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Taps Delayed Lines Architecture Based on Linear Transmit Zero-Forcing Approach for Ultra-Wide Band MIMO Communication Systems

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***Abstract*—In this paper, a transmitter-based multipath processing and inter-channel interference (ICI) cancellation scheme for a ultra-wideband (UWB) spatial multiplexing (SM) multiple input multiple output (MIMO) system is presented. It consists of taps delayed lines and zero-forcing (ZF) filters in the transmitter and correlators in the receiver. For a UWB SM MIMO system with *N* transmit antennas, *M* receive antennas, and *Q* resolvable multipath components, the BER performance of a linear transmit ZF scheme is analyzed in a log-normal fading channel and also compared with that of a receiver-based ICI rejection approach. It is found that when $M \le N$, the transmit ZF processing approach outperforms the ZF receiver while making the mobile units low-cost and low-power.**

***Index Terms*— Ultra-WideBand (UWB), Zero-Forcing (ZF), Rake, Prerake, Taps Delayed Line (TDL), Spatial Multiplexing**

I. INTRODUCTION

ULTRA-wideband (UWB) radio transmission technology has emerged as a potential candidate for future high speed indoor radio communications [1]. To increase the transmission data rate and expand system RF coverage, multiple input multiple output (MIMO) techniques [2] can be used in UWB systems for certain applications such as high definition video transmission. In [2], the performance analysis of UWB MIMO systems over indoor wireless multipath channels has been based on the receiver consisting of zero-forcing (ZF) detectors followed by Rake combiners. This detection architecture is called a RxZF in this work. Multistream communications through spatial multiplexing (SM) require the MIMO detection processing such as ZF filtering at the receiver [2], which is computationally expensive. If this receive processing for downlink can be shifted to the transmitter side, the mobile units become simple and power-efficient [3],[4]. In addition, the Rake processing to obtain path diversity and collect multipath signals in the receiver can be replaced with the Prerake processing in the transmitter [3]. Thus, to significantly reduce the receiver complexity in the mobiles, it is desirable that the Rake processing and ZF filtering units are displaced from the portable receivers to the fixed transmitter. The linear transmit processing for MIMO systems has been examined in [5]. In [6], a decorrelating prefilter is employed for rejecting multiuser interference in a code division multiple access (CDMA) downlink system on a multipath environment.

In this paper, we consider taps delayed lines (TDLs) and linear transmit ZF processing for UWB SM MIMO detection in log-normal multipath fading channels. This technique is called a TxZF. The BER performance of the TxZF systems for UWB SM MIMO detection is analyzed in a log-normal fading channel. The BERs of the TxZF systems are evaluated and compared with those of the RxZF. It is observed that when the number of transmitted multistreams is less than or equal to the number of transmit antennas, the TxZF outperforms the RxZF and its receiver has much lower complexity.

II. LINEAR TRANSMIT ZERO-FORCING APPROACH

In this section, we present a linear transmit ZF scheme for the pulsed UWB SM MIMO systems with binary pulse-amplitude modulation (2PAM) as shown in Fig. 1, where the inter-channel interference (ICI) and multipath effects, respectively, are suppressed and combined by the transmitter through ZF filtering combined with TDLs. The TxZF $M \times N$ MIMO system under Q resolvable multipath where M and N, respectively, are the number of receive and transmit antennas is considered. It is assumed that the transmitter selects only a subset of $Q\left(\le Q_t\right)$ resolvable paths among the total number of Q_t resolvable multipath components. We assume that $M \le N$ for transmit processing. The serial input data are converted into M parallel streams. In the TxZF system's transmitter, Q pulses transmitted in each bit interval are each scaled by factors determined by the pseudoinverse operation of the channel matrix related to the multipath channel

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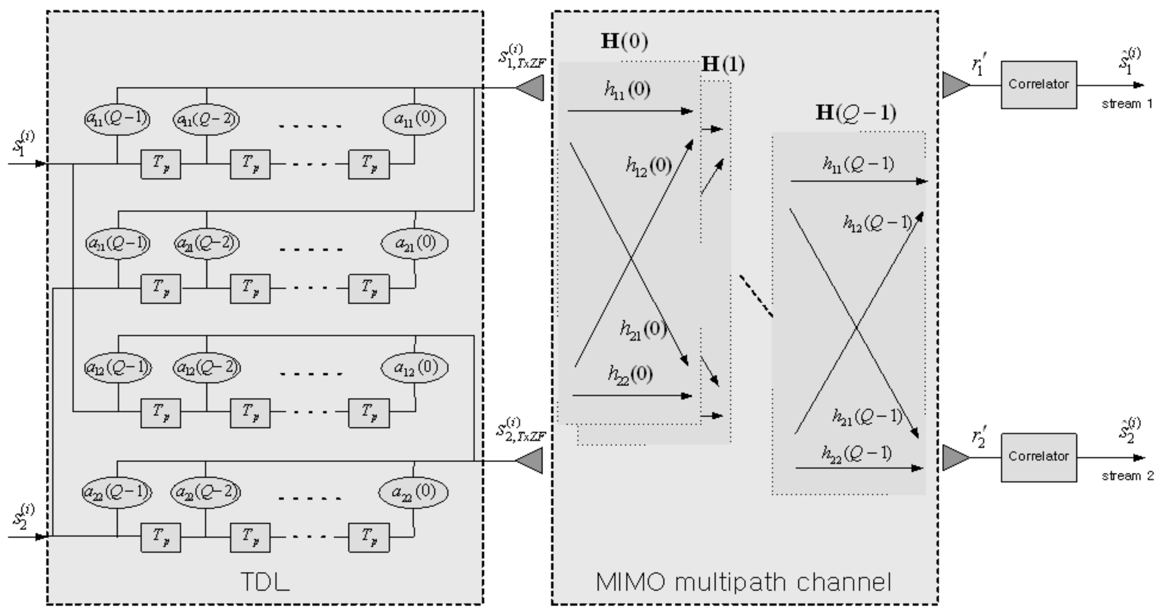

Fig. 1. Linear Transmit Zero-Forcing 2×2 MIMO system with Q resolvable multipath components

coefficients and delayed according to the delay of multipath. Eventually, the independent parallel signals are sent to N transmit antennas for simultaneous transmission to the MIMO multipath channel. In the receiver, only one correlator at each receive antenna is needed for signal detection where the operations of channel estimation and multipath tracking are not required. The transmitted signal from the nth transmit antenna in the ith bit interval in a TxZF SM MIMO system is expressed as

$$s_{n,TxZF}^{(i)}(t)=\sum_{m=1}^{M}\sum_{q=0}^{Q-1}a_{nm}(Q-1-q)s_m^{(i)}(t+qT_p) \tag{1}$$

where $s_m^{(i)}(t)=\sqrt{E_b}b_m(i)p(t-iT_r)$. Here $p(t)$ is the UWB pulse of width T_p with unit energy and E_b is the average bit energy. Compared with the channel delay spread, the pulse repetition period $T_r\left(\gg T_p\right)$ must be large enough to avoid severe inter-symbol interference (ISI) caused by the multipath channel. $b_m(i)\in\{\pm 1\}$ is the ith information bit generated from the mth data stream. The prefilter factor $a_{nm}(q)$ is the (n,m)th element of conjugate transpose of a prefilter matrix. In this work, the ZF filter is employed as a prefilter at the transmitter. The ZF filter matrix with $(\cdot)^{+}$ and $(\cdot)^{H}$ being the pseudoinverse and conjugate transpose, respectively, is defined as

$$\begin{aligned}\mathbf{G}_{TxZF}(q)^H&=\left(\mathbf{H}(q)^{H^{\dagger}}\right)^H\\&=\mathbf{H}(q)^H\left(\mathbf{H}(q)\mathbf{H}(q)^H\right)^{-1}\end{aligned} \tag{2}$$

Where

$$\mathbf{H}(q)=\left[\,\mathbf{h}_1(q)\;\mathbf{h}_2(q)\;\cdots\;\mathbf{h}_N(q)\,\right] \tag{3}$$

$$\mathbf{h}_n(q)=\left[\,h_{1n}(q)\;h_{2n}(q)\;\cdots\;h_{Mn}(q)\,\right]^T \tag{4}$$

Here $h_{mn}(q)$ represents the channel fading coefficient of the qth resolvable path to the mth receive antenna sent from the nth transmit antenna in the multipath channel model, which can be described as a discrete linear filter with a channel impulse response [1]

$$\alpha_{mn}(t)=\sum_{q=0}^{Q_t-1}h_{mn}(q)\delta(t-qT_p) \tag{5}$$

where $\delta(t)$ is the Dirac delta function and T_p is the smallest multipath resolution. The channel gain model can be described as $h_{mn}(q)=\kappa_{mn}(q)c_{mn}(q)$ where $c_{mn}(q)$ is the fading magnitude term with a log-normal fading distribution and $\kappa_{mn}(q)\in\{\pm 1\}$ is the equal probable phase inversion.

After the transmitted signal goes through the multipath fading channel modeled by (5), the received signal to the mth receive antenna from all N transmit antennas is written as

$$\begin{aligned}r'_m(t)=&\sum_{n=1}^{N}\sum_{m'=1}^{M}\sqrt{E_b}b_{m'}(i)\sum_{q=0}^{Q-1}a_{nm'}(Q-1-q)\\&\times\sum_{k=0}^{Q_t-1}h_{mn}(k)p(t-iT_r+qT_p-kT_p)+w_m(t)\end{aligned} \tag{6}$$

where $w_m(t)$ is a real zero-mean white Gaussian noise process with a two-sided power spectral density of $N_0/2$. For detecting the ith bit, only the strongest path signal component is picked up by the correlator receiver at each receive antenna [4]. The correlator receiver output at the mth receive antenna is given by

$$v_m = \int_{-\infty}^{\infty} r'_m(t)\, p(t - iT_r)\, dt \tag{7}$$

It is assumed that time intervals between all neighboring paths are bigger than the UWB pulse width T_p. Thus, in the absence of inter-path interference (IPI), the received signal v_m can be rewritten as

$$\begin{aligned} v_m &= \sum_{n=1}^{N}\sum_{m'=1}^{M}\sqrt{E_b}\, b_{m'} \sum_{q=0}^{Q-1} a_{nm'}(q) h_{mn}(q) + w'_m \\ &= \sqrt{E_b}\sum_{q=0}^{Q-1} \mathbf{h}'_m(q)\left(\mathbf{H}(q)^{H^\dagger} \right)^H \mathbf{b}' + w'_m \end{aligned} \tag{8}$$

where $w'_m = \int_{-\infty}^{\infty} w_m(t)\, p(t - iT_r)\, dt$ is a zero-mean noise component with a variance of $N_0/2$ and $\mathbf{b}' = \left[\, b_1\ b_2 \cdots b_M \,\right]^T$. Here, we omit the bit index i because we are interested in particular bits of the transmitted data streams. Over each bit interval, the received spatial signal vector, $\mathbf{v} = \left[\, v_1\ v_2\ \cdots\ v_M \,\right]^T$, at the correlator output with the received noise vector, $\mathbf{w}' = \left[\, w'_1\ w'_2\ \cdots\ w'_M \,\right]^T$, is written as

$$\begin{aligned} \mathbf{v} &= \sqrt{E_b}\sum_{q=0}^{Q-1} \mathbf{H}(q)\mathbf{G}_{TxZF}^{H}(q)\, \mathbf{b}' + \mathbf{w}' \\ &= Q\sqrt{E_b}\, \mathbf{b}' + \mathbf{w}' \end{aligned} \tag{9}$$

It is evident that the ICI is completely suppressed and thus the signal components in the correlator output at the receiver are perfectly separated. To normalize the total transmission power from different transmit antenna branches for $M \le N$, the power scaling matrix of $\left(\mathbf{G}_{TxZF}(q)\mathbf{G}_{TxZF}(q)^H \right)^{-1/2} \Big/ \sqrt{Q}$ is applied to the desired signal component in the transmitter and then the sufficient statistic vector can be given by

$$\begin{aligned} \mathbf{v}' &= \sqrt{\frac{E_b}{Q}}\sum_{q=0}^{Q-1} \mathbf{H}(q)\mathbf{G}_{TxZF}^{H}(q)\left(\mathbf{G}_{TxZF}(q)\mathbf{G}_{TxZF}^{H}(q) \right)^{-\frac{1}{2}} \mathbf{b}' + \mathbf{w}' \\ &= \sqrt{\frac{E_b}{Q}}\sum_{q=0}^{Q-1} \left(\mathbf{H}(q)\mathbf{H}(q)^H \right)^{\frac{1}{2}} \mathbf{b}' + \mathbf{w}' \end{aligned} \tag{10}$$

Then, the decision variable for any information bit of the mth transmitted data stream is given as $z_m = \mathrm{Re}\{ v'_m \}$ where v'_m is the mth element of the vector $\mathbf{v}'$.

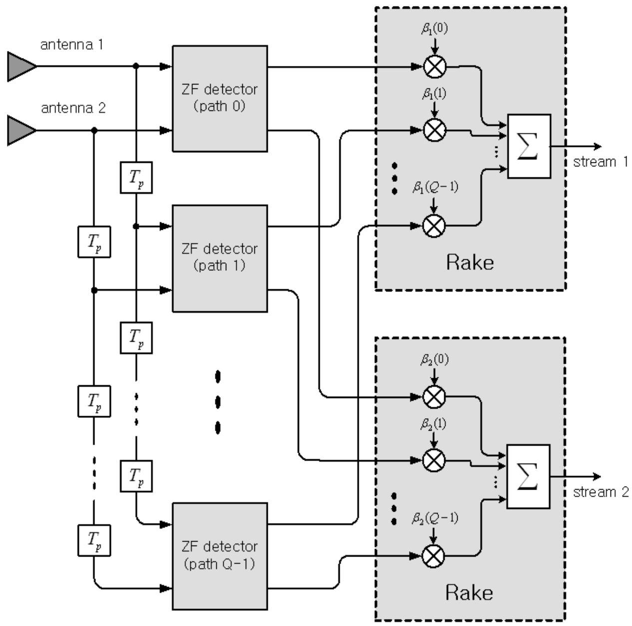

Fig. 2. Linear Receive Zero-Forcing 2×2 MIMO system with Q resolvable multipath components

III. LINEAR RECEIVE ZERO-FORCING APPROACH

In this section, the RxZF scheme examined in [2], which consists of the ZF detection followed by Rake combining as shown in Fig. 2, is briefly introduced for the purpose of comparison with the TxZF approach. It is assumed that $M \ge N$ for receive processing. The decision variable of the Rake output for a particular bit of the nth transmitted data stream is given by

$$z_n = \sum_{q=0}^{Q-1} \beta_n(q)\, y_n(q) \tag{11}$$

where

$$\beta_n(q) = \frac{1}{\lambda_n(q)} \tag{12}$$

$$\lambda_n(q) = \left[\left(\mathbf{H}(q)^H \mathbf{H}(q) \right)^{-1} \right]_{nn} \tag{13}$$

Here, $[\,\cdot\,]_{nn}$ is the nth diagonal element of main diagonal matrix. $y_n(q)$ is the nth element of the output signal vector, $\mathbf{y}(q) = \left[\, y_1(q)\, y_2(q) \cdots y_N(q) \,\right]^T$, of a ZF detector for the qth path of a particular bit, which is given by

$$\mathbf{y}(q) = \mathbf{H}(q)^\dagger\, \mathbf{x}(q) \tag{14}$$

where $\mathbf{x}(q) = \left[\, x_1(q) x_2(q) \cdots x_M(q) \,\right]^T$ is the discrete-time received signal vector at the matched filter output for the qth path and modeled by

$$\mathbf{x}(q) = \sqrt{E_b}\,\mathbf{H}(q)\,\mathbf{b} + \mathbf{w}(q) \tag{15}$$

Here, $\mathbf{b} = \left[\, b_1\; b_2\; \cdots b_N \,\right]^T$ is the information bit vector over N transmit antennas with b_n an information bit from the nth transmit antenna and $\mathbf{w}(q) = \left[w_1(q)\; w_2(q) \cdots w_M(q) \right]^T$ is the received noise vector with $w_m(q)$ a real zero-mean white Gaussian noise of variance $N_0/2$.

IV. NUMERICAL RESULTS

In the simulations, the SNR per bit in decibel is defined as

$$\eta_b = E_b/N_0\,(dB) + 10\log_{10} D \tag{16}$$

where $D = Q\,M$ and $D = Q\,N$ for RxZF and TxZF, respectively. The log-normal fading amplitude $c(q)$ can be expressed as $c(q) = e^{\psi(q)}$ where $\psi(q)$ is a Gaussian random variable with mean $\mu_{\psi(q)}$ and variance σ_ψ^2 (independent of q). We use 5-dB for the standard deviation of $20\log_{10} c(q) = \psi(q)\left(20\log_{10} e\right)$. The requirement such that the average power of the qth path has $E\left[c(q)^2 \right] = e^{-\rho q}$ is $\mu_{\psi(q)} = -\sigma_\psi^2 - \rho q/2$, where σ_ψ is given by $\sigma_\psi = 5 / \left(20\log_{10} e\right)$. Here, the power decay factor is $\rho = 0$. We assume that both the transmitter in the TxZF scheme and the RxZF receiver have perfect knowledge of the channel fading coefficients of the Q resolvable paths. In the plots, (N, M, Q) implies that N transmit antennas, M receive antennas, and Q resolvable paths are used.

Fig. 3, the BER performances of TxZF and RxZF schemes versus the number of resolvable paths, Q, with $N = 3$ and $M = 3$ as a function of SNR per bit η_b are shown. As the number of resolvable paths increases, the TxZF system is getting better performance. Comparing the BERs of TxZF with those of RxZF for the case of $N = M$, the TxZF provides significantly better performance than the RxZF. For a UWB MIMO system with $Q = 4$ resolvable paths, Fig. 4 shows the BERs of TxZF and RxZF systems for various N and M values. It is seen that when the number of transmitted multistreams (number of receive antennas) is less than or equal to the number of transmit antennas (i.e. $M \le N$) with the number of resolvable paths combined at the transmitter fixed, the TxZF systems achieve better performance as the number of transmit antennas increases. In the case of $M \le N$, the TxZF with N transmit antennas, M receive antennas, and Q resolvable paths outperforms the RxZF with M transmit antennas, N receive antennas, and Q resolvable paths over the range of low and medium SNRs. However, for high SNR values the TxZF system with high diversity offers slightly worse than the RxZF. Even if the ZF filtering is performed at the transmitter, the desired signal parts at the TxZF system's receiver still contain the ICI components as seen in the expression of (10). It is due to the effects of the power normalization process at the transmitter, which makes the matrix $\left(\mathbf{H}(q)\mathbf{H}(q)^H \right)^{1/2}$ multiplied to the signal vector $\mathbf{b}'$ to be non-diagonal.

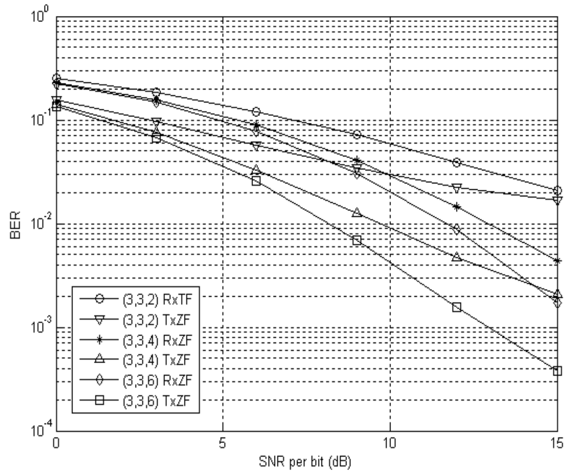

Fig. 3. BER versus SNR per bit for different values of Q in RxZF and TxZF

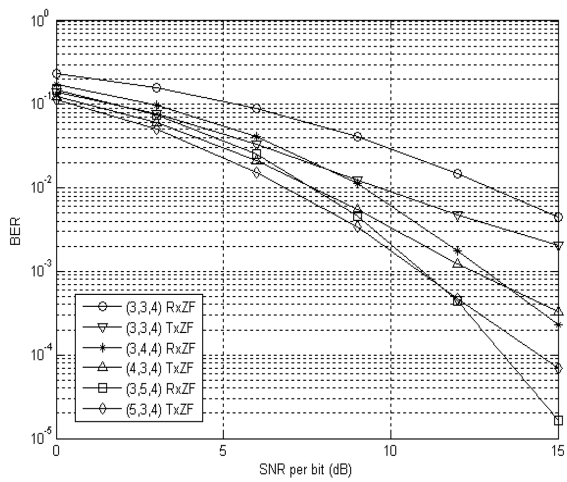

Fig. 4. BER versus SNR per bit for different values of M in RxZF and N in TxZF

V. CONCLUSIONS

The error performance of UWB SM MIMO systems based on the TDLs and linear transmit ZF processing has been presented over indoor log-normal fading channels. The UWB SM MIMO transmitter employs the transmit ZF filters to spatially separate the M multiplexed data streams and tapped delay lines that captures Q resolvable paths. It has been seen that the TxZF scheme offers much better performance than the RxZF receiver over low and medium SNR range when the number of transmitted multistreams (number of receive antennas) is less than or equal to the number of transmit antennas. Moreover, the TxZF system can significantly reduce the computational load in the receiver compared with the RxZF.

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