Alamouti-based space-frequency coding for OFDM

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ABSTRACT

We investigate space-frequency block coding for OFDM systems with multiple transmit antennas, where coding is applied in the frequency domain (OFDM carriers) rather than in the time domain (OFDM symbols). In particular we consider Alamouti's code, which was shown to be the optimum block code for two transmit antennas and time domain coding. We show that the standard decoding algorithm results in significant performance degradation depending on the frequency-selective nature of the transmission channels, such that a low coherence bandwidth results in a huge degradation. The optimum decoding algorithm that alleviates this problem is the maximum-likelihood decoder for joint symbol detection. We present a performance analysis for the investigated space-frequency decoders in terms of the achievable BER results. Furthermore we compare space-time and space-frequency coding and discuss the respective advantages and drawbacks of the different decoding algorithms in terms of their complexity. It should be noted that for the space-time approach we introduce the so-called matched-filter receiver, which shows significantly lower complexity compared to the maximum-likelihood decoder known from literature. The HIPERMAN system serves as an example OFDM system for quantitative comparisons.

Keywords: Space-frequency coding, joint maximum-likelihood detection/decoding, transmit diversity, OFDM

1. INTRODUCTION

The domain of space-frequency coding was introduced in [1]. Space-frequency coding basically extends the theory of space-time coding for narrowband flat fading channels to broadband time-variant and frequency-selective channels. The application of classical space-time coding techniques for narrowband flat fading channels to OFDM seems straightforward, since the individual subcarriers can be seen as independently flat fading channels. However, in [1] it was shown that the design criteria for space-frequency codes operating in the space-time- and frequency domain are different from those for classical space-time codes for narrowband fading channels as introduced in [2].

In this paper we investigate the application of the Alamouti space-time block code [3] in the frequency domain (over two adjacent OFDM carriers), resulting in a space-frequency block coded system with two transmit antennas. We will show that, when operating in frequency-selective fading channels, the application of conventional decoding algorithms results in a significant performance decrease. This is due to the fact that the equivalent channel matrix is no longer orthogonal. Consequently, independent decoding of the two transmitted symbols, as in conventional decoding algorithms, is no longer appropriate. In this paper we are going to outline joint maximum-likelihood decoding for the two transmitted symbols that allows overcoming the drawbacks of conventional decoding algorithms. Based on the HIPERMAN system, an OFDM system with 201 modulated carriers, both decoding approaches are going to be evaluated in terms of their BER performance and their respective complexity.

2. SYSTEM MODEL

We consider a communication system using space-time block coding with two transmit antennas and at least one receive antenna. The block diagram of such a communication system is depicted in Figure 1.

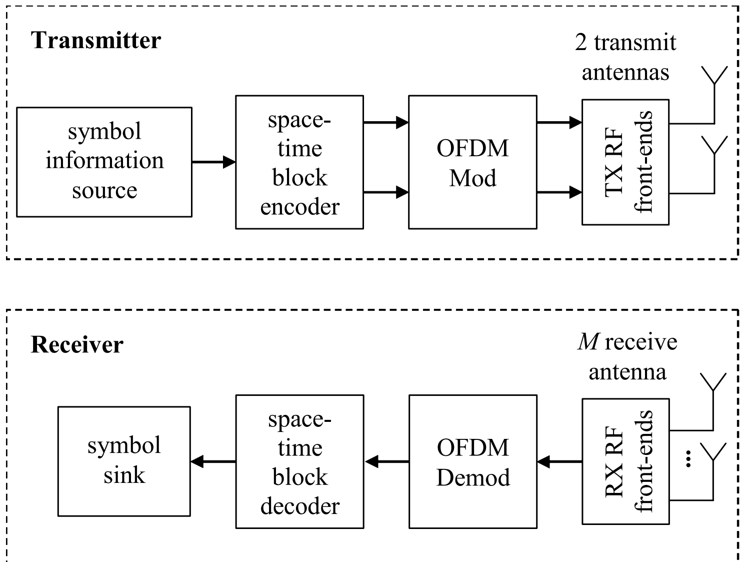

Figure 1. System block diagram.

At the transmitter side the information blocks of symbols are passed to the space-time block encoder, where each block contains two symbols. The space-time block encoder generates the code words of length $M = 2$, where M corresponds to the number of transmit antennas. These code words are passed to the OFDM modulator and the radio frequency (RF) front-ends, which modulate the information onto the carrier frequency. On the receiver side up to N receiver antennas can be used for reception. The RF signals are down-converted and digitized in the RF front-ends and then passed to the OFDM demodulator and the space-time block decoder. The space-time block decoder interprets the received signals and generates estimates for the transmitted information symbols, which are again provided in blocks of two symbols.

2.1. Transmitter processing and signal model

2.1.1. Space Time Coding (STC)

In general, a space–time block code is defined by a $p \times M$ transmission matrix K. The entries of this transmission matrix are linear combinations of the transmitted symbols and their conjugates. The number of transmission antennas is M, and we usually use this parameter as index to separate different codes from each other. For example, K_2 represents a code that utilizes two transmit antennas and is defined by [3]

$$K_2 = \begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix}, \tag{1}$$

We assume that transmission at the baseband employs a signal constellation set with 2^b elements. At each time slot, which consists of p OFDM symbols (e.g. $p = 2$ for K_2), nb bits arrive at the encoder of subcarrier k and constellation symbols $s_{1,k},\ldots,s_{n,k}$ are selected. Setting $x_i = s_{i,k}$ for $i = 1,2,\ldots,n$ in K, we obtain the matrix $Ç_k$ for each subcarrier k, where the matrix entries are linear combinations of $s_{1,k},\ldots,s_{n,k}$ and their conjugates. They are specific constellation symbols that are transmitted from the M transmit antennas. If, for subcarrier k, $c_{t,k}^{(i)}$ represents the element in the t^{th} row and the i^{th} column of $Ç_k$, the entries $c_{t,k}^{(i)}, i = 1,2,\ldots,M$ are transmitted simultaneously from transmit antennas $1,2,\ldots,M$ during the OFDM symbol $t = 1,2,\ldots,p$ *. So, for each individual subcarrier, the i^{th} column of $Ç_k$ represents the transmitted symbols from the i^{th} antenna and the t^{th} row of $Ç_k$ represents the transmitted constellation symbols at OFDM symbol t. Note that $Ç_k$ is basically defined using K, and the orthogonality of K's columns allows a simple decoding scheme which will be explained in Section 2.1. Since p OFDM symbols are used to transmit n constellation symbols, we define the rate of the code to be $R = n/p$. For example, the rate of K_2 is one.

In the sequel we will only consider Alamouti's transmission matrix K_2. For this transmission matrix the signal $y_{t,k}$, which is received at subcarrier k during OFDM symbol t is given by

$$y_{t,k} = \sum_{i=1}^{2} H_{i,k} c_{t,k}^{(i)} + n_{t,k}, \qquad (2)$$

where $H_{i,k}$ is the path gain of the channel transfer function of subcarrier k and transmit antenna i. The noise samples $n_{t,k}$ observed at subcarrier k are independent samples of a zero-mean complex Gaussian random variable with variance M/γ, where γ is the SNR†. The average energy of the symbols transmitted from each antenna is normalized to be one, so that the average power of the received signal at each receive antenna is M and the resulting signal-to-noise ratio is γ.

The transmission model of eq. (2) can also be expressed by means of a vector signal model. Without loss of generality we set $t = 1$ and by using the following equivalence

$$K_2 = \begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix} \rightarrow \begin{pmatrix} s_{1,k} & s_{2,k} \\ -s_{2,k}^* & s_{1,k}^* \end{pmatrix} = \begin{pmatrix} c_{t,k}^{(1)} & c_{t,k}^{(2)} \\ c_{t+1,k}^{(1)} & c_{t+1,k}^{(2)} \end{pmatrix}, \qquad (3)$$

the received signal vector $\boldsymbol{Y}_k = \begin{bmatrix} y_{1,k} & y_{2,k}^* \end{bmatrix}^T$ can be expressed as follows:

$$\boldsymbol{Y}_k = H_k \boldsymbol{X}_k + \boldsymbol{N}_k, \qquad (4)$$

where $\boldsymbol{X}_k = \begin{bmatrix} s_{1,k} & s_{2,k} \end{bmatrix}^T$ and $\boldsymbol{N}_k = \begin{bmatrix} n_{1,k} & n_{2,k}^* \end{bmatrix}^T$ are the transmitted and additive noise signal vectors, respectively and the channel matrix, which is imposed by the Alamouti code matrix, is given by

$$H_k = \begin{pmatrix} H_{1,k} & H_{2,k} \\ H_{2,k}^* & -H_{1,k}^* \end{pmatrix}. \qquad (5)$$

Note that transmission model assumes that the channel does not vary during the two consecutive OFDM symbols, e.g. $H_{i,k}[t] = H_{i,k}[t+1]$, which results in an orthogonal channel matrix.

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* Note that the entries $c_{t,k}^{(i)}, i = 1,2,\ldots,M$ of all $k = 1,2,\ldots,N_{SC}$ modulated subcarriers are transmitted in parallel during one OFDM symbol period.

† Note that the increase of the noise energy by M corresponds to an increase of the transmit energy since each of the M transmit antennas is transmitting with the same power. Due to the increase of noise energy the received SNR is kept constant, which in turn ensures a fair comparison with systems employing one single transmit antenna.

2.1.2. Space Frequency Coding (SFC)

Let us now consider the application of the Alamouti code matrix [3] with coding in the frequency domain across two adjacent subcarriers and thus within one OFDM symbol time. The received signal vector $\boldsymbol{Y}_k' = \begin{bmatrix} y_k & y_{k+1}^* \end{bmatrix}^T$ now contains two adjacent subcarrier signals and may be represented in the vector signal model as[‡]:

$$\boldsymbol{Y}_k' = H_k' \boldsymbol{X}_k' + \boldsymbol{N}_k', \tag{6}$$

where $\boldsymbol{X}_k' = \begin{bmatrix} s_k & s_{k+1} \end{bmatrix}^T$ and $\boldsymbol{N}_k' = \begin{bmatrix} n_k & n_{k+1}^* \end{bmatrix}^T$ are the transmitted and additive noise signal vectors, respectively and the channel matrix is given by

$$H_k' = \begin{pmatrix} H_{1,k} & H_{2,k} \\ H_{2,k+1}^* & -H_{1,k+1}^* \end{pmatrix}. \tag{7}$$

Note that for frequency-selective channels the channel coefficients of the two adjacent OFDM carriers k and $k+1$ are, in general, not equal ($H_{i,k} \neq H_{i,k+1}$) such that the channel matrix H_k is no longer orthogonal.

2.2. Receiver processing

2.2.1. Space Time Coding (STC)

The space–time block code K_2 uses the transmission matrix in (1) for each individual subcarrier k. Consequently, maximum likelihood detection for the transmission model (4) leads to minimizing the decision metric

$$D\left(s_{1,k}, s_{2,k}\right) = \left|\boldsymbol{Y}_k - H_k \boldsymbol{X}_k\right|^2 = \left|y_{1,k} - H_{1,k} s_{1,k} - H_{2,k} s_{2,k}\right|^2 + \left|y_{2,k} + H_{1,k} s_{2,k}^* - H_{2,k} s_{1,k}^*\right|^2 \tag{8}$$

for all possible values of $s_{1,k}$ and $s_{2,k}$, thus for all carriers $k = 1, 2, \ldots, N_{SC}$. The minimizing values are the receiver estimates of $s_{1,k}$ and $s_{2,k}$, respectively. By expanding the above metric and by deleting the terms that are independent of the codewords, we observe that the above metric can be decomposed in to two separate parts for detecting each individual symbol: minimizing the decision metric

$$\left|\left(y_{1,k} H_{1,k}^* + y_{2,k}^* H_{2,k}\right) - s_{1,k}\right|^2 + \left(\sum_{i=1}^{2} \left|H_{i,k}\right|^2 - 1\right) \left|s_{1,k}\right|^2 \tag{9}$$

for detecting $s_{1,k}$ and the decision metric

$$\left|\left(y_{1,k} H_{2,k}^* - y_{2,k}^* H_{1,k}\right) - s_{2,k}\right|^2 + \left(\sum_{i=1}^{2} \left|H_{i,k}\right|^2 - 1\right) \left|s_{2,k}\right|^2 \tag{10}$$

for decoding $s_{2,k}$. This is the simple decoding scheme described in [3]. Note that the decomposition and the consequent individual decoding of the two transmitted symbols is possible due to the fact that the channel matrix (5) is orthogonal (quasi-static channel assumption with constant path gains over two consecutive OFDM symbols).

[‡] Note that the index t is no longer required since all constellation symbols are transmitted within the same OFDM symbol and the carrier index k is a sufficient identifier. Hence, the index t was suppressed for simplicity and ease of reading.

Next we will show that the maximum likelihood decoding solution of (9) and (10) is equivalent to the solution of the standard matched filter receiver. The zero-forcing equalizer for the signal model in (4) provides the unbiased estimates $\hat{\boldsymbol{X}}_k = \left[\hat{s}_{1,k} \quad \hat{s}_{2,k}\right]^T$ and is given by

$$\hat{\boldsymbol{X}}_k = \left(H_k^H H_k\right)^{-1} H_k^H \boldsymbol{Y}_k . \tag{11}$$

Note that the first term $\left(H_k^H H_k\right)^{-1}$ in the above equation corresponds to the equalization part whereas the term H_k^H is the actual matched filter part of the matched filter receiver for the flat fading vector signal model (4). By using the structure of the channel matrix in (5), the entire matched filter matrix can be written as

$$\left(H_k^H H_k\right)^{-1} H_k^H = \frac{1}{\sum_{i=1}^{2}\left|H_{i,k}\right|^2}\begin{pmatrix} H_{1,k}^* & H_{2,k} \\ H_{2,k}^* & -H_{1,k} \end{pmatrix}. \tag{12}$$

Consequently, we obtain the estimates $\hat{s}_{i,k}$ for the transmitted signal $s_{1,k}$ and $s_{2,k}$ as

$$\begin{aligned} \hat{s}_{1,k} &= \frac{1}{\sum_{i=1}^{2}\left|H_{i,k}\right|^2}\left(H_{1,k}^* y_{1,k} + H_{2,k} y_{2,k}^*\right) \\ \hat{s}_{2,k} &= \frac{1}{\sum_{i=1}^{2}\left|H_{i,k}\right|^2}\left(H_{2,k}^* y_{1,k} - H_{1,k} y_{2,k}^*\right) \end{aligned}, \tag{13}$$

which provides the detailed description for the implementation of the matched filter receiver. From (13), we observe that the following equation holds:

$$-\left(H_{1,k}^* y_{1,k} + H_{2,k} y_{2,k}^*\right) + \hat{s}_{1,k}\sum_{i=1}^{2}\left|H_{i,k}\right|^2 = 0$$

and

$$-\left(H_{2,k}^* y_{1,k} - H_{1,k} y_{2,k}^*\right) + \hat{s}_{2,k}\sum_{i=1}^{2}\left|H_{i,k}\right|^2 = 0 .$$

We deduce that for $\hat{s}_{i,k} = s_{i,k}$ the previous two metrics show a minimum. By expanding the previous expressions and by deleting the terms that are independent of the codewords the above metrics decompose into (9) and (10), which are the decision metrics of the maximum likelihood receiver. This goes to show that the matched filter solution proposed in (12) is identical to the maximum likelihood detector proposed in [3].

This observation is not surprising since it is well known that, in general, the ZF equalizer provides the optimum solution for flat fading channels. Hence, for the transmission model (4) the ZF solution is equivalent to the maximum-likelihood solution.

2.2.2. Space Frequency Coding (SFC)

In the same way as before we can now derive the matched filter receiver for the SFC signal model (6). The zero-forcing equalizer providing the unbiased estimates $\hat{\boldsymbol{X}}_k = \left[\hat{s}_k \quad \hat{s}_{k+1}\right]^T$ can be written as:

$$\left(H_k'^H H_k'\right)^{-1} H_k'^H = \frac{1}{H_{1,k}H_{1,k+1}^* + H_{2,k}H_{2,k+1}^*}\begin{pmatrix} H_{1,k+1}^* & H_{2,k} \\ H_{2,k+1}^* & -H_{1,k} \end{pmatrix} \tag{14}$$

Consequently, we obtain the estimates $\hat{s}_k$ and $\hat{s}_{k+1}$ for the transmitted signal s_k and s_{k+1} as:

$$\begin{aligned} \hat{s}_k &= \frac{1}{H_{1,k}H_{1,k+1}^* + H_{2,k}H_{2,k+1}^*}\left(H_{1,k+1}^* y_k + H_{2,k} y_{k+1}^*\right) \\ \hat{s}_{k+1} &= \frac{1}{H_{1,k}H_{1,k+1}^* + H_{2,k}H_{2,k+1}^*}\left(H_{2,k+1}^* y_k - H_{1,k} y_{k+1}^*\right) \end{aligned}. \tag{15}$$

Note that for the flat fading case, where $H_{i,k+1} = H_{i,k}$, the above matched filter receiver for the frequency-selective case converges towards solution (13). However, for the frequency-selective fading case the channel matrix H_k' is no longer orthogonal and we consequently deduce that the matched filter receiver no longer provides the optimum solution.

The maximum likelihood decision metric, which yields the optimum signal detector, now detects the symbols s_k and s_{k+1} jointly and is given by

$$D\left(s_k, s_{k+1}\right) = \left|\boldsymbol{Y}_k' - H_k'\boldsymbol{X}_k'\right|^2 = \left|y_k - H_{1,k}s_k - H_{2,k}s_{k+1}\right|^2 + \left|y_{k+1} - H_{2,k+1}s_k^* + H_{1,k+1}s_{k+1}^*\right|^2 \tag{16}$$

where the minimization over all possible combinations of s_k and s_{k+1} provides the estimates for the two transmitted symbols. Note that, since the channel matrix H_k' is no longer orthogonal, the maximum likelihood decision metric does not decompose into two separated metrics for detection of s_k and s_{k+1}[§]. Hence, the two symbols have to be detected jointly. By neglecting the terms that are independent of the transmitted symbols s_k and s_{k+1} the above decision metric reduces to

$$\begin{aligned} D\left(s_k, s_{k+1}\right) = &\left(\left|H_{1,k}\right|^2 + \left|H_{2,k+1}\right|\right)\left|s_k\right|^2 + \left(\left|H_{1,k+1}\right|^2 + \left|H_{2,k}\right|^2\right)\left|s_{k+1}\right|^2 - \\ &2\cdot\Re\Big\{\left(H_{1,k}y_k^* + H_{2,k+1}^* y_{k+1}\right)s_k + \left(H_{2,k}y_k^* - H_{1,k+1}^* y_{k+1}\right)s_{k+1} + \\ &\left(H_{1,k+1}H_{2,k+1}^* - H_{1,k}H_{2,k}^*\right)s_k s_{k+1}^*\Big\} \end{aligned} \tag{17}$$

3. PERFORMANCE EVALUATION

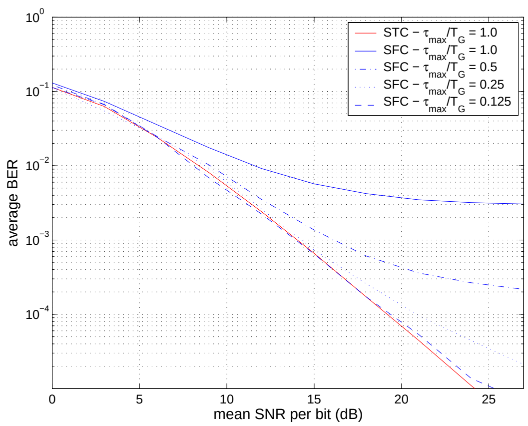
Simulation results for the simple and the joint maximum-likelihood decoding approach are presented in the following figures. The BER for HIPERMAN system parameters and Alamouti-based space-frequency coding are shown for channels with different coherence bandwidths. The guard interval was set to its maximum value $T_G = TN/4$ (T is the DFT sampling time and N is the FFT size) and the maximum excess delay $\tau_{\max}$ of the channel was varied in order to produce channels with different coherence bandwidth. In other words, channels with long excess delay times, e.g. $\tau_{\max} = T_G$, exhibit a lower coherence bandwidth – corresponding to higher frequency-selectivity – than shorter channels. All channels are based on an exponentially decaying power delay profile. The performance results of the conventional Alamouti-based space-time coding on an OFDM carrier basis[**] and with matched filter based decoding is given for comparison purposes. Results for space-

[§] The same applies for STC in case where the transmission channel is not constant over two consecutive OFDM symbols.

[**] For these results we assume that the channel does not change over the duration of the transmission of the space-time code, which is two consecutive OFDM symbols in the case of the Alamouti code.

frequency coding are referred to by the acronym SFC whereas the space-time coding results are referred to by the term STC. Performances were investigated for QPSK, 16QAM and 64QAM modulations. Figure 2 shows the results for QPSK modulations. From the results for the joint maximum-likelihood decoding it can be seen that with the new decoding approach the frequency-selective nature of the channel no longer degrades the performance as with the traditional matched filter approach, in the sense that the error floor for the BER that appears for the conventional approach can be eliminated through joint maximum likelihood decoding. Similarly, this observation also holds for 16QAM and 64QAM modulations.

The above figures illustrate that Alamouti-based space-frequency coding can achieve the same performance as Alamouti-based space-time coding if joint symbol detection is employed. This observation can be extended to other block-coding schemes, e.g. involving more than two transmit antennas.



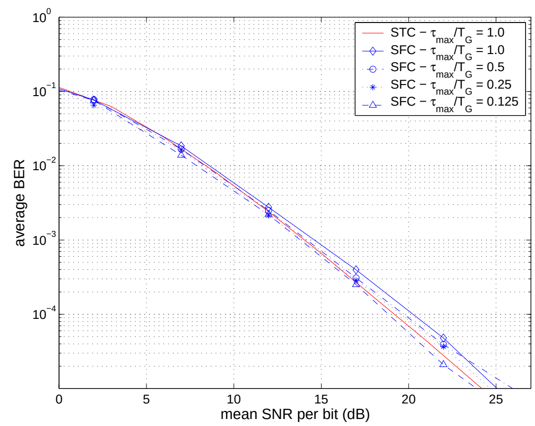


Figure 2. Alamouti-based SFC for QPSK modulation and different frequency selective transmission channels and with matched filter (left) and advanced joint maximum likelihood decoding (right).

4. COMPLEXITY EVALUATION

For the complexity analysis we are going to compare the number of complex-valued additions/subtractions and multiplications as well as the number of real-valued metrics that have to be computed for the maximum-likelihood decoders. We assume that the received signals and the channel estimates are available in the form complex numbers represented through their real and imaginary part. Furthermore, we assume that the values for the symbols s_i as well as the values for $|s_i|^2$ and $s_i s_j^*$ are stored in the receiver and hence directly available for decoding. As for complex- and real-valued operations we assume that one complex-valued addition is equal to two real-valued additions and that a complex-valued multiplication accounts for four real-valued multiplications. Furthermore, the multiplication of a real-valued quantity by the factor 2, like e.g. the term on the right hand side of (17), is implemented by means of one real-valued addition. Note furthermore, that the number of complex-valued memory cells that are required to store the constellation symbols, e.g. s_i and $s_i s_j^*$, is going to be referred by the term symbol memory.

Let us first consider space-time coding and the complexity of the two decoding algorithms presented in Section 2.2.1, namely the maximum-likelihood decoder given by (9) and (10) and the matched-filter decoder (13). The discussion of the complexity of the maximum-likelihood decoder is presented in the following paragraphs. The complexity of the matched-filter receiver is straight-forward. For each of the bracketed terms in (13) we require one complex-valued addition as well as two complex-valued multiplications. The first term in (13) is common to both symbols that need to be decoded and this term requires one real-valued addition (or equally 0.5 complex-valued additions) as well as two complex-valued multiplications. In total this comes to 2.5 complex-valued additions and six complex-valued multiplications. Furthermore, we require one real-valued division for the first term in (13). It is important to note, that the complexity of the matched-filter receiver does not depend on the symbol constellation, a characteristic that does not apply to the maximum-likelihood decoder. In Table 1 quantitative results of the complexity of both decoders are presented for different common symbol constellation, namely QPSK, 16QAM, and 64QAM.

By comparing the results in Table 1 for the maximum-likelihood and the matched-filter receiver, we can see that the matched-filter receiver shows a significantly lower complexity as far as additions and multiplications are concerned. This is especially true for higher order modulations like, e.g., 64QAM. Moreover, this solution does not require to store neither symbols nor metrics. The only drawback of the matched-filter decoder is the need for one real-valued division.

	QPSK			**16QAM**			**64QAM**		
	ML	**MF**	**decrease ML → MF**	**ML**	**MF**	**Decrease ML → MF**	**ML**	**MF**	**decrease ML → MF**
real decision metrics	8	0	∞	32	0	∞	128	0	∞
additions/subtractions	16	2.5	**6.4**	52	2.5	**20.8**	196	2.5	**78.4**
complex multiplications	18	6	**3.0**	48	6	**8.0**	168	6	**28.0**
real divisions	0	1	**N.A.**	0	1	**N.A.**	0	1	**N.A.**
symbol memory cells	6	0	∞	24	0	∞	96	0	∞

Table 1. Complexity analysis for space-time coding with comparison of maximum-likelihood (ML) and matched-filter (MF) decoders, results for QPSK, 16QAM, and 64QAM.

Let us next consider the complexity of the decoding algorithms for space-frequency decoding as outlined in Section 2.2.2. We are interested in the complexity of the joint maximum likelihood decoding as described by (17) and the disjoint maximum likelihood decoding outlined in (9) and (10). Note that the latter corresponds to the complexity for the space-time coding scheme mentioned above, e.g. in Table 1.

For the disjoint detection approach we need to compute 2^b metrics for each of the two transmission symbols, where b is the number of bits per modulated symbol, e.g. 2 in the case of QPSK modulation. Joint detection, in contrast, requires 2^{2b} metrics in order to determine the two symbols that jointly minimize the metric in (17). In order to minimize the number of additions, subtractions, and multiplications it is advantageous to arrange the expressions for the decision metrics such that terms that are independent of the symbols s_i are grouped together, e.g. as in (17). The number of necessary complex-valued additions/subtractions and multiplications are provided in Table 2. For the disjoint detection approach we need to store 2^b complex-valued symbols s_i as well as their real-valued squared amplitude values $|s_i|^2$. Due to their real-valued nature we require only half the storage place for the latter values. For the joint detection approach we need to foresee additional 2^{2b} complex-valued storage cell for the values $s_i s_j^*$. The number of real-valued decision metrics as well as the number of complex-valued operations and memory cells is summarized in the following table.

	disjoint detection	joint detection
real decision metrics	$2 \cdot 2^b$	2^{2b}
additions/subtractions	$4+3 \cdot 2^b$	$4+3.5 \cdot 2^{2b}$
complex multiplications	$8+2.5 \cdot 2^b$	$10+4 \cdot 2^{2b}$
symbol memory cells	$1.5 \cdot 2^b$	$1.5 \cdot 2^b + 2^{2b}$

Table 2. Number of complex-valued operations and memory cells for space-frequency coding with comparison of disjoint and joint maximum-likelihood decoding.

In Table 3 the results of the complexity analysis in terms of the complexity increase, which is given as the fraction of the number for joint versus the number for the disjoint decoding approach, are provided for QPSK, 16QAM and 64QAM modulation. Although the complexity increase is quite moderate for QPSK, the increase, especially of the required number of complex-valued multiplications, might be prohibitive for implementation of higher order modulation schemes like, e.g., 64QAM. It should be noted, however, that in the system context the complexity of the space-frequency decoder needs to be put in relation to the complexity of other functional blocks. Let for example consider the FFT operation at the receiver. The number of complex multiplications required for an N point FFT equals to $N \cdot \log_2 N$, which comes to 2048 complex multiplications for the HIPERMAN system, where $N = 256$. As such, the optimal space-frequency decoder based on joint maximum likelihood decoding is approximately higher times as complex as the FFT operation.

	QPSK (b=2)			16QAM (b=4)			64QAM (b=6)		
	disjoint	joint	increase	disjoint	joint	increase	disjoint	joint	increase
real decision metrics	8	16	**2.0**	32	256	**8.0**	128	4096	**32.0**
additions/subtractions	16	60	**3.8**	52	900	**17.3**	196	14340	**73.2**
complex multiplications	18	74	**4.1**	48	1034	**21.5**	168	16394	**97.6**
symbol memory cells	6	22	**3.7**	24	280	**11.7**	96	4192	**43.7**

Table 3. Complexity analysis for space-frequency coding with comparison of disjoint and joint maximum-likelihood decoders, results for QPSK, 16QAM, and 64QAM and presentation of required increase of complexity for joint ML decoder.

5. CONCLUSIONS

Significant attention has been given to the well-known Alamouti code since the diversity gain is achieved without loss in bandwidth efficiency. In contrast to conventional space-time block coding with coding in the temporal domain, this paper also addresses space-frequency coding, i.e. coding across the OFDM subcarriers within one OFDM symbol. The study considers the Alamouti code as a study case. The BER performance of different detection methods has been investigated. First, we show that for space-time coding the conventional scalar (disjoint) maximum likelihood detection is equivalent to the matched-filter receiver. It was shown that the complexity of the newly introduced matched-filter solution is independent of the symbol constellation size and results in a significantly lower complexity in terms of the required additions and multiplications. Second, we show that the use of conventional disjoint maximum likelihood detection for space-frequency decoding leads to a significant performance degradation, which is due to the frequency selective nature of the transmission channel. This performance degradation can be avoided by using joint maximum likelihood detection. The complexity in terms of the number of required operations and the required memory has been studied for this detector and we show that the complexity increase with respect to the standard (disjoint) maximum likelihood detection might be prohibitive for large constellation sizes.

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BIOGRAPHIES

Andreas A. Hutter received the Dipl.-Ing. (electrical engineering) and the Dr.-Ing. degree from Munich University of Technology (TUM) in 1997 and 2001, respectively. From 1997 to 2000 he was with the research and engineering department (FIZ) of BMW at Munich where he was project leader for the broadband wireless data initiative. In 2000 he was visiting researcher at Stanford University and in 2001 he joined the Swiss Center for Electronics and Microtechnology (CSEM) as senior R&D engineer. His research interests include the characterization of the propagation characteristics of mobile communication channels, signal-processing techniques for multiple antenna systems and the different aspects related to the design of ultra wideband systems. Andreas A. Hutter is co-recipient of the VTC-Fall 1999 best paper award.

Selim Mekrazi received his M.Sc in Digital Communications, Signal Processing and Telecommunications from Université de Rennes 1, Ecole Nationale Supérieure de Télécommunications de Bretagne (ENST Bretagne) and Ecole Supérieure d'Electricité (Sup'Elec), France, in 2003. In October 2003, he joined Eurecom Institue (Sophia-Antipolis, France) where he is currently pursuing his Ph.D in Electrical Engineering. His general interests lie in the areas of information theory, signal processing, digital communications, micro-electronics and public safety systems. Current researches focus on physical layer transmission techniques and implementation aspects for high-throughput, reconfigurable and rapidly deployable systems.

Beza Negash Getu was born in Bahir Dar, Ethiopia, on May 27, 1975. From 1992 to 1997, he followed Addis Abeba University and he completed his B.Sc. degree in Electrical Engineering. Following his graduation, he was employed at Bahir Dar University as an assistant lecturer. From June 1998 to August 2000, he studied the Master of Science Program of Delft University of Technology, the Netherlands and obtained his M.Sc. degree in Electrical Engineering. In October 2000, he joined the Antennas and Propagation group of Prof. Dr. Techn. Jørgen Bach Andersen at the Center for Person Kommunikation (CPK) of Aalborg University, Denmark as a Ph.D. student.

He worked in the field of Wireless Communications focusing on smart antennas and MIMO systems. The subject are encompasses communication theory, propagation and antenna research with the goal of optimizing link spectral efficiency and bit error rate. During 2002-2003, he spent six months in the Wireless Communications group at CSEM, Neuchâtel, Switzerland, working on the same area. Beza N. Getu received the Ph.D. degree from Aalborg University in 2003.

Fanny Platbrood received her Dipl.Eng. (M.Sc.EE) Degree in Electrical Engineering from the "Faculté Polytechnique" (Mons-Belgium) in 1996 after having made some research at the University of Rochester (NY-USA) and at the Swiss Federal Institute of Technology (EPFL). She worked as ASIC designer in the VLSI Design Department of Alcatel Bell, Belgium until end 1998. She was more particularly responsible for the ASIC development and testing for ADSL. In fall 1998, she joined the CSEM to work on research and development as an expert in wireless communications. She worked then on the PHY layer development for the WLAN standard HiperLAN Type-2 (H/2). From September 1999 until September 2001, she worked in the ESPRIT SLATS European project where she began first as workpackage leader of the GSM and WCDMA PHY software module development to become later the SLATS project manager for CSEM. In 2001, she was responsible for a UMTS concept study in a receiver structure. Up to end 2003, she worked on the IST SCOUT project on software architectures for re-configurable baseband systems and APIs definition. From 2002 to September 2004, she was responsible for the IST STRIKE project where she worked on Multiple Transmit Multiple Receive (MTMR) coding techniques applied to BFWA systems (HIPERMAN). From 2001 to 2004, she was the technical project manager of the IST PRODEMIS project. From 2003, she is task leader in the IST MAGNET project. She is presently project manager at CSEM and her areas of expertise are in ASIC design, digital and mobile communications. She published conference and magazine papers.