DETECTION OF STBC SIGNAL IN FREQUENCY SELECTIVE FADING CHANNELS

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Abstract

Transmit diversity through two-transmit and one-receive antennas (DISO) with Alamouti's Space-Time block coding (STBC) assumes the channel be flat and static over the coding block. Lindskog and Paulraj have extended this scheme for a time dispersive channel by doing the space-time coding over a large data block such as one burst. This helps to achieve diversity gain on a time-dispersive channel. However, there are two issues: 1) The receiver front-end consists of channel response filters and combining followed by a sequence detector. The channel filters on receive signal double the channel response size, consequently, increase the complexity of the sequence detector. 2) Channel needs to be static over the entire larger block of coded data as opposed to only two coded-symbol periods as per Alamouti's scheme. This paper describes a new receiver that is based on MLSE with Euclidean metric for time-dispersive channel that provides joint STBC decoding and symbol detection. The proposed receiver assumes transmission with original STBC over two-symbol blocks to overcome the above restrictions and limitations. In addition, this paper tries to evaluate the impact on complexity compared to a one-transmit and one-receive (SISO) system. This proposed solution is more suitable for narrowband systems such as GSM/EDGE.

Introduction

Diversity techniques are well known to combat multipath fading in wireless channels. Receive-antenna diversity that require many wavelengths of antenna separation for obtaining statistically independent channels are more suitable for uplink. For downlink, although studies have shown that the receive-antenna diversity is feasible with less than a wavelength of separation, it is still not preferred for small and compact terminals. Transmit diversity with Space-Time coding is being adopted by various broadband wireless systems such as WCDMA, CDMA-2000, and in various proposed forms for narrowband systems such as GSM/EDGE.

Transmit diversity scheme was originally formulated in a coding framework [1], and later trellis-based Space-Time codes that achieve both coding and full diversity gain were found [2]. Alamouti invented a simple Space-Time block code over two symbols that achieves full diversity gain [3].

Problem statement

Original transmit diversity using STBC requires the channel to be flat and static over the coding block period that is two symbols long [3]. Lindskog and Paulraj [4] extended this scheme for time-dispersive channel. According to their scheme, the space-time coding is done over two large blocks of data symbols instead of just two symbols as in originally proposed scheme. At the receiver, space-time decoding involves convolution with channel responses and combining, followed by a maximum-likelihood sequence estimator (MLSE) with an Euclidean metric. Consequently, the front-end convolution almost doubles the overall system channel response, thus increases the complexity of the MLSE detector. Also, the noise gets colored by the channel response filters that make the MLSE based detector a sub-optimal. Kambiz et. al. [5] have identified the above issues, and proposed a whitening filter followed by a Decision Feedback Sequence Estimation (DFSE) detector for lower complexity.

The above solutions for time-dispersive channels increase the complexity of the receiver due to increase in channel response by the front-end convolution. More importantly, these schemes assume the channel to be static over large block of data, usually one burst period, compared to just two symbol periods in the original scheme. Constant channel assumption over the whole burst is not valid for the high speed mobile terminals, even for GSM burst.

Due to the above reasons, this paper proposes transmitter with original space-time coding that is over every two-symbol blocks. For time-dispersive channels, an MLSE with Euclidean metric type receiver is shown for joint space-time decoding and symbol detection for maximum diversity gain. MLSE detector for SISO system can be extended for this DISO system with no or some complexity increase based on system requirements.

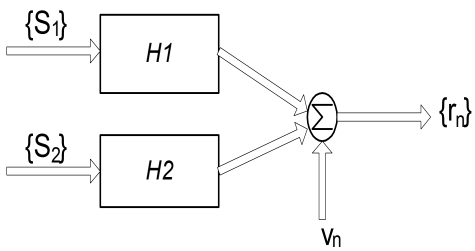

Figure 1. DISO baseband system model

System Model

In a SISO system, symbol spaced baseband received samples is given as

$$r_n = \sum_{k=0}^{L-1} h_k s_{n-k} + v_n$$

where { s_n } is the transmit symbol sequence, { h_n } is the channel impulse response of length *L*, and v_n is the sampled baseband additive white Gaussian noise.

In matrix form,

$$r_n = \overline{H}\,\overline{S}_n + v_n$$

$\overline{H}$ is channel weights row vector of size (1xL), and $\overline{S}_n$ is a column vector of size (Lx1) with the last *L* data symbols at time *n.*

For a DISO system, the receive signal is the sum of the signals from two transmitters and an additive noise, and is given as

$$r_n = \overline{H}_1 \overline{S}_{n,1} + \overline{H}_2 \overline{S}_{n,2} + v_n$$

where the suffixes *1* and *2* stand for transmitter 1 and transmitter 2, respectively.

This can also be viewed as a SISO system, especially for joint channel estimation, as

$$r_n = \overline{H}\,\overline{S}_n + v_n = [\overline{H}_1 \quad \overline{H}_2] \begin{bmatrix} \overline{S}_{n,1} \\ \overline{S}_{n,2} \end{bmatrix} + v_n$$

System Description

A burst or a block of *2N* data symbols that needs to be transmitted is

$$\overline{S} = [s_0 \quad s_1 \; s_2 \ldots\ldots s_{2N-1}]$$

After space-time coding, the transmit data symbols vector on two transmit antennas are

$$\overline{S}_{Tx1} = [s_0 \; -s_1^* \quad s_2 \; -s_3^* \ldots\ldots s_{2N-2} \; -s_{2N-1}^*]$$
$$\overline{S}_{Tx2} = [s_1 \quad s_0^* \quad s_3 \quad s_2^* \ldots\ldots s_{2N-1} \quad s_{2N-2}^*]$$

where $\overline{S}_{Tx1}$, $\overline{S}_{Tx2}$ are the coded symbol sets that are transmitted through antenna #1 and antenna #2, respectively. Note that every two symbols are space-time coded, and hence there are *N* coded blocks in the burst. For convenience, these transmit vectors are represented as *N* coded blocks as

$$\overline{S}_{Tx1} = [s_{0,0} \; -s_{1,0}^* \quad s_{0,1} \; -s_{1,1}^* \ldots\ldots s_{0,N-1} \; -s_{1,N-1}^*]$$
$$\overline{S}_{Tx2} = [s_{1,0} \quad s_{0,0}^* \quad s_{1,1} \quad s_{0,1}^* \ldots\ldots s_{1,N-1} \quad s_{0,N-1}^*]$$

Suffix *(m,n)* stands for *m'th* symbol in coded block *n*. Note that the coded block period is only two symbols long.

Time-dispersive, symbol-spaced channel vector for the link with two transmit antennas are

$$\overline{H}_{Tx1} = [h_{0,1} \; h_{1,1} \ldots\ldots h_{L-1,1}]$$
$$\overline{H}_{Tx2} = [h_{0,2} \; h_{1,2} \ldots\ldots h_{L-1,2}]$$

$\overline{H}_{Tx1}$ *and* $\overline{H}_{Tx2}$ are the discrete-time channel responses from transmit antenna #1 and #2, respectively. Suffix *(x, y)* stands for *x'th* channel tap from transmitter *y*. The channel response that is *L* taps long is the composite responses of transmit filter, channel, and receive filter, sampled at symbol rate. It is assumed that the amount of dispersion on the channel responses is the same for the two links.

At the receiver, the received signal samples at *n'th* coded block time are

$$r_{0,n} = \left[s_{0,n} \; -s^*_{1,n-1} \quad s_{0,n-1} \cdots \right] \left[h_{0,1} \; h_{1,1} \ldots\ldots h_{L-1,1} \right]^T +$$

$$\underbrace{\left[s_{1,n} \quad s^*_{0,n-1} \quad s_{1,n-1} \cdots \right] \left[h_{0,2} \; h_{1,2} \ldots\ldots h_{L-1,2} \right]^T}_{\text{signal term}} + \underbrace{v_{0,n}}_{\text{noise term}}$$

$$r_{1,n} = \left[-s^*_{1,n} \quad s_{0,n} \; -s^*_{1,n-1} \cdots \right] \left[h_{0,1} \; h_{1,1} \ldots\ldots h_{L-1,1} \right]^T +$$

$$\underbrace{\left[s^*_{0,n} \quad s_{1,n} \quad s^*_{0,n-1} \cdots \right] \left[h_{0,2} \; h_{1,2} \ldots\ldots h_{L-1,2} \right]^T}_{\text{signal term}} + \underbrace{v_{1,n}}_{\text{noise term}}$$

where

$r_{0,n}, r_{1,n}$ – samples at first and second symbol periods in coded block n

$v_{0,n}, v_{1,n}$ – filtered white noise samples

These receive samples over *n'th* coded block can be arranged in matrix form as,

$$\begin{bmatrix} r_{0,n} \\ r_{1,n} \end{bmatrix} = \begin{bmatrix} s_{0,n} & s_{1,n} \\ -s^*_{1,n} & s^*_{0,n} \end{bmatrix} \begin{bmatrix} h_{0,1} \\ h_{0,2} \end{bmatrix} +$$

$$\underbrace{\begin{bmatrix} -s^*_{1,n-1} & s^*_{0,n-1} \\ s_{0,n} & s_{1,n} \end{bmatrix} \begin{bmatrix} h_{1,1} \\ h_{1,2} \end{bmatrix} + \ldots + \overline{M} \begin{bmatrix} h_{L-1,1} \\ h_{L-1,2} \end{bmatrix}}_{\text{signal term}} + \underbrace{\begin{bmatrix} v_{0,n} \\ v_{1,n} \end{bmatrix}}_{\text{noise term}}$$

where

$$\overline{M} = \begin{cases} \begin{bmatrix} s_{0,(2n-L+1)/2} & s_{1,(2n-L+1)/2} \\ -s^*_{1,(2n-L+1)/2} & s^*_{0,(2n-L+1)/2} \end{bmatrix} & \text{if L is odd} \\ \begin{bmatrix} -s^*_{1,(2n-L)/2} & s^*_{0,(2n-L)/2} \\ s_{0,(2n-L+2)/2} & s_{1,(2n-L+2)/2} \end{bmatrix} & \text{if L is even} \end{cases}$$

Let's represent above equation as sum of signal and noise term,

$$\begin{bmatrix} r_{0,n} \\ r_{1,n} \end{bmatrix} = \begin{bmatrix} x_{0,n} \\ x_{1,n} \end{bmatrix} + \begin{bmatrix} v_{0,n} \\ v_{1,n} \end{bmatrix}$$

MLSE with Euclidean metric can be used to jointly decode both data symbols within each coded block, still preserving the diversity gain. For ($s_{0,n}$, $s_{1,n}$) hypothesized symbols set at block *n*,

$$\begin{cases} \left| r_{0,n} - \hat{x}_{0,n} \right|^2 \Rightarrow \text{Euclidean distance metric} \\ \qquad\qquad \text{for first sample in n'th block} \\ \left| r_{1,n} - \hat{x}_{1,n} \right|^2 \Rightarrow \text{Euclidean distance metric} \\ \qquad\qquad \text{for second sample in n'th block} \end{cases}$$

where

$\hat{x}_{0,n}$ *and* $\hat{x}_{1,n}$ are hypothesized signal terms using estimated channel vector

Hence,

$$metric\ (s_{0,n}, s_{1,n}) = \sum_{k=0}^{1} \left| r_{k,n} - \hat{x}_{k,n} \right|^2$$

where *metric*( ) is the Euclidean metric estimates for all possible symbol sequences corresponding to state transitions in the MLSE detector.

The first and second receive samples of a coded block are a function of *n'th* block symbols ($s_{0,n}$, $s_{1,n}$) and *(L-1)/2* previous block symbols, for *L* odd numbered channel taps. *(L-1)/2* previous blocks decide the number of trellis states that lead to m^{L-1} states where *m* is the symbol alphabet size which is the same as a SISO system. For example, GSM GMSK signal with 3 channel taps will have 4 (= 2^{3-1}) state trellis. Since the first and second samples of the coded block depend on both symbols of *n'th* block, the combined decoder will have m^2 transitions from each state. This compares with *2m* transitions over two symbol periods for a SISO system. For the above example of GSM, this gives 4 transitions from each state at each trellis stage. Hence, the overall trellis will be of four states with four transitions per state for each coded blocks (two symbols period) [Fig. 2]. If it were a SISO system, the trellis size would have been four states with two transitions per state at each symbol interval. This is equivalent to four transitions per state over two symbol intervals.

For *L* even numbered channel taps, the first receive sample of an *n'th* coded block is a function of symbols in that block and *L/2* previous blocks, since the

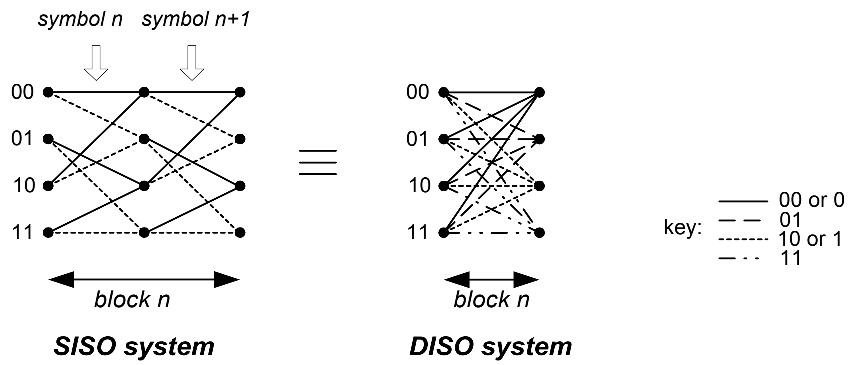

Figure 2. Trellis for a binary system with 3 channel taps

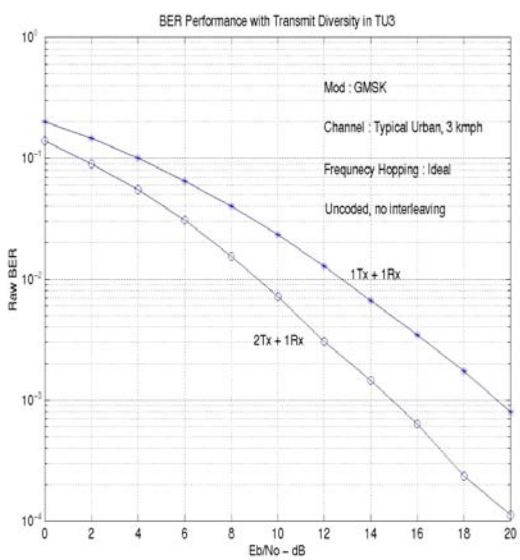

Figure 3. Raw BER (uncoded) performance of GSM GMSK signal with STBC, TU3, 5 channel taps.

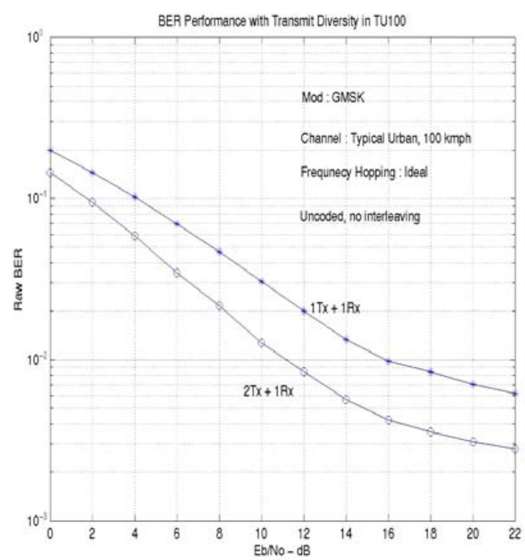

Figure 4. Raw BER (uncoded) performance of GSM GMSK signal with STBC, TU100, 5 channel taps.

decoding is done in blocks of two symbols. Hence, the trellis will be m^L states. This is equivalent to a SISO system with an additional channel tap. The number transitions per state will be the same as before. For GSM GMSK signal with $L=4$ channel taps, trellis will be of 16 states with 4 transitions per state for each coded block interval. This compares with 8 states and (2*2) transitions per state over two symbol interval for a SISO system.

To summarize, DISO system requires the same number of states as SISO system (assuming odd number of channel taps), but the number of transitions per state will be m^2 as opposed to $2m$ for SISO over the block period. This leads to an increase in complexity for EDGE signal, and no significant complexity increase for GSM GMSK signal in terms of states and transitions.

Simulated Performance

In this section, the simulation of the proposed scheme is presented. GSM GMSK burst and signal format are assumed for this simulation. GSM burst consists of 116 bits of data with 26 bits of training sequence at the middle, and 3 tail bits at the ends. No channel coding or interleaving of the data bits are assumed. The entire burst is space-time coded as explained earlier. Statistically independent channels ($\overline{H}_{Tx1}$ *and* $\overline{H}_{Tx2}$) are modeled for Typical Urban channels (TU) of GSM at 3 kmph and 100 kmph vehicle speeds. Ideal frequency hopping between bursts is assumed. At the receiver, an MLSE equalizer with 16 states and four transitions per state is used for 5 channel taps. This simulated performance is compared with non-STBC signal performance, and the results are presented in Figures 3 and 4. As can be seen, the proposed solution provides significant diversity gain compared to non-STBC signal. Notice that the gains are consistent for both low and high speed mobiles.

Conclusion

A new detection method for STBC signal through frequency selective channels is discussed. Unlike the previously proposed approaches given in [4] and [5] where the channel is assumed to be constant over the whole GSM burst, the proposed scheme uses the originally proposed STBC over two symbol periods with extended MLSE based receiver structure. The complexity of the proposed receiver is evaluated and compared with SISO system. It is shown that employing STBC in frequency selective fading channels with the proposed approach is practical, and it works well even for high speed mobiles.

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