Abstract: The 2 × 2 MIMO profiles included in Mobile WiMAX specifications are Alamouti’s space-time code (STC) for transmit diversity and spatial multiplexing (SM). The former has full diversity and the latter has full rate, but neither of them has both of these desired features. An alternative 2×2 STC, which is both full rate and full diversity, is the Golden code. It is the best known 2×2 STC, but it has a high decoding complexity. Recently, the attention was turned to the decoder complexity, this issue was included in the STC design criteria, and different STCs were proposed. In this thesis, a full-rate full-diversity 2 × 2 STC design leading to substantially lower complexity of the optimum detector compared to the Golden code with only a slight performance loss. We provide the general optimized form of this STC and show that this scheme achieves the diversity-multiplexing frontier for square QAM signal constellations. Then, we present a variant of the proposed STC, which provides a further decrease in the detection complexity with a rate reduction of 25% and show that this provides an interesting trade-off between the Alamouti scheme and SM.

1. INTRODUCTION

Multiple-Input Multiple-Output (MIMO) concepts have been under development for many years for both wireless and wire-line systems. One of the earliest MIMO-to-wireless communications applications came in 1984 with groundbreaking developments by Jack Winters of Bell Laboratories. Digital communication using MIMO or also called volume to volume wireless links is emerging as one of the most promising research areas in wireless communications. The MIMO approach can also be used for spatial diversity rather than multiplexing to benefit the Bit Error Rate (BER) performance of the wireless communication link. A full-rate full-diversity 2 × 2 STC design leading to substantially lower complexity of the optimum detector compared to the Golden code with only a slight performance loss.

Overview

In wireless MIMO the transmitting end as well as the receiving end is equipped with multiple antenna elements, as such MIMO can be viewed as an extension of the very popular ‘smart antennas’. In MIMO though the transmit antennas and receive antennas are jointly combined in such a way that the quality (Bit Error Rate) or the rate (Bit/Sec) of the communication is improved.

At the system level, careful design of MIMO signal processing and coding algorithms can help increase dramatically capacity and coverage and thus can improve the economics of network deployment for operators. Today, MIMO wireless is widely recognized as one of three or four key technologies in the forthcoming high-speed high-spectrum efficiency wireless networks (4G, and to some extent 3G). Applications also exist in fixed wireless and wireless Local Area Network (LAN) networks. This MIMO pioneer described ways to send data from multiple users on the same frequency/time channel using multiple antennas at the transmitter and receiver. Multiple-input/multiple-output (MIMO) technology offers tremendous performance gains for wireless systems at relatively low cost. Any system with multiple inputs into the receiver and multiple outputs to the transmitter is a MIMO system.

Wireless system designers are facing a number of challenges. These include the limited availability of the radio frequency spectrum and a complex space–time varying wireless environment. In addition, there is an increasing demand for higher data rates, better quality of service, and higher network capacity. In recent years, MIMO systems have emerged as a most promising technology in these measures. The most effective technique to accomplish reliable
communication over a wireless channel is diversity which attempts to provide the receiver with independently faded copies of the transmitted signal with the hope that at least one of these replicas will be received correctly.

One of the major advantages of MIMO systems is the substantial increase in the channel capacity, which immediately translates to higher data throughputs. Another advantage of MIMO systems is low bit error rates. These advantages are achievable without any expansion in the bandwidth or increase in the transmit power. Since the information is transmitted through different paths, a MIMO system is capable of exploiting transmitter and receiver diversity, hence maintaining reliable communications.

2. 2X2 FRFD STC

Code Design Criteria

A finite set of complex matrices is a STBC. A n×n linear STBC is obtained starting from an n × n matrix consisting of arbitrary linear combinations of k complex variables and their conjugates, and letting the variables take values from complex constellations. The rate of such a code is k/n complex symbols per channel use. We consider Rayleigh quasi-static flat fading MIMO channel with full channel state information (CSI) at the receiver but not at the transmitter. For 2×2 MIMO transmission, we have

\[ Y = HS + N \]  \hspace{1cm} (1)

**Code rate**

If there are k independent information symbols in the codeword which are transmitted over T channel uses, then, for an n×n MIMO system, the code rate is defined as k/T symbols per channel use. If k = n_{min}T, where n_{min} = min(n_t,n_r), then the STBC is said to have full rate.

Considering ML decoding, the decoding metric that is to be minimized over all possible values of codewords S is given by

\[ M(S) = ||Y - HS||^2 \]  \hspace{1cm} (2)

for, number of symbols = 4, Time slots =2, (4/2) = 2 symbols for channel used and with 2 transmit, 2 receive antennas n_{min} = min(n_t,n_r) = 2, shows full rate.

**Decoding complexity**

The ML decoding complexity (number of metrics) is given by the minimum number of symbols that need to be jointly decoded in minimizing the decoding metric. This can never be greater than k, in which case, the decoding complexity is said to be of the order of M^k. If the decoding complexity is lesser than M^k, the code is said to admit simplified decoding. If, constellation size is 16 then metrics computed are (16^4) = 256.

To achieve maximum diversity, the codeword difference matrix (X−^\hat{X}) must be full rank for all possible pairs of codewords and the diversity gain is given by n_t n_r.

**The Golden Code**

This is a full-rate and full diversity 2×2 linear dispersion algebraic space-time code with unprecedented performance based on the Golden number (1+\sqrt{5})/2.

\[ X = \begin{bmatrix} \alpha(s_1 + s_2 \theta) & \alpha(s_3 + s_4 \theta) \\ \overline{\alpha}(s_3 + s_4 \overline{\theta}) & \overline{\alpha}(s_1 + s_2 \overline{\theta}) \end{bmatrix} s_1,s_2,s_3,s_4 \in Z[i] \]  \hspace{1cm} (3)

where \( \alpha = 1+i(1-\theta), \theta = (1+\sqrt{5})/2 \) and \( \overline{\theta} = (1-\sqrt{5})/2 \) and we assume the signal constellation S to be a 2^b-QAM, with in-phase and quadrature components equal to ±1,±3, . . . and b bits per symbol.

\[ M(s_1,s_2,s_3,s_4) = ||Y - HX||^2 \]  \hspace{1cm} (4)

As four symbols are transmitted in two time slots, it is a full rate code. This STBC seems to be a combination of two Alamouti STC. For M constellation size and 4 symbols, ML exhaustive search takes M^4 metrics to be computed.
3. PROPOSED - LOW COMPLEXITY FRFD STC

Now, we present our approach to full-rate 2x2 STC design which attempts to maximize both the diversity gain and the coding gain, while leading to an optimum detector of reduced complexity. More specifically, the proposed STC is a full-rate, full-diversity 2x2 space-time code whose optimum receiver has a complexity that is only proportional to M^2, where M is the size of the signal constellation. Thus, the number of Euclidean distance computations in the optimum detector is reduced to 16^2 = 256 for a 16-QAM signal constellation and to 64^2 = 4,096 for a 64-QAM signal constellation. Comparing these numbers to those associated to the Golden code (or Matrix C), it becomes clear that this code makes the implementation of full-rate, full-diversity 2x2 STCs with optimum receiver realistic. We now give a general description of the proposed code. A group of 4 data symbols (s_1, s_2, s_3, s_4) in the proposed code design is transmitted as follows:

\[
X_{\text{new}} = \begin{bmatrix}
    a s_1 + b s_3 \
    a s_2 + b s_4 \
    -c s_2^* - d s_4^* \
    c s_1^* + d s_3^*
\end{bmatrix}
\]  

(5)

Where \(a, b, c, \) and \(d\) are complex-valued design parameters and the star designates complex conjugate.

In this matrix representation, the first column represents the symbol combinations transmitted during a first symbol interval t1 and the second column represents the symbol combinations transmitted during a second symbol interval t2. The first row of the matrix gives the symbol combinations transmitted from the first Tx antenna, and second row of the matrix gives the symbol combinations transmitted from the second Tx antenna. In other words, \(a s_1 + b s_3\) is transmitted from Tx antenna 1 during the first symbol interval t1, \(a s_2 + b s_4\) is transmitted from Tx antenna 2 during the first symbol interval t1, \(-c s_2^* - d s_4^*\) is transmitted from Tx antenna 1 during the second symbol interval t2, and \(c s_1^* + d s_3^*\) is transmitted from Tx antenna 2 during the second symbol interval t2.

On the first receive antenna, the two signals received at the first and second symbol intervals are:

\[
r_1 = h_{11} (a s_1 + b s_3) + h_{12} (a s_2 + b s_4) + n_1,
\]

(6)

\[
r_2 = h_{11} (-c s_2^* - d s_4^*) + h_{12} (c s_1^* + d s_3^*) + n_2,
\]

(7)

Similarly, we have on the second Rx antenna:

\[
r_3 = h_{21} (a s_1 + b s_3) + h_{22} (a s_2 + b s_4) + n_3,
\]

(8)

\[
r_4 = h_{21} (-c s_2^* - d s_4^*) + h_{22} (c s_1^* + d s_3^*) + n_4,
\]

(9)

Where \(n_i, \) for \(i = 1...4,\) are the additive noise terms.

The maximum likelihood (ML) detector makes an exhaustive search over all possible values of the transmitted symbols and decides in favor of the quadruplet \((s_1, s_2, s_3, s_4)\) which minimizes the Euclidean distance:

\[
D(s_1, s_2, s_3, s_4) = \left\{ \left| r_1 - h_{11} (a s_1 + b s_3) - h_{12} (a s_2 + b s_4) \right|^2 + \left| r_2 - h_{11} (-c s_2^* - d s_4^*) - h_{12} (c s_1^* + d s_3^*) \right|^2 + \left| r_3 - h_{21} (a s_1 + b s_3) - h_{22} (a s_2 + b s_4) \right|^2 + \left| r_4 - h_{21} (-c s_2^* - d s_4^*) - h_{22} (c s_1^* + d s_3^*) \right|^2 \right\}
\]

(10)

An exhaustive search clearly involves the computation of \(M^4\) metrics and \(M^4-1\) comparisons, which is excessive for the 16-QAM and 64-QAM signal constellations. But the proposed STC design lends itself to a low-complexity implementation of the ML detector as we now show.

From the received signal samples \((r_1, r_2, r_3, r_4),\) let us compute the following signals:

\[
w_1 = r_1 - b (h_{11} s_3 + h_{12} s_4) = a (h_{11} s_1 + h_{12} s_2) + n_1
\]

(11)

\[
w_2 = r_2 - d (h_{12} s_3^* - h_{11} s_4^*) = c (h_{12} s_1^* - h_{11} s_2) + n_2
\]

(12)

\[
w_3 = r_3 - b (h_{21} s_3 + h_{22} s_4) = a (h_{21} s_1 + h_{22} s_2) + n_3
\]

(13)
\[ w_{d} = r_{d} - d(h_{22}s_{3}^{*} - h_{21}s_{4}^{*}) = c(h_{22}s_{1}^{*} - h_{21}s_{2}^{*}) + n_{d} \]  
\[ (14) \]

Next, from \((w_{1}, W_{2}, W_{3}, W_{4})\), we compute:

\[ h_{11}^{*}w_{1} = a |h_{11}|^{2}s_{1} + h_{11}^{*}h_{12}s_{2} + h_{11}^{*}n_{1} \]  
\[ (15) \]

\[ h_{12}w_{2}^{*} = c^{*}(h_{12} |s_{1} - h_{11}^{*}h_{12}s_{2}, h_{12}n_{2}^{*}) \]  
\[ (16) \]

\[ h_{21}w_{3}^{*} = a(h_{21}|s_{1} + h_{21}^{*}h_{22}s_{2} + h_{21}^{*}n_{3}) \]  
\[ (17) \]

\[ h_{22}w_{4}^{*} = c^{*}(h_{22}|s_{1} - h_{21}^{*}h_{22}s_{2} + h_{22}n_{4}^{*}) \]  
\[ (18) \]

From those signals, we next compute the signal \(Y_{1}\) given by:

\[ y_{1} = (h_{11}^{*}w_{1} + h_{21}^{*}w_{3}/a + (h_{12}w_{2}^{*} + h_{22}w_{4}^{*}/c^{*}) = (|h_{11}|^{2} + |h_{12}|^{2} + |h_{21}|^{2} + |h_{22}|^{2})s_{1} + n_{3} \]  
\[ (19) \]

With \(n_{3} = (h_{11}^{*}n_{1} + h_{21}^{*}n_{3}/a + (h_{12}n_{2}^{*} + h_{22}n_{4}^{*}/c^{*})\)

It can be seen that the signal \(y_{1}\) has no terms involving symbol \(s_{2}\), and the coefficient of the term in \(s_{1}\) clearly indicates that estimation of \(s_{1}\) benefits from 4th-order detector, we get the ML estimate of symbol \(s_{1}\) conditional on \((s_{3}, s_{4})\). Note that the elimination of the terms involving \(s_{2}\) is possible if and only if the respective coefficients of the symbols \(s_{1}\) and \(s_{2}\) in each column of the code matrix are identical.

Similarly, we compute the intermediate signals \(h_{12}^{*}w_{1}, h_{11}w_{2}, h_{22}w_{3}, h_{21}w_{4}\) and then,

\[ Y_{2} = (h_{11}^{*}w_{1} + h_{22}^{*}w_{3}/a - (h_{11}w_{2}^{*} + h_{21}w_{4}^{*}/c^{*}) = (|h_{11}|^{2} + |h_{12}|^{2} + |h_{21}|^{2} + |h_{22}|^{2})s_{2} + n_{2} \]  
\[ (20) \]

With \(n_{2} = (h_{11}^{*}n_{1} + h_{22}^{*}n_{3}/a + (h_{11}n_{2}^{*} + h_{22}n_{4}^{*}/c^{*})\)

As previously, signal \(Y_{2}\) has no terms involving symbol \(s_{1}\) and the coefficient of the term in \(s_{2}\) shows that estimation of \(s_{2}\) benefits from 4th-order spatial diversity. By sending \(Y_{2}\) to a threshold detector, we get the ML estimate of symbol \(s_{2}\) conditional on \((s_{1}, s_{4})\). ML estimation of \(s_{1}\) and \(s_{2}\) conditional on \((s_{3}, s_{4})\) is illustrated in Figure 5.2. In this way, for a given symbol pair \((s_{1}, s_{2})\), we get the ML estimate of \((s_{1}, s_{2})\), which we denote \((s_{1}^{ML}, s_{2}^{ML})\). Now, instead of computing the metric \(D(s_{1}, s_{2}, s_{3}, s_{4})\) for all \((s_{1}, s_{2}, s_{3}, s_{4})\) values, we only need to compute it for \((s_{1}^{ML}, s_{2}^{ML}, s_{3}, s_{4})\), with \(s_{3}\) and \(s_{4}\) spanning the signal constellation. Specifically, let \((s_{2}^{k}, s_{4}^{l})\) indicate that symbol \(s_{3}\) takes the \(k\)th point of the signal constellation and symbol \(s_{4}\) takes the \(l\)th point of the signal constellation.

The optimum receiver computes the metric \(D(s_{1}, s_{2}, s_{3}, s_{4})\) for \((s_{1}^{ML}, s_{2}^{ML}, s_{3}^{k}, s_{4}^{l})\), where \(k, l = 1, 2, \ldots, M\). This procedure, which is illustrated in Figure 2, reduces the ML receiver complexity from \(M^2\) to \(M^2\).

![Figure 1](image-url)  
**Figure 1** Processing of the received signals to determine the ML estimate of symbols \(s_{1}\) and \(s_{2}\) conditional on a particular combination of symbols \(s_{3}\) and \(s_{4}\).
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Figure 2 Second stage of the estimator.

Note that the special structure of figure 1 allows the ML detector also to work the other way round: Instead of deriving the ML estimate of \((s_1, s_2)\) conditional on \((s_3, s_4)\) and then computing the metric \(D(s_1, s_2, s_3, s_4)\) for \((s_{1k}, s_{2l}, s_{3k}, s_{4l})\), we can first estimate \((s_3, s_4)\) conditional on \((s_{1k}, s_{2l})\), then compute the metric \(D(s_1, s_2, s_3, s_4)\) for \((s_{1k}, s_{2l}, s_{3ML}, s_{4ML})\), for \(k, l = 1, 2, \ldots, M\), and select the quadruplet \((s_1, s_2, s_3, s_4)\) minimizing the metric.

It is instructive to point out here that the described detector is optimum only when the magnitudes of \(a\) and \(c\) (alternatively the magnitudes of \(b\) and \(d\) for the reverse detection order) are equal. This can be easily seen by looking at the SNR at the receiver input and then at the threshold detector input. Indeed, these two SNR values are the same if and only if \(|a| = |c|\) for forward detection and \(|b| = |d|\) for reverse detection.

The \(a, b, c, d\) parameters in the code matrix are design parameters to be optimized in order to obtain full-diversity STC with large coding gain. However, this task is infeasible especially for higher constellation sizes. Fortunately, the transmit power constraints can further decrease the number of parameters to be optimized.

In terms of the transmitted power, the desired conditions can be expressed as

\[
|a|^2 + |b|^2 = 1 = |c|^2 + |d|^2
\]
\[
|a|^2 + |c|^2 = 1 = |b|^2 + |d|^2
\]

The first condition ensures an equal transmit power at each symbol time, while the second condition ensures that equal total power is transmitted for each symbol. These equalities together with the constraint \(|a| = |c|\) for optimal detection lead immediately to the fact that all the design parameters should have the same magnitude, i.e., \(|a| = |c| = |b| = |d| = 1/\sqrt{2}\)

Without any loss of generality, we take \(a = c = 1/\sqrt{2}\) (this allows to decrease the number of unknown parameters without affecting the coding gain) and make an exhaustive search to optimize the parameters \(b\) and \(d\).

**Rate-3/4 2 \times 2 STC**

The STC given in (5.5) can be modified for a further reduction in the optimum detector complexity. More specifically, by setting \(s_4 = s_3\) and scaling the energy of this symbol, we obtain the following 2 \times 2 code with rate 3/4:

\[
X_{new}^{3/4} = \begin{bmatrix}
    a \hat{s}_1 + b \hat{s}_3/\sqrt{2} & -c \hat{s}_2^* - d \hat{s}_3^*/\sqrt{2} \\
    a \hat{s}_2 + b \hat{s}_3/\sqrt{2} & c \hat{s}_1^* + d \hat{s}_3^*/\sqrt{2}
\end{bmatrix}
\]

(21)

Where the notation \(X_{new}^{3/4}\) is used to distinguish the proposed code \(X_{new}(5.5)\) from its reduced-rate version.

In order to detect the symbols transmitted, the full ML detector makes an exhaustive search over all possible values of the transmitted symbols and decides in favor of the triplet \((\hat{s}_1, \hat{s}_2, \hat{s}_3)\) which minimizes the Euclidean distance that we denote by \(D(\hat{s}_1, \hat{s}_2, \hat{s}_3)\). Specifically, this exhaustive search involves the computation of \(M^3\) metrics and \(M^3 - 1\) comparisons, which is also excessive for the 16-QAM and 64-QAM signal constellations. Now, dropping the symbol \(s_4\) lends itself to a lower-complexity implementation of the ML detector at the price of transmission rate reduction.
More precisely, following the same procedure as that presented for the full-rate case, it can be seen that the signals $u_k$, $k = 1, 2$, will have only terms involving the respective symbol $s_k$ and the estimation of symbols $s_k$, $k = 1, 2$, will benefit from full fourth-order spatial diversity. By sending the signals $u_1$ and $u_2$ to a threshold detector, we get the ML estimate of symbol $s_1$ and $s_2$ conditional only on the symbol $s_3$. Note that, as a natural consequence of similarity to the full-rate case, the elimination of the terms involving $s_3$ can be possible if the coefficients $a$ and $c$ have the same magnitude. In this way, for a given value of symbol $s_3$, we get the ML estimate of $(s_1, s_2)$, which we denote $(s_{1,ML}, s_{2,ML}|s_3)$. Now, instead of computing the metric $D(s_1, s_2, s_3)$ for all $(s_1, s_2, s_3)$ values, we only need to compute it for $(s_{1,ML,ML}|s_3, s_3)$. In other words, the optimum receiver computes the metric $D(s_1, s_2, s_3)$ for $(s_{1,ML,ML}|s_3, s_3)$, $l = 1, \ldots, M$. This procedure evidently reduces the ML receiver complexity from $M^3$ to $M$. Optimization of the parameters in the reduced-rate case can be performed similarly to the full-rate case. The parameters $a$ and $c$ can be set to $1/2$ without any loss of generality.

4. RESULTS

In this section, we present some comparisons between the aforementioned new STCs and the existing alternatives. The simulations were carried out for the QPSK, 16-QAM and 64-QAM signal constellations, and the results are obtained for an uncorrelated Rayleigh fading channel with $E[|h|^2] = 1$ for all $k, l$. Two receive antennas were used in all cases.

**Complexity, Rate and Diversity comparison**

| Table 1: Complexity(no. of metrics), Rate and Diversity comparison for different STCs |
|---------------------------------|-------|-------|-------|-------|
|                                | QPSK  | 16 QAM | 64 QAM | Rate  | Diversity            |
| Alamouti STC                   | 16    | 256    | 4096   | 1/2   | Full (4th order)     |
| SM                              | 16    | 256    | 4096   | 1(full) | Half (2nd order)     |
| Golden Code                    | 256   | 65536  | 16777216 | 1    | Full (4th order)     |
| NewFRFD STC(low)               | 16    | 256    | 4096   | 1    | Full (4th order)     |
| Rate 3/4                       | 4     | 16     | 64     | 3/4   | full                 |

Complexity is determined by number of metrics computed for symbol estimation. Table 1 shows Rate $3/4$ code is least complex than Alamouti STC, Spatial Multiplexing, Golden Code because it requires only 4 metrics with QPSK constellation, 16 metrics with 16-QAM constellation and 64 metrics with 64-QAM constellation, it is not full rate (only 3 symbols are transmitted in two time slots). In general Rate $3/4$ code complexity is $M(\text{constellation size})$. So in practical implementation of view Rate $3/4$ code is implementable with less hardware (less chip area), even it can use for higher constellations 512, 1024-QAM. But practical implementation is expensive for Golden Code even it is full rate and full diversity STC at higher constellation sizes.

**Performance Comparison in the Full-Rate Case**

Performance comparisons between the low complexity full-rate full diversity $2 \times 2$ STC and the Alamouti STC. Figure 7.1 shows the BER performance as a function of $E_b/N_0$, where $E_b$ denotes the average signal energy per bit, and provides comparisons between $X_{new}$, namely, the new STC, and $X_s$ (the Golden code). It can be seen that $X_{new}$ achieves the same diversity gain and gives
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essentially the same results as $X_g$ at substantially lower complexity. Indeed, their conclusion is that the performance of low complexity full-rate full diversity $2 \times 2$ STC is marginally very close to that of $X_g$.

![Figure 3 Performance comparison between low complexity FRFD STC and Alamouti STC](image)

The complexity reduction can be observed from table 3, low complexity full-rate full diversity $2 \times 2$ STC results in a considerable reduction in the number of computations. These results indicate that $X_{new}$ enables to reduce the hardware complexity without any significant performance degradation.

![Figure 4 Comparison of Bit Error Rate(BER) performance, SNR for different QPSK constellations of Alamouti, SM and Rate3/4 STCs](image)

**Performance Comparison in the Rate-3/4 Case**

We now provide a performance comparison between $X_{new}^{3/4}$, namely, the proposed rate-3/4 STC, and the two MIMO schemes in current mobile WiMAX system specifications (Alamouti’s STC and SM). With the optimized values, the proposed STC maximizes the diversity gain and, therefore, it achieves the same BER curve slope as Alamouti’s STC with constant coding gain.
independent of the constellation size. This is a crucial property as in the full-rate case, since we do not want vanishing determinants.

The results shown in Figure 4 indicate that the Alamouti scheme achieves a BER of $10^{-1}$ with an SNR of 10 dB for QPSK. Next, we can observe that SM achieves BER of $10^{-0.782}$ with an SNR of 10 dB for QPSK. Finally, our Rate $\frac{3}{4}$ STC achieves a BER of $10^{-1.107}$ with an SNR of 10 dB for QPSK. Clearly, the Alamouti scheme has the best BER performance, but also the lowest bit rate on a given channel bandwidth. The SM scheme doubles the bit rate, but it involves a strong SNR loss, which increases at lower BER values. As evidenced from these results, the proposed rate-$\frac{3}{4}$ scheme is an interesting alternative to these two MIMO schemes, as it substantially improves BER performance compared to SM, and it increases the bit rate by 25% compared to the Alamouti scheme at the price of some SNR loss.

5. CONCLUSION AND FUTURE SCOPE

In this thesis, I have presented a new low complexity full-rate full-diversity $2 \times 2$ STC leading to an inherent low-complexity optimum decoder. We have compared its performance to those of the STCs included in the IEEE 802.16e-2005 specifications, and the results indicated that the proposed scheme achieves the performance of the best known code while reducing the decoder complexity by an order of magnitude in QPSK, two orders of magnitude in 16-QAM, and four orders of magnitude in 64-QAM based MIMO systems. a full-diversity rate-$\frac{3}{4}$ $2 \times 2$ STC whose optimum decoder complexity grows only linearly with the number of constellation points. We have compared its performance to the two MIMO schemes included in the IEEE 802.16e-2005 specifications, and the results indicated that it stands as an interesting alternative providing further tradeoffs between performance and spectral efficiency.

In future the low complexity STC design can implement for $4 \times 4$, $6 \times 6$, and different antenna configurations. Thus, the novel STC design opens up new perspectives for future evolutions of WiMAX systems, as well as for other wireless systems.

References


