An Ordered Successive Interference Cancellation Scheme in UWB MIMO Systems

Jinyoung An and Sangchoon Kim

ABSTRACT—In this letter, an ordered successive interference cancellation (OSIC) scheme is applied for multiple-input multiple-output (MIMO) detection in ultra-wideband (UWB) communication systems. The error rate expression of an OSIC receiver on a log-normal multipath fading channel is theoretically derived in a closed form solution. Its bit error rate performance is analytically compared with that of a zero forcing receiver in the UWB MIMO detection scheme followed by RAKE combining.

Keywords—Ultra-wideband (UWB), MIMO, zero forcing (ZF), ordered successive interference cancellation (OSIC).

I. Introduction

Ultra-wideband (UWB) radio transmission technology has attracted enormous interests for applications requiring high data rates over short ranges. To increase the transmission data rate through spatial multiplexing (SM) and extend system RF coverage, multiple-input multiple-output (MIMO) techniques can be employed in UWB systems for certain applications such as high-definition video transmission [1]. In [1], the error performance of UWB MIMO systems over indoor wireless multipath channels was analyzed. A zero forcing (ZF) receiver is applied to separate the spatially multiplexed data on a path-by-path basis. The zero-forced resolvable paths are then combined by a RAKE. This detection structure is called a ZF-RAKE. Also, the minimum mean square error (MMSE) detector for MIMO systems on flat Rayleigh channels was considered in [2] and [3].

Manuscript received April 5, 2009; revised April 29, 2009; accepted May 11, 2009.

This work was supported by Korea Research Foundation Grant funded by the Korean Government (MOEHRD, Basic Research Promotion Fund) (KRF-2008-331-D00382).

Jinyoung An (phone: +82 51 200 5589, email: sirius0422@nate.com) and Sangchoon Kim (phone: +82 51 200 7795, email: sckim@dau.ac.kr) are with the Department of Electronics Engineering, Dong-A University, Busan, Rep. of Korea.

doi:10.4218/etrij.09.0209.0140

Performance of the ZF detector is known to be similar to that of the MMSE receiver at high signal-to-noise ratios (SNRs). This work focuses on the analysis of a ZF related scheme. In this letter, we present the error performance of a ZF scheme with an ordered successive interference cancellation (OSIC) for UWB SM MIMO detection in a log-normal fading channel. We analytically investigate the theoretic bit error rate (BER) performance of the ZF-OSIC architecture followed by RAKE combining, called a ZF-OSIC-RAKE, in an UWB SM MIMO receiver. The diversity order that the ZF-OSIC-RAKE offers for detecting the n-th transmitted data stream is obtained for UWB MIMO systems. It is also seen that the ZF-OSIC-RAKE scheme outperforms the ZF-RAKE.

II. System Model

A multiple antenna UWB communication system with N_T transmit and N_R receive antennas $(N_R \geq N_T)$ is considered. Independent N_T data streams modulated with the pulse-amplitude of a short-duration UWB pulse are sent on different transmit antennas. To avoid severe intersymbol interference, we assume that the pulse repetition interval is sufficiently large compared with the channel delay spread. We also assume that the signals put through the multipath channel, and only L resolvable paths for the symbol-by-symbol detection are exploited by the receiver at each receive antenna. The received signal of the l-th propagation path at the m-th receive antenna denoted by $x_m(l)$, $l = 0, 1, \cdots, L-1$, can be obtained by passing through a UWB pulse matched filter. The discrete-time received signal vector, $\mathbf{x}(l) = \left[x_1(l)\ x_2(l)\ \cdots\ x_{N_R}(l)\right]^{\mathrm{T}}$, at the matched filter output for the l-th path can be written as

$$\mathbf{x}(l) = \sqrt{E_s}\ \mathbf{H}(l)\ \mathbf{a} + \mathbf{w}(l), \tag{1}$$

where E_s is the average symbol energy. Here, the information symbol vector over N_T transmit antennas is denoted by $\mathbf{a}=[a_1\ a_2 \cdots a_{N_T}]^T$, where a_n is an information symbol from the n-th transmit antenna. The received noise vector is represented by $\mathbf{w}(l)=\left[w_1(l)\ \ w_2(l)\ \cdots\ w_{N_R}(l)\right]^T$, where $w_m(l)$ is a real zero-mean white Gaussian noise of variance $N_0/2$. For the l-th delayed path, an $N_R \times N_T$ discrete-time channel matrix $\mathbf{H}(l)$ can be expressed as $\mathbf{H}(l)=\left[\mathbf{h}_1(l)\ \mathbf{h}_2(l) \cdots \mathbf{h}_{N_T}(l)\right]$ where the column vector $\mathbf{h}_n(l)$ is given as $\mathbf{h}_n(l)=\left[h_{1n}(l)\ h_{2n}(l) \cdots h_{N_R n}(l)\right]^T$. Here, $h_{mn}(l)$, $m=1,2,\cdots,N_R$, $n=1,2,\cdots,N_T$, is the channel coefficient of the l-th path for the signal from the n-th transmit antenna to the m-th receive antenna. It is given by $h_{mn}(l)=\zeta_{mn}(l)\,g_{mn}(l)$ [4], where $\zeta_{mn}(l)\in\{\pm 1\}$ with equal probability is a discrete random variable (RV) representing pulse-phase inversion, and $g_{mn}(l)$ is the fading magnitude term with a log-normal distribution.

III. Ordered Successive Interference Cancellation

In the ZF-RAKE examined in [1], ZF detection is performed by the filter matrix $\mathbf{F}_{ZF}(l)=\mathbf{H}(l)^+$, where $(\cdot)^+$ denotes the pseudoinverse. The output signal vector, $\mathbf{y}(l)=\left[\ y_1(l)\ \ y_2(l)\ \cdots\ y_{N_R}(l)\ \right]^T$, for the l-th path of a particular bit is given by $\mathbf{y}(l)=\mathbf{F}_{ZF}(l)\mathbf{x}(l)$. The decision variable (DV) of the MRC output for a particular bit of the n-th transmitted data stream can be written as

$$z_n=\sum_{l=0}^{L-1}\alpha_n^2(l)\,y_n(l), \tag{2}$$

where $\alpha_n^2(l)=1/\upsilon_n(l)$ and $\upsilon_n(l)=\left[\mathbf{H}(l)^H\mathbf{H}(l)\right]_{nn}^{-1}$. Here, $(\cdot)^H$ is the conjugate transpose. The diversity order of an (N_R, N_T, L) MIMO system based on the ZF-RAKE scheme is given by the parameter $D_{ZF}=L(N_R-N_T+1)$.

In a ZF-RAKE, a bank of separate filters has been considered to estimate the N_T data substreams. However, the output of one of the filters can be exploited to help the operation of the others. In the OSIC technique [5], the signal with the highest post-detection SNR is first chosen for processing and then cancelled from the overall received signal vector. This reduces the burden of inter-channel interference on the receivers of the remaining data substreams. To apply OSIC to the ZF-RAKE architecture for multipath channels, the OSIC procedure for each path is performed prior to RAKE combining. The zero-forced signal is first obtained in the detection step of each propagation path signal. Then, the zero-forced signals are combined to detect each substream. This detection scheme is called a ZF-OSIC-RAKE. The OSIC algorithm for the l-th path signal is shown in Fig. 1. Here,

for $i=1, 2, \cdots, N_T$
$\quad \mathbf{F}_{ZF,i}(l)=\mathbf{H}_i(l)^+$
$\quad k_i(l)=\arg\min_j \left\| \mathbf{f}_i^{(j)}(l) \right\|^2$
$\quad y_{k_i(l)}(l)=\mathbf{f}_i^{(k_i(l))}(l)\,\mathbf{x}_i(l)$
$\quad \mathbf{x}_{i+1}(l)=\mathbf{x}_i(l)-\mathbf{h}_{k_i(l)}(l)\cdot Quan\left[y_{k_i(l)}(l)\right]$
$\quad \mathbf{H}_{i+1}(l)=\mathbf{H}_i^{\overline{k}_i(l)}(l)$
end
Permutate the elements of $\mathbf{y}(l)\left(=\left[\ y_{k_1(l)}(l)\ \ y_{k_2(l)}(l)\ \cdots\ y_{k_{N_R}(l)}(l)\ \right]^T\right)$ according to the original sequence $(1,2,\cdots,N_T)$.

Fig. 1. ZF-OSIC algorithm for each path in ZF-OSIC-RAKE.

$(\mathbf{H}_i(l), \mathbf{F}_{ZF,i}(l), \mathbf{x}_i(l))$ denotes the specific variables in the i-th detection step. Initial values are set to be $\mathbf{H}_1(l)=\mathbf{H}(l)$, $\mathbf{F}_{ZF,1}(l)=\mathbf{F}_{ZF}(l), \mathbf{x}_1(l)=\mathbf{x}(l)$; $\mathbf{f}_i^{(j)}(l)$ and $Quan[\bullet]$ indicate the row j of $\mathbf{F}_{ZF,i}(l)(=\mathbf{H}_i(l)^+)$ and quantization operation, respectively; and $\mathbf{H}_{i+1}(l)=\mathbf{H}_i^{\overline{k}_i(l)}(l)$ depicts the nulling of column $k_i(l)$ of the channel matrix $\mathbf{H}_i(l)$.

To obtain the single zero-forced signal of the l-th path signal in OSIC step i, the row of $\mathbf{F}_{ZF,i}(l)$ with the minimal norm is chosen. After getting the zero-forced signal vector associated with each multipath component, we reorder the elements of the zero-forced vector according to the original sequence of the transmitted data streams. Then, RAKE combining of the zero-forced signals is accomplished to detect each data stream and leads to obtaining a DV for each transmitted data stream. The DV of the RAKE combiner output for a particular bit of the n-th transmitted data stream can be written as (2).

The instantaneous SNR γ_n of the DV of the RAKE combiner output for a particular bit of the n-th data stream is given by $\gamma_n=2\lambda\sum_{l=0}^{L-1}\alpha_n^2(l)$ with $\lambda=E_s/N_0$. By following an analysis similar to that in [1], we obtain the quadratic form, $\alpha_n^2(l)=\mathbf{h}_n(l)^H\mathbf{S}_n(l)\mathbf{h}_n(l)$, where $\mathbf{S}_n(l)$ is a $N_R \times N_R$ non-negative Hermitian matrix constructed from $\mathbf{h}_{n+1}(l), \mathbf{h}_{n+2}(l), \cdots, \mathbf{h}_{N_T}(l)$. Then, γ_n can be expressed as $\gamma_n=2\lambda\sum_{i=1}^{LN_R}\lambda_i^{(n)}{q_i^{(n)}}^2=2\lambda\gamma_n'$, where $q_i^{(n)}$ and $\lambda_i^{(n)}$, respectively, are the zero-mean unit-variance Gaussian RV and eigenvalue of the matrix $\mathbf{S}_n=diag\left[\mathbf{S}_n(0)\ \mathbf{S}_n(1)\cdots\ \mathbf{S}_n(L-1)\right]$. Thus, the eigenvalues of $\mathbf{S}_n(l)$ with N_R-N_T+n are equal to 1, and the others with N_T-n eigenvalues are 0. Hence, the matrix $\mathbf{S}_n$ has LN_R eigenvalues, among which the $L(N_R-N_T+n)$ eigenvalues are equal to 1, and the other $L(N_T-n)$ eigenvalues are equal to 0; therefore, $\gamma_n'=\sum_{i=0}^{L(N_R-N_T+n)}\left|q_i^{(n)}\right|^2$. The variable γ_n' is a central chi-square distributed RV with $D_{ZF\text{-}OSIC}^{(n)}=L(N_R-N_T+n)$ degrees of freedom, which has a probability density function (PDF) of

$$f_{\gamma_n'}(t)=\left(0.5^{\kappa}/\Gamma(\kappa)\right)t^{\kappa-1}e^{-t/2}, \tag{3}$$

where $\kappa = 0.5\, D_{\text{ZF-OSIC}}^{(n)}$, and $\Gamma(\cdot)$ is the gamma function defined as $\Gamma(b) = \int_0^\infty \tau^{b-1} e^{-\tau} d\tau$, $b > 0$. Thus, by averaging the conditional BER $P_r^{(n)}(t) = Q\left(\sqrt{2\lambda t}\right)$ over the PDF $f_{\gamma_n'}(t)$, the average BER for a particular bit of the *n*-th transmitted data stream over log-normal fading channels can be calculated as

$$P_{b,\text{ZF-OSIC}}^{(n)} = \int_0^\infty P_r^{(n)}(t) f_{\gamma_n'}(t)\, dt . \qquad (4)$$

To derive the average BER in a closed form, a tightly approximated expression of *Q*-function such as $Q(\beta) \simeq (1/12)e^{-\beta^2/2} + (1/6)e^{-2\beta^2/3}$ [2] is employed. Then, the conditional BER, given $\lambda = E_s/N_0$, can be approximated as

$$P_r^{(n)}(t) \simeq (1/12)e^{-\lambda t} + (1/6)e^{-4\lambda t/3} . \qquad (5)$$

By using (3) and (5) in (4), the average BER for the ZF-OSIC-RAKE expressed in a closed form can be obtained as

$$P_{b,\text{ZF-OSIC}}^{(n)} \simeq \frac{(\kappa-1)!}{12(2\lambda+1)^{\kappa}\,\Gamma(\kappa)} + \frac{(\kappa-1)!}{6((8/3)\lambda+1)^{\kappa}\,\Gamma(\kappa)} . \quad (6)$$

Here, $(\cdot)!$ is the factorial defined as $s! = s\cdot(s-1)\cdots 2\cdot 1$. The average BER over all N_T data streams can be given by $BER_{\text{ZF-OSIC}} = (1/N_T)\sum_{n=1}^{N_T} P_{b,\text{ZF-OSIC}}^{(n)}$. Note that to obtain the theoretical performance of the ZF-RAKE, the same formula as that of ZF-OSIC-RAKE can be used except $\kappa = 0.5\, D_{\text{ZF}}$.

IV. Simulation Results

Consider the binary PAM scheme (E_s=E_b). In the simulations, the SNR per bit in decibels is defined as $\eta_b = 2\lambda(\text{dB}) + 10\log_{10} L(N_R - N_T + 1)$ for (N_R, N_T, L) systems. The log-normal fading amplitude $g(l)$ can be expressed as $g(l) = e^{\psi(l)}$, where $\psi(l)$ is a Gaussian RV with mean $\mu_{\psi(l)}$ and variance σ_ψ^2. We assume that the standard deviation of $20\log_{10} g(l) = \psi(l)(20\log_{10} e)$ is 5 dB. To satisfy $E[g(l)^2] = e^{-\rho l}$ representing the average power of path *l*, $\mu_{\psi(l)} = -\sigma_\psi^2 - \rho l/2$ is required, where σ_ψ is obtained as $\sigma_\psi = 5/(20\log_{10} e)$. Here, the power decay factor $\rho = 0$ is used. The receiver is assumed to have perfect knowledge of the channel fading coefficients of *L* resolvable paths. In the figures, theoretic and simulated BER curves of the UWB MIMO system are shown with lines and markers, respectively.

For a (4, 3, *L*) MIMO system, Fig. 2(a) shows the analytical and simulated BERs of the ZF-RAKE and ZF-OSIC-RAKE as a function of SNR per bit in decibels. The analytical results are similar to the simulated BER performance and demonstrate the performance boost in using the ZF-OSIC-RAKE over the ZF-RAKE. This boost in BER performance comes from a

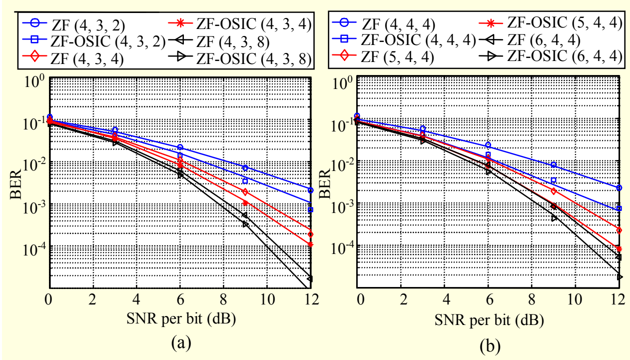

Fig. 2. BER vs. SNR for different values of (a) *L* and (b) N_R.

power gain due to the increase of diversity order in each increasing cancellation step. Also, multipath combining yields enhanced performance in both ZF-OSIC-RAKE and ZF-RAKE. In an (N_R, 4, 4) case, Fig. 2(b) shows the analytical and simulated results of BER as a function of SNR per bit for various N_R. The increase in N_R increases the diversity order of the ZF-OSIC-RAKE receiver as well as that of the ZF-RAKE.

V. Conclusion

The error performance of a UWB MIMO system over indoor log-normal fading channels has been analyzed. The UWB MIMO receiver employs ZF-OSIC detectors and then a RAKE temporal combiner. It has been shown that the ZF-OSIC for detecting the *n*-th transmitted data stream in an (N_R, N_T, *L*) MIMO system achieves a diversity order of $L(N_R - N_T + n)$. The ZF-OSIC offers better performance than the ZF receiver.

References

[1] H. Liu, R.C. Qiu, and Z. Tian, “Error Performance of Pulse-Based Ultra-Wideband MIMO Systems over Indoor Wireless Channels,” *IEEE Trans. Wireless Commun.*, vol. 4, no. 6, Nov. 2005, pp. 2939-2944.

[2] N. Kim, Y. Lee, and H. Park, “Performance Analysis of MIMO System with Linear MMSE Receiver,” *IEEE Trans. Wireless Commun.*, vol. 7, no. 11, Nov. 2008, pp. 4474-4478.

[3] A. Zanella, M. Chiani, and M.Z. Win, “MMSE Reception and Successive Interference Cancellation for MIMO Systems with High Spectral Efficiency,” *IEEE Trans. Wireless Commun.*, vol. 4, no. 3, May 2005, pp. 1244-1253

[4] A.F. Molish, J.R. Foerster, and M. Pendergrass, “Channel Models for Ultra-Wideband Personal Area Network,” *IEEE Wireless Commun.*, vol. 10, no. 6, Dec. 2003, pp. 14-21.

[5] P.W. Wolniansky et al., “V-BLAST: An Architecture for Realizing Very High Data Rates Over the Rich-Scattering Wireless Channel,” *Proc. IEEE ISSSE*, Sept. 1998, pp. 295-300.