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A novel weighted multilevel space–time trellis coding scheme

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ABSTRACT

It has been shown that when perfect or partial channel state information (CSI) is available at the transmitter, the performance and capacity of a space–time coded system can be further improved. Multilevel space–time trellis codes (MLSTTC's) are capable of simultaneously providing bandwidth efficiency, diversity improvement and coding gain with significantly reduced decoding complexity, especially for larger constellations and higher throughputs. In this paper, we present a design of combined multilevel space–time trellis codes and beam forming based on feedback from receiver, henceforth referred to as weighted multilevel space–time trellis codes (WMLSTTCs). The channel profile is used to provide a beam forming scheme that achieves the better performance by properly weighting transmitted signals. Weights are selected that matches best with a channel profile feedback from the receiver indicative of long-term characteristics of the wireless channel. WMLSTTCs provide improvement in performance of MLSTTCs.

1. Introduction

During the last decade, there has been a steady increase in the demand of high data rates that are to be supported by wireless communication applications. Recently, there has been a trend towards use of multilevel signals in order to increase transmission speed. However, multilevel signals are more likely to be corrupted by channel noise. Thus a digital communication system using multilevel signals generally requires an error-correcting scheme. Several multilevel coding methods using error-correcting have been previously suggested. However, an efficient system generally could not be constructed using a binary error-correcting code. Hideki Imai and Shuji Hirakawa [1] proposed a new multilevel coding method that used several error-correcting codes. The transmission symbols are constructed by combining binary error-correcting codes having different error-correcting capabilities to construct an efficient system.

The ability to detect and/or correct errors is provided by the additional transmission of redundant bits, and thus by lowering the effective information rate per transmission bandwidth. Gottfried Ungerboeck [2] solved this problem by providing a coding technique which improves error performance without sacrificing data rate or requiring more bandwidth by coding channel with expanded sets of multilevel/phase signals and Soft maximum-likelihood (ML) decoding using the Viterbi algorithm. The technique works by partitioning the signal constellation into a multilevel hierarchy and defining a code over each level and maximizing the minimum intra-subset Euclidean distance. The binary partitions of Ungerboeck have been generalized by Calderbank [3] which permits multidimensional signal sets to be conveniently partitioned by extending the multilevel coding method to coset codes and to calculate minimum squared distance and path multiplicity in terms of the norms and multiplicities of the different cosets. The multilevel structure allows the redundancy in the coset selection procedure to be allocated efficiently among the different levels. Pottie and Taylor [4] proposed a hierarchy of codes to match the geometric partitioning of a signal set and combinations of such codes in a multilevel scheme to reduce coding complexity.

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Motivated by the work of Imai and Hirakawa [1], and by Calderbank [2], Nambi Seshadri [5,6] proposed 8-phase shift keyed multilevel code constructions for the Rayleigh fading channel utilizing maximum free Hamming distance binary convolutional codes as building blocks. A new multistage decoding procedure that makes use of interleaving/deinterleaving resulting in a decoder that is robust to error bursts of decoding from previous stages and provides unequal error protection can be obtained in a flexible manner by choosing the component codes appropriately.

The combined coding and modulation introduced by Ungerboeck consists of using an equi-resolution modulation and a unique code to increase the Euclidean distance between two transmitted sequences [4]. Khaled Fazel [6] proposed combining multilevel coding and multiresolution modulation, and tried to optimize the different parameters of such a combination under different conditions in a Rician fading channel. The constellation of an MRM consists of clusters of points spaced by different distances. Each cluster may itself have sub-clusters, and so on. The distance between two clusters is higher than the distance between two sub-clusters. Then the basic idea is to assign the most significant information bits to the clusters and the less significant information bits to the sub-clusters.

Among the multilevel coding that has been proposed by the research community to cope with this new demand, the utilization of multiple antennas arises as one of the best candidates due to the fact that it provides both an increase in reliability and also in information transmission rate. Combining multiple antennas both at the transmitter and receiver ends gives rise to the so called multi-input multi-output (MIMO) channels. For MIMO systems, many codes have been devised for increasing the information transmission rate. More specifically, space–time codes (STCs) have been introduced by Tarokh [7] to provide improved error performance and increasing information transmission rate. Space–time block codes (STBC) operate on a block of input symbols producing a matrix output whose columns represent time and rows represent antennas providing full diversity with extremely low encoder/decoder complexity. Space–time trellis codes operate on one input symbol at a time producing a sequence of vector symbols whose length represents antennas and provides diversity and coding gain. Ever since Tarokh [7] has proposed design criteria for space–time trellis codes (STTCs), a number of codes [8–12] with improved performance have been presented.

The space–time coding schemes mentioned above do not exploit the channel knowledge at the transmitter. However, if CSI is made available at the transmitter, a beam forming scheme can provide the same diversity order as well as significantly more array gain [13,14] than the traditional space–time codes. A combined beam forming and space–time block code was proposed in [15,16], while combined beam forming and space–time trellis codes were developed in [17–19]. It has been shown that such schemes can attain both coding gain and beam forming gain, thus improving system performance dramatically. Li and Vucetic [20] evaluated the performance of space–time turbo trellis codes (STTTC) combined with ideal beam forming over slow fading channels. Simulation results exhibit the significant performance improvement of the proposed scheme. Jöngren [21] presented the code design using side information in the form of channel estimates at the transmitter in conjunction with certain space–time codes. The transmission scheme combines the benefits of transmit beam forming and orthogonal space–time block coding.

Yonghui Li and Branka Vucetic [22] present the design of combined space–time trellis codes and beam forming to design weighted space–time trellis codes (WSTTCs). It was shown that WSTTCs can achieve not only a full spatial diversity order, but also an additional time diversity order as well as an additional received SNR gain, referred to as the weighting gain and a significant coding gain, relative to the standard STTC.

Tarokh et al. [7] combined multilevel codes (MLCs) or coset codes in a space–time (multiple transmit antenna) environment. MLCs are concatenated with orthogonal space–time block codes (OSTBCs). They consider various partitioning schemes, including Ungerboeck set partitioning, mixed partitioning and block partitioning. The concatenation of multilevel codes and a space–time block code (STBC) can achieve full diversity gain and improved coding gain. In the traditional MLC designs [23,24], the code rates of component codes are determined by the capacity rule. Shang-Chih Ma [25] proposed a multilevel concatenated STBC scheme by combing component codes in conjunction with set partitioning of the expanded signal constellation such that the CGD at each level is maximized to achieve better error performance in comparison with the traditional MLC scheme.

Baghaie [26] developed multilevel space–time trellis codes (MLSTTC's) by combining multilevel coding, STTC and multiresolution modulation (MRM) which is capable of simultaneously providing bandwidth efficiency, diversity improvement and coding gain with significantly reduced decoding complexity, especially for larger constellations and higher throughputs. Baghaie [27] designed grouped multilevel STTCs that can provide the high throughput of multi-layered schemes while realizing larger diversity gains.

In this paper we extend the work of Baghaie [26] in which MLSTTC was designed by assuming perfect CSI available at receiver. A novel design criterion for MLSTTC with CSI at transmitter is proposed, that incorporates the statistical information concerning the channel estimates. New improved Weighed MLSTTC codes are obtained using a novel combination of beam forming and MLSTTC. We show that Weighed MLSTTC codes provide improved performance than MLSTTC. This paper is organized as follows. In Section 2, we present a system model for Weighed MLSTTC codes. In Section 3 performance analysis of Weighed MLSTTC codes is provided. Finally, some concluding remarks are provided in Section 4.

2. System model

A block diagram of a weighted MLSTTC system is presented in Fig. 1. For simplicity, we consider N transmit antennas and only one receive antenna.

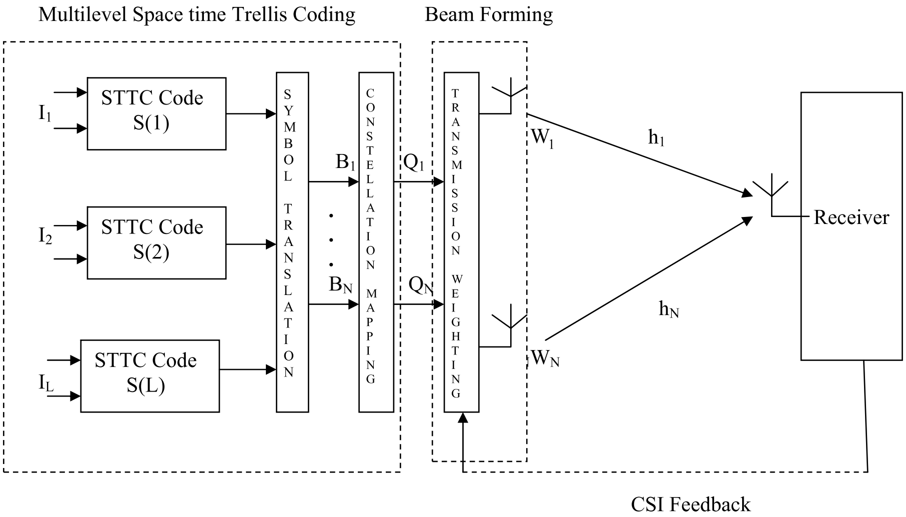

Fig. 1. Block diagram of weighted MLSTTC system.

The information bits at each time t are encoded first by the space–time trellis encoder designed for the QAM constellation. These generate N 4-QAM symbols per time slot per STTC code. The STTC codes denoted by $S(1)\cdots S(L)$ are given by:

$$S(1) = x_t(1)^1, x_t(1)^2, \ldots, x_t(1)^N \tag{1}$$

$$S(L) = x_t(L)^1, x_t(L)^2, \ldots, x_t(L)^N. \tag{2}$$

Symbols in $S(1)$ are used to determine the most significant bits of the new QAM symbols. Similarly the N symbols generated by $S(L)$, are used to determine the least significant bits.

The output of space–time trellis code is fed into a symbol translator, which generates N new symbols. These N new QAM symbols are denoted

$$B(1) = x_t(1)^1, x_t(2)^1, \ldots, x_t(L)^1 \tag{3}$$

$$B(N) = x_t(1)^N, x_t(2)^N, \ldots, x_t(L)^N. \tag{4}$$

These N new QAM symbols are then input into constellation mapper to map M-QAM constellation.

The MLSTTC system works by partitioning the underlying signal constellation using the multi-resolution modulation (MRM) approach. The constellation of a MRM consists of clusters of points spaced by different distances. Each cluster may itself have sub-clusters, and so on. The distance between two clusters is higher than the distance between two sub-clusters. Then the basic idea is to assign the most significant information bits to the clusters and the less significant information bits to the sub-clusters. Ultimately, the last bits choose a signal point within the underlying constellation.

Fig. 2 illustrates an MRM with 64-points. The 64 points in the constellation are first divided into four clusters and then each cluster itself consists of 4 sub-clusters. Each sub-cluster is made up of four points. This clusterization allows one to have three resolutions: the most significant bits are mapped to the clusters, the middle significant bits to sub-clusters and the least significant bits to the points of the sub-clusters. The performance of these resolutions depend strongly on the distances d_1, d_2 and d_3. The performance of such a modulation can be estimated easily by averaging over all points of the clusters. The basic idea is to use different codes for each bit assigned to each partition level.

Fig. 3a shows the application of MRM to a 16-QAM constellation and that is used for set portioning of 64-QAM as shown in Fig. 3b. As can be seen, the 64 points in the underlying constellation are divided into 4 clusters and each cluster into 4 sub-clusters and so forth. Thus, we use a 4-cluster as the basic unit of resolution. We partition [68] the multi-resolution constellation by treating each cluster as a 4-QAM constellation. This enables us to directly use CSTTCs designed for 4-QAM in our mapping process. The labeling of the signal constellation points based on the partitioning is also shown in Fig. 3, where we have 3 clusters, each having 4 sub-clusters. The circles in the figure denote one sub-cluster of each cluster.

This clusterization allows us to have L resolutions for a M-QAM constellation, with $M = 4L$. We then map the output of the first STTC $S_m(1)$, to the clusters and the output of the next component code $S_m(2)$ to the sub-clusters and so forth with the output of $S(L)$ selecting the actual constellation points to be transmitted. For the 64-QAM constellation of Fig. 3, this results in $L = 3$ levels.

The output of $S(L)$, denoted as $x_t(L)$, gets mapped to the actual constellation points, while the outputs generated by $S(1)$ to $S(L-1)$, denoted $x_t(1), \ldots, x_t(L-1)$ respectively, get mapped to the virtual cluster center points (centroids) as shown in Fig. 2, and thus label the corresponding clusters. These STTC are all designed for a 4-QAM constellation and their coded output symbols, $x_t(1)$ to $x_t(L)$, are all drawn from 4-QAM constellations and can be represented in complex form as

$$x_t(l) = a + jb \quad a, b\varepsilon\{1, -1\}. \tag{5}$$

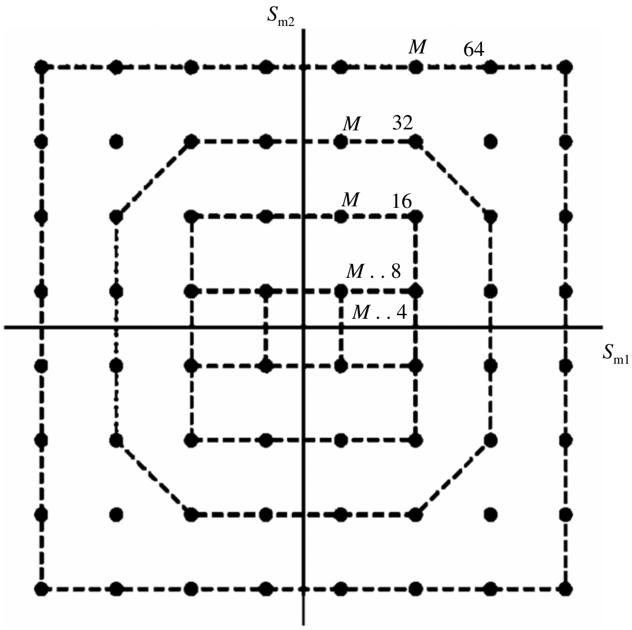

Fig. 2. MRM with 64 points.

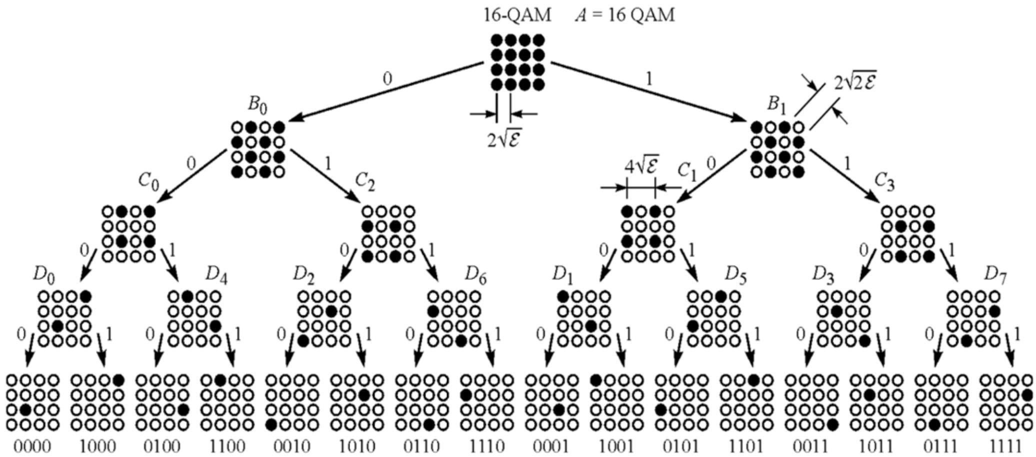

Fig. 3a. 16-QAM set partitioning using convolution code.

Each point in the underlying M-QAM constellation can be represented in complex form as

$$Q = d_{x(1)}x(1) + d_{x(2)}x(2) + \cdots + d_{x(L)}x(L) \tag{6}$$

where $d_{x(1)}, \ldots, d_{x(L)}$ are the cluster distances corresponding to $x_t(1), \ldots, x_t(L)$, as shown in Fig. 2.

The output of constellation mapper is then weighted by weighting parameter based on the feedback channel information from the receiver.

The symbol transmitted at time t by the jth transmit antenna is denoted by $Q_t j$, for $1 = j = N$. We assume

- the channel exhibits quasi-static frequency flat Rayleigh fading over a frame duration. Thus, it is constant over one frame and varies independently between frames.
- Perfect CSI is available at the receiver and transmitter.

In this paper we use beam forming coefficients as

$$w_i(t) = h_t{}^* \left(\sum_{j=1}^{n} \left| h_j \right|^2 \right)^{-1/2}. \tag{7}$$

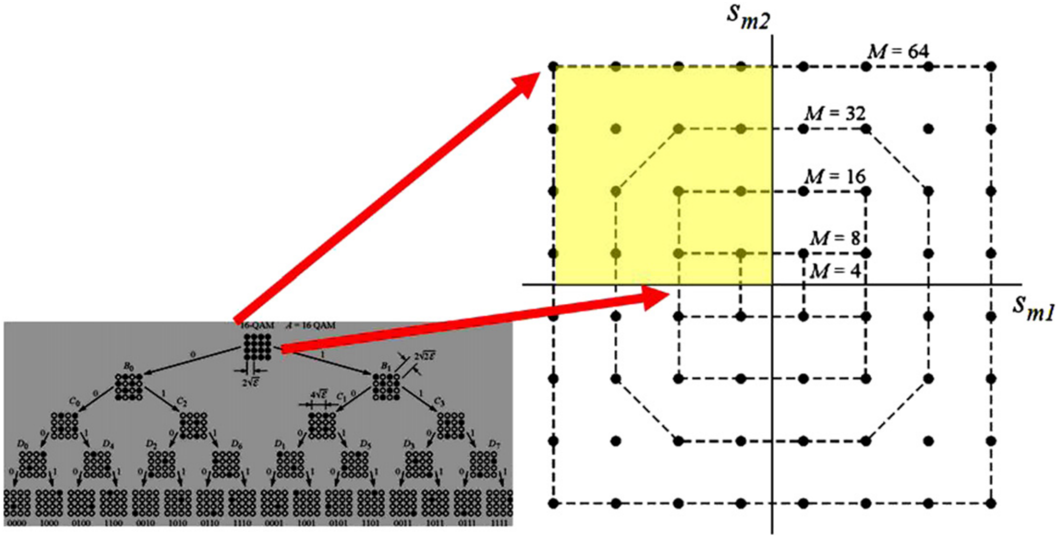

Fig. 3b. 64-QAM set partitioning.

It also satisfies the power normalization constraint

$$\sum_{i=n}^{n} |w_i|^2 = 1.$$

At time t, the symbol transmitted from the jth transmit antenna can be represented by

$$Qj^t = w_j\{d_{x(1)}x_t(1)^j + d_{x(2)}x_t(2)^j + \cdots + d_{x(L)}x_t(L)^j\} \quad j = 1, \ldots, N \tag{8}$$

where $x_t(l)^j$ is the 4-QAM symbol generated by the lth STTC code. w_j is weight for jth antenna.

3. Decoding of WMLSSTC

The decoder starts by decoding the output of the Lth component code. The estimated values of $x(L)$, $\hat{x}(L)$, are then passed to the next decoding stage and are used to decode the values of $x(L-1)$ and so forth. The final stage of the decoder uses the estimates obtained from levels L to 2, namely $\hat{x}(L), \hat{x}(L-1), \ldots, \hat{x}(2)$ to obtain $\hat{x}(1)$. We derive decoding metrics for the proposed multi-stage decoder. The derivation is carried out for an arbitrary number of transmit and one receive antenna. We start by looking at the form of the received signal. Based on (6) and (9), the received signal at the receive antenna at time t is given by

$$r_t = \sum_{l=1}^{L} \sum_{j=1}^{N} h_j^t d_{x(l)} x_t(l)^j W_j + n_t. \tag{9}$$

The conditional probability density function (pdf), of r_t conditioned on the channel matrix and all L encoder outputs, may then be written as

$$P\left(r_t | x_t(1), \ldots, x_t(L), H_t\right) \tag{10}$$

where $x_t(l) = (x_t(l)^1, x_t(l)^2, \ldots, x_t(l)^N)$, for $l\varepsilon\{1, 2, \ldots, L\}$.

We apply a multi-stage decoder, where $x_t(L)$ is decoded first and is passed to the next stage, where $x_t(L-1)$ is decoded and so forth.

3.1. Decoding of stage L

In the Lth stage, the aim is to decode $x_t(L)$. We make use of the Viterbi algorithm, and thus hypothesize the value of vector $x_t(L)$. The decoder in Stage L, performs a search to maximize the likelihood function over the hypnotized values of $x_t(L)$. The values of $x_t(1)$ to $x_t(L-1)$ are unknown at this stage and thus we treat them as "nuisance" variables and average them out.

$$P\left(r_t | x_t(L), H_t\right) = \sum_{\substack{x_t(l),\\ l=1,\ldots,L-1}} P(x_t(1), \ldots, x_t(L-1))|x_t(L), H_t P\left(r_t | x_t(1), \ldots, x_t\left(L-1\right), H_t\right). \tag{11}$$

Since the component encoders are assumed to be independent of each other, we consider $x_t(1), \ldots, x_t(L-1), x_t(L)$ to be mutually independent. Considering that the channel is also assumed to be independent of the component codes, and that different values of $x_t(l)$ have the same probability of being transmitted, the probability $P(x_t(1), \ldots, x_t(L-1)|x_t(L), H_t)$ in the expression on the right hand side of (10) reduces to a constant and can be ignored when maximizing the likelihood function. Now focusing on the second term, on the right hand side of (10) we can write

$$P\left(r_t|x_t(1), \ldots, x_t\left(L-1\right), x_t(L), H_t\right) = \frac{1}{(\sqrt{2\pi}\sigma_n{}^2)} \exp\left[\frac{\left|r_t - \sum_{l=1}^{L}\sum_{j=1}^{N} h_t^j d_{x(l)} x_t(l)^j w_j\right|^2}{2\sigma_n{}^2}\right]. \tag{12}$$

Substituting (9) into (10) and ignoring the constant term, we obtain the likelihood function in the form

$$L\left(x_t(L)\right) = P\left(r_t|x_t(L), H_t\right) \propto \sum_{\substack{x_t(l),\\ l=1,\ldots,L-1}} \exp\left[\frac{\left|r_t - \sum_{l=1}^{L}\sum_{j=1}^{N} h_t^j d_{x(l)} x_t(l)^j w_j\right|^2}{2\sigma_n{}^2}\right] \tag{13}$$

$$\log \sum_{\substack{x_t(l),\\ l=1,\ldots,L-1}} \exp\left[\frac{\left|r_t - \sum_{l=1}^{L}\sum_{j=1}^{N} h_t^j d_{x(l)} x_t(l)^j w_j\right|^2}{2\sigma_n{}^2}\right] \tag{14}$$

as the corresponding branch metric. As mentioned previously, Stage L decodes the subset labels $x(L)$ using the Viterbi algorithm. More specifically, assuming that r_t is the received signal at the receive antenna at time t, the branch metric for a transition labeled $x_t(L)^1, x_t(L)^2, \ldots, x_t(L)^j$ is given by (13). The Viterbi algorithm is used to compute the path with the largest accumulated metric over the duration of a data frame.

The decoded values of $x(L)$, denoted $\hat{x}(L)$, are assumed to be correct and are passed on to the decoder of Stage $L-1$ and used in decoding $x(L-1)$ and so forth. In the next section we will look at stage k of decoding for $1 < k < L$.

3.2. Decoding of intermediate stage k

The decoding of the kth stage is similar to the decoding of stage L. Now the aim is to decode $x_t(k)$. We make use of the Viterbi algorithm and hypothesize the value of vector $x_t(k)$. Similar to what we had before, the decoder performs a search to maximize the likelihood function over the hypothesized values of $x_t(k)$. This time however, assuming the decoding starts from stage L, and $1 < k < L$, the outputs of the stage L to $k+1$ decoders (i.e. $\hat{x}(L), \ldots, \hat{x}(k+1)$) are available. Therefore, we can take these decisions into account.

The values of $x_t(1), \ldots, x_t(k-1)$ are still unknown at this stage and thus are treated as"nuisance" variables and averaged out. The branch metric is derived using the same procedure as before. For a transition labeled $x_t(k)^1, x_t(k)^2, \ldots, x_t(k)^j$ the branch metric at stage k, where $1 < k < L$, can be calculated as

$$\log\left(\sum_{\substack{x_t(l),\\ l=1,\ldots,k-1}} \exp\left[\frac{\left|r_t - \sum_{l=1}^{k}\sum_{j=1}^{N} h_t^j d_{x(l)} x_t(l)^j w_j - \sum_{p=k+1}^{L}\sum_{j=1}^{N} h_t^j d_{x(p)} \hat{x}_t(p)^j w_j\right|^2}{2\sigma_n{}^2}\right]\right). \tag{15}$$

The appearance of the decisions $\hat{x}_t(p);\ p = k+1, \ldots, L$ in the third term of the exponent. We next look at the final stage of decoding where the estimated values of $x(L), \ldots, x(2)$ are available to the decoder.

3.3. Decoding of stage 1

Stage 1 is the final stage of decoding, where we use the Viterbi algorithm to decode the STTC generated by the $S(1)$ encoder. Following the same procedure as for decoding of stage L, the branch metrics for

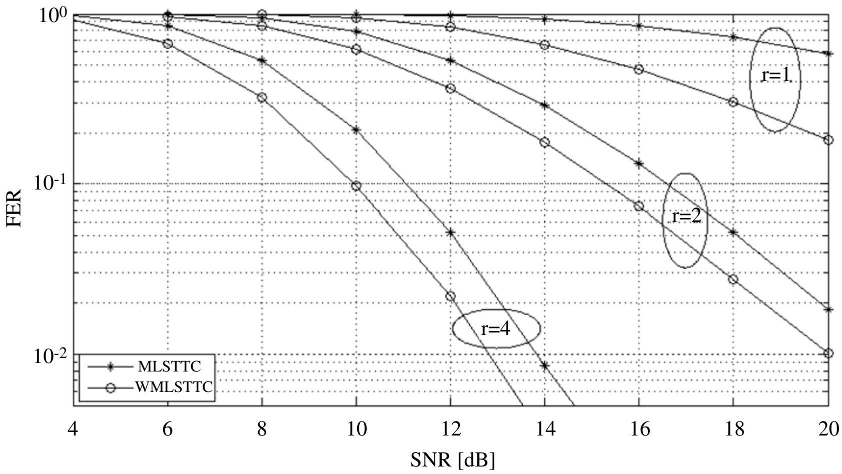

Fig. 4. Comparison of simulation results for MLSTTC and proposed WMLSTTC.

Stage 1 with transition labeled $x_t(1)^1, x_t(1)^2, \ldots, x_t(1)^j$ is given by

$$\log\left(\sum_{\substack{x_t(l),\\ l=2,\ldots,L}} \exp\left[\frac{\left|r_t - \sum_{l=1}^{L}\sum_{j=1}^{N} h_t^j d_{x(l)} x_t(l)^j w_j\right|^2}{2{\sigma_n}^2}\right]\right). \tag{16}$$

The decoded (estimated) values of $x(L), \ldots, x(2)$, denoted by $\hat{x}(L), \ldots, \hat{x}(2)$, are now available and are passed to the Stage 1 decoder, as shown in Fig. 4(1). Therefore, we no longer average over $x(L), \ldots, x(2)$ but instead insert the corresponding decisions $\hat{x}(L), \ldots, \hat{x}(2)$, denoted by $\hat{x}(l)^j$, directly into the above expression. Doing so eliminates the summation over $\hat{x}(l)$ and allows (13) to be reduced to

$$\left|r_t - \sum_{j=1}^{N} h_j^t \left(d_{x(1)} x_t(1)^j w_j - \sum_{l=2}^{L} d_{x(l)} \hat{x}(l)^j\right)\right|^2. \tag{17}$$

Simulation results:

In this section, we consider a simple WMLSTTC system with two transmit ($n = 2$) and one receive antenna as shown in Fig. 2. It uses L (=2) level coding and set partitioning to partition a 16-QAM constellation 2 times using 4-way partitions into subset of constellation points [4]. Each level uses a 4-state 4-QAM STTCs designed using the trace criterion [2]. At time t, the output of the Lth level code, denoted by $\mathbf{X}_t(L) = x_t(L)^j | 1 \le j \le n$, selects the subset of constellation points on the Lth partitioning level [4]. The 16-QAM symbol at time t at the stream j, denoted Q_t^j, is collectively defined by the 4-QAM symbols from the 2-levels as

$$Q_t^j = d_1 x_t(1)^j + d_2 x_t(2)^j \tag{18}$$

where d_1, d_2 are the subset distances corresponding to $x_t(1)^j, x_t(2)^j$ (for all) [4]. Finally, the weighted 16-QAM symbol transmitted through antenna j at time t is given by, $c_t^j = w_t^j Q_t^j$.

At the receiver, a 2-stage decoder starts at stage 2 by decoding the level 2 code. The decision $\hat{\mathbf{X}}_t(2)$ on $\mathbf{X}_t(2)$ is passed to the stage 1 to decode the values of $\mathbf{X}_t(1)$. Each stage uses the Viterbi algorithm for decoding. The branch metric is created using a max–log approximation to the likelihood function [4].

We present simulation results for WMLSTTC system shown in Fig. 2 with two transmit and r (=1, 2, 4) receive antennas, over a quasi-static Rayleigh fading channel, for which the fading coefficients are constant within one frame but vary independently from one frame to another. We used $d_1 = 2$, $d_2 = 1$ and a frame size of 130 symbols. We assume that the CSI is perfectly known at both the transmitter and the receiver. Each STTC provides a throughput of 2 bits/s/Hz, resulting in an overall throughput of 4 bits/s/Hz.

For comparison purposes, we have used an MLSTTC system with the same specifications as the above WMLSTTC system but without weighting.

Fig. 4 exhibits the frame error rate (FER) performance of WMLSTTC and MLSTTC system plotted against signal to noise ratio (SNR). It can be noted that the performance of the MLSTTC system is dramatically improved by the weighting. It can be seen that for two transmit and two receive antennas WMLSTTC is superior to MLSTTC by about 1.3 dB at the FER of 10^{-1} and for two transmit and four receive antennas WMLSTTC is superior to MLSTTC by about 1 dB at the FER of 10^{-2}.

4. Conclusion

This paper has shown that if perfect channel state information (CSI) is available at the transmitter, the performance of a space–time coded system can be further improved by weighting the transmitted signals. In this reported work, we have evaluated the performance of multilevel space–time trellis codes (MLSTTCs) combined with ideal beam forming over slow fading channels. Simulation results showed that the proposed scheme has considerably outperformed the conventional MLSTTCs without weighting.

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