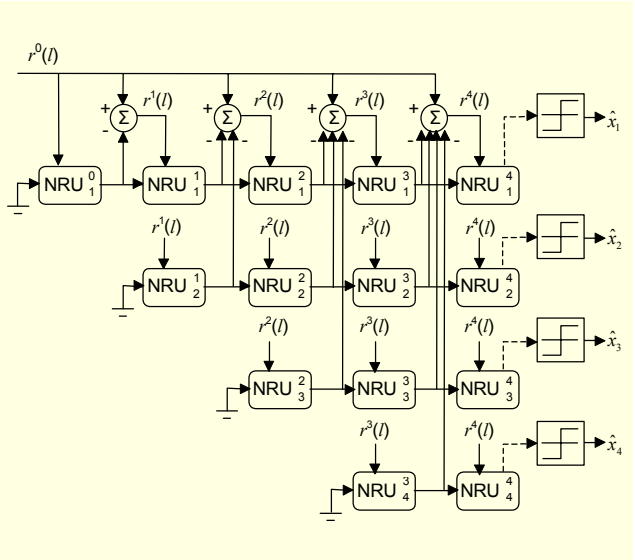

Fig. 2. The hybrid detection scheme for VBLAST system under $N_t = 4$ and $N_r = 4$.

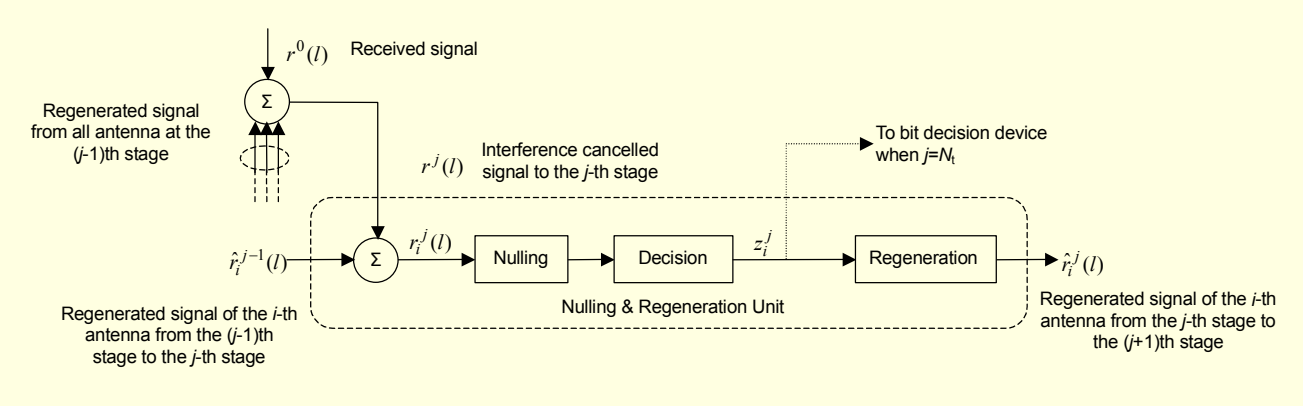

Fig. 3. The nulling and regeneration unit at the *j*-th stage of *i*-th antenna at time *l*.

VI. Modified Nulling of the Hybrid VBLAST Detection

For the $N_t \times 1$ vector of transmitted data symbols at time l, denoted as $\mathbf{x} = [x_1(l) x_2(l) \cdots x_{N_t}(l)]^T$, the corresponding received $N_r \times 1$ vector is

$$\mathbf{r}^0 = \mathbf{H}\mathbf{x} + \mathbf{v}, \tag{21}$$

where $\mathbf{H} = [\mathbf{h}_1 \mathbf{h}_2 \cdots \mathbf{h}_{N_r}]^T$ is a multi-path channel with each element of $\mathbf{h}_n = [h_{n1} h_{n2} \cdots h_{nN_t}]$, and $\mathbf{v} = [v_2(l)\ v_1(l) \cdots v_{N_t}(l)]^T$ is the AWGN vector with variance σ_v^2.

We perform the sequential detection of the elements in $\mathbf{x}$. Note that we do not need to detect element x_i in the order of $i = 1, 2, \cdots, N_t$. Therefore, the optimal ordering to minimize the detection error is founded. It turns out that we can obtain the optimal ordering by selecting the minimum-norm column of $\mathbf{H}^{\dagger}$, where $(\cdot)^{\dagger}$ indicates a pseudo-inverse. Let the optimal detection ordering be $[x_{k_1} x_{k_2} \cdots x_{k_{N_t}}]$.

To detect the i-th element of $\mathbf{x}$, x_{k_i}, we perform zero-forcing nulling. We find the minimum norm weight vector $\mathbf{w}_{k_i}$ such that

$$\mathbf{w}_{k_i}^H \mathbf{h}_{k_i} = \delta_{k_i}, \tag{22}$$

where $(\cdot)^H$ means the complex conjugate transpose. Weight vector $\mathbf{w}_{k_i}$ can be obtained from the pseudo-inverse of $\mathbf{H}$, and using its estimator, $\hat{\mathbf{H}}$, we obtain the weight vector,

$$\hat{\mathbf{w}}_{k_i} = (\hat{\mathbf{G}}_i)_{k_i}^*, \tag{23}$$

where

$$\hat{\mathbf{G}}_{i+1} = \hat{\mathbf{H}}_{k_i-}^{\dagger} \tag{24}$$

and the notation $\mathbf{H}_{k_i-}^{\dagger}$ denotes the matrix obtained by zeroing columns $k_1, k_2, \cdots, k_i$ of $\mathbf{H}$.

When the i-th symbol is detected, the received vector $\mathbf{r}^j$ after cancelling x_{k_l}, $l = 0, 1, \cdots, j-1$, becomes

$$\mathbf{r}^j = \mathbf{r}^0 - \sum_{l=0}^{j-1} \hat{x}_{k_l} \hat{\mathbf{h}}_{k_l} + \mathbf{v}. \tag{25}$$

Assume that no detection error of x_{k_l} gives

$$\begin{aligned} \mathbf{r}^j &= \mathbf{H}\mathbf{x} - \sum_{l=0}^{j-1} x_{k_l} \hat{\mathbf{h}}_{k_l} + \mathbf{v} \\ &= \mathbf{H}_{k_i-}\mathbf{x}_{k_i-} - \sum_{l=0}^{j-1} x_{k_l} \mathbf{h}_{k_l} + \mathbf{v} \\ &= \hat{\mathbf{H}}_{k_i-}\mathbf{x}_{k_i-} - \Delta\mathbf{H}_{k_i-}\mathbf{x}_{k_i-} - \sum_{l=0}^{j-1} x_{k_l} \mathbf{h}_{k_l} + \mathbf{v}, \end{aligned} \tag{26}$$

where

$$\begin{aligned} \mathbf{x}_{k_i-} &= [x_{k_i} x_{k_{i+1}} \cdots x_{k_{N_t}}], \\ \hat{\mathbf{H}}_{k_i-} &= [\hat{\mathbf{h}}_{k_i} \hat{\mathbf{h}}_{k_{i+1}} \cdots \hat{\mathbf{h}}_{k_{N_t}}], \\ \Delta\mathbf{H}_{k_i-} &= [\Delta\mathbf{h}_{k_i} \Delta\mathbf{h}_{k_{i+1}} \cdots \Delta\mathbf{h}_{k_{N_t}}]. \end{aligned} \tag{27}$$

If $\Delta\mathbf{H} = 0$, $\hat{\mathbf{w}}_{k_i}$ in (23) is an optimal nulling vector. But in a real system, there is nonzero matrix $\Delta\mathbf{H}$, and $\hat{\mathbf{w}}_{k_i}$ is not the optimal nulling vector. We derive hereafter the new nulling vector which minimizes the unexpected effect of the channel estimation error. Without loss of generality, the new nulling vector $\overline{\mathbf{w}}_{k_i}$ is written as

$$\overline{\mathbf{w}}_{k_i} = \hat{\mathbf{w}}_{k_i} + \Delta\mathbf{w}_{k_i}, \tag{28}$$

where $\Delta\mathbf{w}_{k_i}$ indicates the difference between original nulling vector $\hat{\mathbf{w}}_{k_i}$ and new nulling vector $\overline{\mathbf{w}}_{k_i}$. With the new nulling vector, we obtain the new decision statistic, and the estimate of $\hat{x}_{k_i}$ is as follows:

$$\hat{x}_{k_i} = Q(\overline{\mathbf{w}}_{k_i}^H \mathbf{r}_i^j), \tag{29}$$

where $Q(\cdot)$ is the quantization operation appropriate to the constellation in use.

VII. Simulation Results and Discussions

In this section, we evaluate the detection performances of both a perfect channel and LMMSE channel estimations for the various VBLAST detection schemes for a Rayleigh fading channel with a multi-path length of $L = 16$. The channel model is assumed to be an equal and exponential multi-path power delay profile.

In Fig. 4, the channel estimation is done by using the pilot preamble described, and the performance of the LMMSE estimator is measured using MSE $= E[(\mathbf{H} - \hat{\mathbf{H}}_{\text{LMMSE}})^2]$. This figure validates the effectiveness of the investigated preamble structure, which is designed based on the assumption that the length of the guard interval in both ETSI HIPERLAN/2 and IEEE802.11a standards is 16. In general, this means that time delay L (or, delay spread) is less than 16. So, multi-paths of 9 and 16 are corresponding to the optimum and worst cases, respectively. The worst case is a 1.5 dB degradation compared with the optimum case. As the propagation length of L exceeds the window size of 16, the MSE of the LMMSE estimator increases and the MSE performance of the exponential power delay profile is less decreased than that of the equal power delay profile. When the SNR is less than 30dB, fortunately, the MSE degradation in the

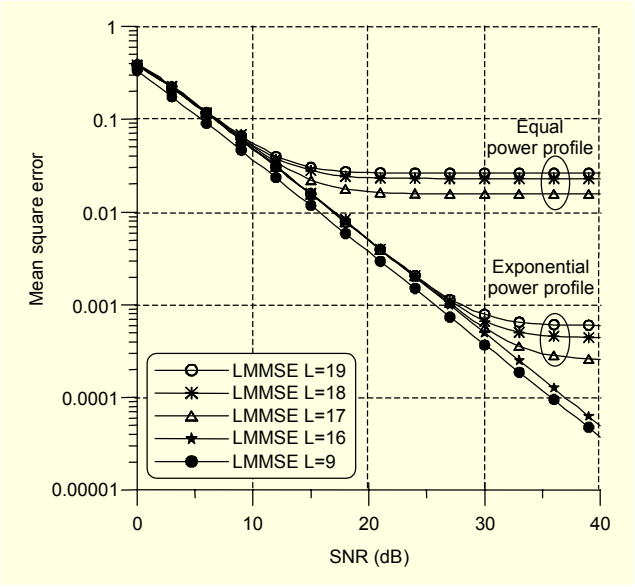

Fig. 4. MSE performance of LMMSE estimation for pilot preamble according to the type of delay profile with $N_t = 4$ and $N_r = 4$.

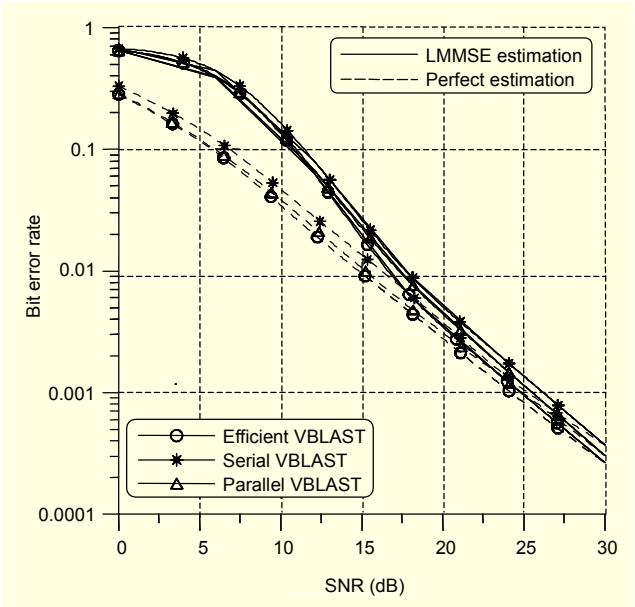

Fig. 5. BER performance of various OFDM-VBLAST detection schemes based on both perfect and LMMSE estimations in the case of $N_t = 2$ and $N_r = 2$.

case of the exponential delay profile is negligible for some extra channel propagations. However, there is an irreducible MSE floor due to imperfect windowing for a relatively high SNR.

As for the channel estimation in the OFDM system with multiple antennas, some works formulated the training preamble sequences by minimizing the error covariance of the time-domain channel response [3], [4], which is basically similar to our approach. So, both approaches attain the same MSE bound [3]. If the condition of $\lfloor K / L_t \rfloor \geq L$ is satisfied

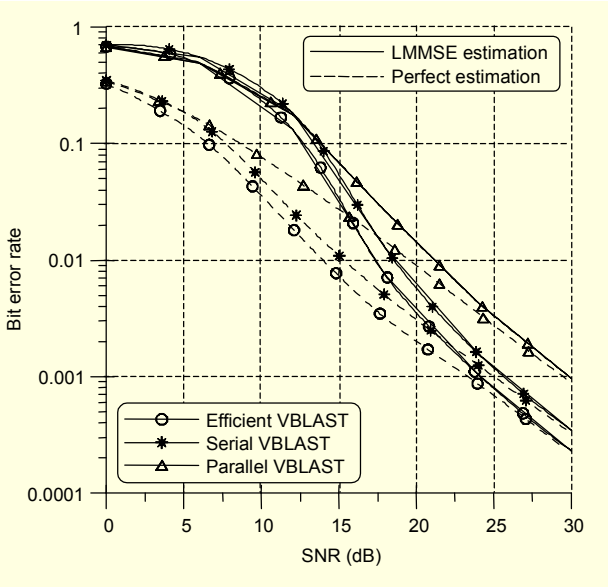

Fig. 6. BER performance of various OFDM-VBLAST detection schemes in the case of of N_t=4 and N_r=4.

and some parameters are suitably chosen, our approach in the time domain can give the same MSE performance bound to the above mentioned approaches. On the other hand, our approach is simpler for analysis than others [3], [4] because the channel estimation is simply done by only shifting by n_i, as we can see from (14) and (15).

Figures 5 and 6 show the evaluated bit error rate (BER) performances of both perfect and LMMSE channel estimations for the various VBLAST detection schemes of N_t=N_r=2 and N_t=N_r=4. In Fig. 6, we can observe that the hybrid detection method provides improvements of approximately 2 dB and 7 dB over the classical methods with serial and parallel detection, respectively, when the required BER is 10^{-3}. Also, it can be easily observed from the BER results of Figs. 5 and 6 that the performance of parallel detection is markedly better than that of classical VBLAST detection for N_t=2. In the case of N_t=3, however, its performance becomes worse. And the ordinary VBLAST based on serial and parallel detections requires N_t and $2N_t$ nulling operations, respectively. The hybrid VBLAST detection scheme needs $N_t(N_r+1)/2$ nulling operations, which results in additional complexity. On the other hand, the total number of rows used to obtain the pseudo-inverse matrix of $\mathbf{H}_{k_i-}$ is calculated as $N_t(N_t+1)/2$, $N_t(N_t+1)$ and $N_t(N_t+1)(N_t+2)/6$ for serial, parallel and hybrid detection, respectively. In the case of N_t=N_r=4, for example, the total number of rows for hybrid detection is equal to that of parallel detection [7] and double that of sequential detection [2].

Figure 7 illustrates the effect of the number of transmitting antennas on the BER performance of the perfect estimation

with SNR as a parameter. As the number of transmitting antennas increases, as expected, the BER increases. Regardless of the number of transmitting antennas and SNR, the proposed hybrid detection scheme appears to be a stable tendency over other reference VBLAST systems [2], [7].

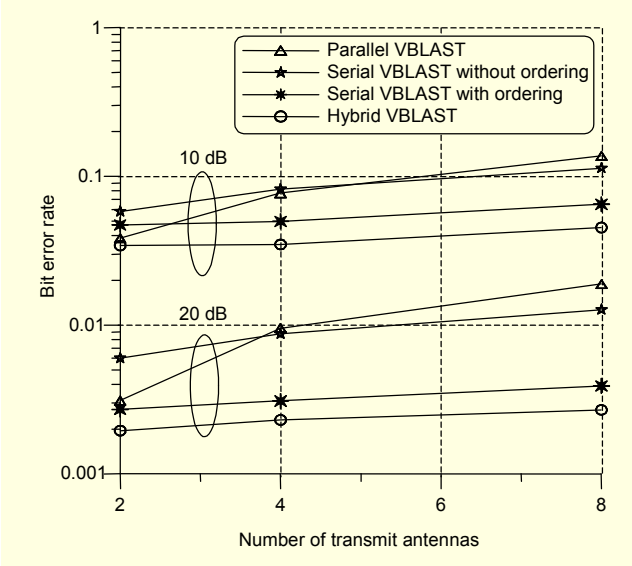

Fig. 7. BER performance of perfect estimations according to the number of transmitting antennas for various MIMO-BLAST detection schemes.

VIII. Conclusions

In this paper, OFDM with multiple transmitting and receiving antennas has been used to form an OFDM-based VBLAST system to increase system capacity. An efficient detection algorithm for a pilot symbol assisted interference nulling vector and cancellation has been studied for OFDM-based VBLAST systems. We have shown that the BER performance of hybrid detection outperforms ordinary OFDM-VBLAST based on serial and parallel detections at the expense of a small hardware complexity.

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