

Linear maximum likelihood decoding of space – time block coded OFDM systems for mobile communications

G.T.F. de Abreu, H. Ochiai and R. Kohno

Abstract: A new linear decoder for space–time block codes is presented, which enables linear maximum-likelihood decoding in scenarios beyond those described by block-fading models. The new decoder employs linear combinations of received signals with weights designed so as to yield orthogonal estimates for all encoded symbols regardless of channel variations within the block transmission, provided that the channel is known. The combination weights are scalar functions of the channel estimates, obtained through systematic rotations over the decoding matrices of the linear maximum-likelihood decoder for the block-fading case. The proposed decoder yields good performance with low-complexity and is therefore suitable for low-cost portable receivers supporting space–time transmit diversity at the downlink of high data-rate wireless mobile communication systems. Combined with OFDM technology, the technique yields a low-complexity receiver that is effective in combating both time and frequency selectivities, and provides a tool to add flexibility to wireless local area network systems and to offer high data-rate services with support to mobility.

1 Introduction

Orthogonal block space–time transmit diversity, also known as orthogonal space–time block codes (STBCs), has attracted increasing attention since the proposal of the Alamouti scheme [1] and its generalisation by Tarokh [2, 3]. In fact, when combined with maximum ratio combiner (MRC) receive diversity schemes, this technique can transfer a significant amount of the system complexity to the base stations.

Although various space–time codes have since been invented, orthogonal STBCs remain attractive for their excellent balance of performance and low complexity. Orthogonal STBCs were originally developed for high-data-rate systems in flat, block-fading channels, based on the assumptions that symbol rates are much larger than typical fading rates and that frequency selectivity can be neglected. In such channels, a fundamental advantage of the block structure, in comparison to the alternative trellis structure such as those proposed in [4, 5], is the possibility of decoupling symbol estimates through simple linear combinations of the received signals, enabling low-complexity symbol-by-symbol maximum likelihood decoding (MLD). This technique, hereafter referred to as linear maximum likelihood decoding (LMLD), was one of the main appeals of Alamouti's original scheme [1] and has been recently generalised [6] to orthogonal space–time block codes of any

order [2]. Linear maximum likelihood decoding of orthogonal space–time block codes brings the complexity involved in the application of such codes to implementable levels (even with higher order modulation schemes). Unfortunately, this attractive technique is presently known only for space–time block codes in block-fading channels, which inherently limits its applications to low mobility scenarios.

In the ever more information-hungry modern and future societies, however, high-data-rate services are demanded not only by stationary users, but also by users on the move. A prominent example of practical interest is public wireless local area networks (WLANs) increasingly found in public places such as airports and train stations, and utilised by people moving at different speeds, on foot and travelers, as well as bicycles, karts and shuttles. Another is private WLANs supporting seamless indoor–outdoor transitions so as to serve users not only inside buildings but also outside.

However, in such systems with support for moderate mobility, the effect of time-varying frequency selectivity over space–time code structures cannot be neglected. In a single-carrier high-data-rate wireless system applying STBCs, for instance, frequency selectivity causes encoded blocks to overlap with random delays. The resulting intersymbol interference (ISI) destroys the orthogonality of the code and impedes (conventional) linear maximum likelihood decoding.

Fortunately, frequency selectivity can be effectively mitigated with orthogonal frequency-division multiplexed (OFDM) technology [7]. Nevertheless, there are intrinsic characteristics of OFDM systems that make the introduction of space–time block codes into such systems less straightforward than might be expected at first glance. For instance, the symbol periods in OFDM systems are much longer than that of a single-carrier system with the same data-rate, such that a fading process slow enough to be considered block-fading in a single carrier system, might not be so in a system with an OFDM architecture. Similarly, in the frequency domain, implementation issues limit the total

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number of carriers of OFDM systems so that adjacent carriers are still significantly separated, which implies that 'space–frequency' implementations of space–time coding techniques (as suggested in [8]) are also subject to a departure from the original assumption of a quasi-static channel. To summarise, there are applications when neither in the time nor in the frequency domains, can the OFDM channel be considered quasi-static so as to support a straightforward implementation of space–time block codes.

Aiming at combining the advantages of OFDM and STBCs (with LMLD), so as to design low-cost high-data-rate wireless communication systems with support for mobility, applications of the Alamouti scheme to OFDM systems were investigated in [8–11]. Due to the fact that OFDM channels are not quasi-static, the results reported in [8–11] were, however, less than satisfactory. Indeed, it was observed that the bit error rate performances of space–time block encoded OFDM systems saturated (reached error floors) at higher signal-to-noise ratios. The same conclusion was reached in a previous work where the Alamouti scheme was applied in a non-block-fading channel [12], tracked with a (computationally costly!) Kalman filter.

What went unnoticed in all such works (i.e., [8–12] and their like), is that the error floors at high signal-to-noise ratios observed with STBCs in non-quasi-static fading channels are not an intrinsic problem of the codes, but a consequence of inadequate decoding. In fact, strict MLD is theoretically always possible (see [3]), at the price of a higher complexity (although this alternative is in most cases prohibitively complex).

In addition to the low decoding complexity involved, there are other good reasons that stimulate us to consider LML decoded STBCs for low-cost, portable receivers in mobile wireless OFDM systems in urban environments (with moderate, yet non-block, fading channels). For instance, highly accurate, low-complexity, two-dimensional channel estimation techniques are available [13–15] for OFDM systems in these scenarios. Since decoding orthogonal STBCs is essentially a coherent process and as such, dependent on the accuracy of channel estimation, this property of OFDM systems highly favour STBCs as opposed to space–time trellis codes. In fact, while the fading processes typically faced by mobile wireless OFDM systems in urban environments is fast enough to impair the introduction of (non-coherent) differential space–time block coding [16, 17], these are not fast enough to favour the use of more complex trellis-based space–time diversity methods [4, 5], which are, in general the best fit for fast-fading channels. Finally, unlike the latter, STBCs with LMLD enable real-time decoding of transmitted signals, required by some applications.

2 Channel model

Consider an M -carrier OFDM system with bandwidth B and center frequency f_C in a multipath mobile wireless channel with delay spread $\Delta\tau$. The complex baseband representation of such a channel is given by

$$h(t) = \sum_k \gamma_k(t) \delta(t - \tau_k) \quad (1)$$

where $\gamma_k(t)$ and τ_k are the complex amplitude and the normalised delay of the k th path, respectively.

Owing to motion, $\gamma_k(t)$ are wide-sense stationary complex gaussian processes, independent for each k . In addition, both their average power σ_k^2 , and probability of occurrence, decay exponentially with τ_k [18].

If v is the relative speed between transmitter and receiver, the Doppler spread at the m th carrier is given by

$$f_D(m) = \frac{v}{c} \left(m \frac{B}{M-1} + f_C - \frac{B}{2} \right) \quad (2)$$

where $m=0, \dots, M-1$.

Typically, however, $B \ll f_C$ and $f_C \gg 0$ that $f_D(m)$ can be considered the same at all carriers. The time varying frequency response of the channel, $H(t, f)$, can then be written as

$$H(t, f) = \sum_k \gamma_k(t) e^{-j2\pi f \tau_k} \quad (3)$$

The two-dimensional complex fading process $H(t, f)$ has an envelope autocorrelation well approximated by [19]

$$R_H(\Delta t, \Delta f) = \sigma_H^2 \frac{J_0^2(2\pi \Delta t f_D)}{1 + (2\pi \Delta \tau \Delta f)^2} \quad (4)$$

where σ_H^2 is the total average power of the channel and $J_0(\cdot)$ is the 0th order Bessel function of the first kind. An illustrative example of channels with this description is shown in Fig. 1.

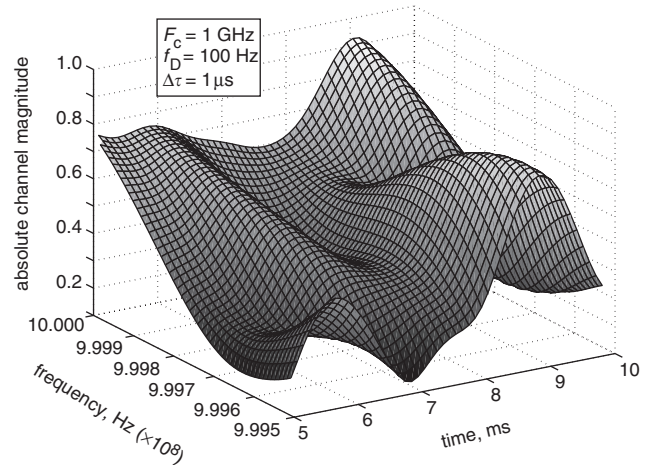


Fig. 1 Typical OFDM channel through a 500ms-by-500kHz window

It is known that the two-dimensional autocorrelation envelope of the time and frequency fading processes observed in the OFDM channel given by (4) can be separated if the normalised autocorrelation functions of all $\gamma_k(t)$ are the same [14]. In this case it can be shown that [19]

$$R_H(\Delta t, \Delta f) = \sigma_H^2 R_t(\Delta t) R_f(\Delta f) \quad (5)$$

where

$$R_t(\Delta t) = J_0(2\pi \Delta t f_D) \quad (6)$$

$$R_f(\Delta f) = \frac{J_0(2\pi \Delta \tau \Delta f)}{1 + (2\pi \Delta \tau \Delta f)^2} \quad (7)$$

Comparing (1) and (3) we observe that for any given carrier f_m , the processes $h(t)$ and $H(t, f_m)$ have similar statistics, since the terms $e^{-j2\pi f_m \tau_k}$ only affect the phases of $\gamma_k(t)$. Indeed, in both cases, $R_t(\Delta t)$ is parameterised only by the Doppler spread f_D . In other words, the time-domain fading process observed at each carrier of an OFDM system is equivalent to that observed in a low-data-rate single-carrier system.

On the other hand, for a fixed instant t , all $\gamma_k(t)$ are constant, so that the frequency-domain fading process depends on the instantaneous distribution of the

magnitudes of $\gamma_k(t)$. The fading observed across the carriers in an OFDM system is therefore not the same as the time-domain fading process observed in narrowband single-carrier systems, despite having the same (Rayleigh) distribution. This is rather evident from the fact that the autocorrelation function $R_f(\Delta f)$ is parameterised both by the Doppler f_D spread and the delay spread $\Delta\tau$.

As an illustration to the fact that a typical OFDM channel is not accurately described by a block-fading model, consider the downlink of a wireless system operating at the 5 GHz band towards a mobile with relative speed of 11 km/h, so that the Doppler frequency spread experienced is 50 Hz.

If the system is single-carrier and transmits symbols taken from an 8PSK constellation with a data-rate of 10 Mbit/s, the symbol period is of approximately 0.3 μ s. Assuming Rayleigh fading, this means that the autocorrelation of channel samples taken every, say, 100 symbols is virtually 1. In other words, the channel is quasi-static and the block-fading assumption holds.

However, in a typical urban environment with a delay spread of 3 μ s [18], the single carrier system employing STBC will experience intersymbol interference. Suppose that in order to combat intersymbol interference a 512-carrier OFDM system is set up. The symbol period in this system is approximately 0.1536 ms and the carrier separation will be around 6.5 kHz. In this case, the normalised time domain autocorrelation of channel samples taken at every symbol drops to 0.9999. This figure indicates that the channel can be considered constant over the transmission of a symbol. Over the transmission of an eight-symbol long STBC, e.g. \mathbf{G}_3 and \mathbf{G}_4 , the figure falls to 0.9907 and for samples taken every 16 symbols (block-length of codes \mathbf{G}_5 – \mathbf{G}_8) the number is 0.9631. Similarly, while the autocorrelation of channel samples taken at every carrier is 1, the figure drops to 0.9976 for samples taken every 8 carriers, and to 0.9903 for samples taken every 16 carriers.

In other words, even in typical urban channels with relatively small Doppler frequencies and moderate delay spreads, the fading experienced by OFDM signals is slightly (yet not negligibly) different in both the time domain (from symbol to symbol) and the frequency (from carrier to carrier) domain.

Two important things are learned from the numerical example given above. The first is that typical channel variations are still slow enough that the channel can be considered constant over a symbol. This implies that the problem of co-channel (carrier-to-carrier) interference due to symbol distortion can be neglected (in fact dealt with independently) when considering decoding of STBCs in OFDM systems. The second is that while smooth channel variation is observed across a block of a few symbols, this is typically not strong enough to favour space–time diversity methods designed for the fast-fading scenario.

Given the above discussion, the model used throughout this paper to describe the fading process faced by OFDM signals in the time (across successive symbols) and frequency (across adjacent carriers) domains is that of highly correlated, symbol-by-symbol and carrier-by-carrier Rayleigh fading. The accuracy of this model is further enforced by the fact that a guard interval between consecutive symbols is commonly used in OFDM systems.

In the following, it will also be assumed that the channels observed across different transmit antennas (space domain) are statistically independent and identically distributed (i.i.d.).

3 Decoding of Alamouti STBC in non-quasi-static channels

Block space–time diversity schemes in OFDM systems can be applied both in space–time and space–frequency arrangements. In a space–time arrangement, the STBC structure is applied independently at each carrier, so that codewords are transmitted over a number of transmit intervals. In a space–frequency arrangement, codewords are transmitted over a number of adjacent carriers simultaneously, so that at each transmit interval a new coded block is transmitted.

It was shown in [11] and [8] that depending on channel conditions (determined by the delay spread $\Delta\tau$ and the Doppler frequency f_D) and on system parameters such as the symbol period Δt and the distance between adjacent carriers Δf , one arrangement might yield better results than the other. These works show that in fact these two arrangements are equivalent in the sense that the performances of the space–time and space–frequency arrangements in a certain channel are equivalent to those of corresponding space–frequency and space–time arrangements in another channel where the fading processes in frequency and time domains have inverted properties. For these reasons we shall hereafter, without loss of generality, use notation and language only based on a space–time framework, i.e., assuming that the STBC is applied to every carrier using N -transmit antennas. It is clear that the results hold also for OFDM systems employing space–time codes in a space–frequency arrangement.

3.1 Conventional combiner

Let S_i denote the data symbol transmitted by the i th transmit antenna (through any arbitrary carrier) and $H_i[t]$ its channel state at the t th time interval.

As discussed in Section the preceding, it is assumed that $H_i[t]$ is constant during the transmission of S_i , but changes to $H_i[t+1]$ at the next transmit interval. In this case, the received signals at times t and $t+1$ become

$$Y[t] = H_1[t]S_1 + H_2[t]S_2 + W[t] \quad (8)$$

$$Y[t+1] = H_2[t+1]S_1^* - H_1[t+1]S_2^* + W[t+1] \quad (9)$$

where W denotes zero-mean random white Gaussian complex noise with independent in-phase and quadrature components each with variance $\sigma_{W/2}^2$.

Assuming that the channel is perfectly estimated, the question that arises is what combination of received signals and channel values must be used in order to obtain orthogonal estimates of S_i .

In previous works [8–12], the difference between $H_i[t]$ and $H_i[t+1]$ is neglected and the pairs $(H_1[t]^*, H_2[t])$ and $(H_2[t]^*, -H_1[t])$ are used to combine the received signals for estimates of the transmit symbols, resulting in:

$$\begin{aligned} \hat{S}_1 &= H_1[t]^* Y[t] + H_2[t] Y[t+1]^* \\ &= (|H_1[t]|^2 + H_2[t]H_2[t+1]^*)S_1 + H_2[t](H_1[t]^* \\ &\quad - H_1[t+1]^*)S_2 + W[t]H_1[t]^* + W[t+1]^*H_2[t] \end{aligned} \quad (10)$$

$$\begin{aligned} \hat{S}_2 &= H_2[t]^* Y[t] - H_1[t] Y[t+1]^* \\ &= (|H_2[t]|^2 + H_1[t]H_1[t+1]^*)S_2 + H_1[t](H_2[t]^* \\ &\quad - H_2[t+1]^*)S_1 + W[t]H_2[t]^* - W[t+1]^*H_1[t] \end{aligned} \quad (11)$$

This is equivalent to what a system designed for the block-fading channel and employing the Alamouti scheme would obtain as estimates in the presence of a non-quasi-stationary fading process. Equations (10) and (11) clearly show that if the channel values vary slightly from one transmission instant to the other ($H_1[t] \neq H_1[t+1]$), the estimates obtained with the conventional linear decoder for Alamouti's scheme are (strictly) no longer orthogonal.

Let us quantify the average effect, in terms of intrasymbol interference, that is caused by such a block-fading-oriented decoding in the presence of a non-quasi-static fading channel. Assuming that the fading rate is not too fast, consecutive samples of the same process can be related by an auto-regressive model [12]

$$H_i[t+1] = \alpha_i H_i[t] + V_i[t] \quad (12)$$

where V_i are zero-mean complex Gaussian random variables uncorrelated to H_i , and α_i is given by

$$\alpha_i = \frac{E[H_i[t]H_i[t+1]^*]}{\sigma_{H_i}^2} = R_i(\Delta t) = J_0(2\pi\Delta t f_D) \quad (13)$$

In the equation above, $E[\cdot]$ denotes the expected value, $\sigma_{H_i}^2$ the variance (power) of the i th diversity branch H_i and Δt is the time lag between successive transmissions at t and $t+1$, that is, the symbol period.

Strictly speaking, α_i take complex values. However, if the channel variation from instant t to $t+1$ is not too fast, α_i have very small phase and can be approximated by real numbers [19]. Then, the following approximation holds:

$$\begin{aligned} & E\left[|H_i[t]|^2 + H_j[t]H_j[t+1]^*\right] \\ & \approx (\sigma_{H_i}^2 + J_0(2\pi\Delta t f_D)\sigma_{H_i}^2) \end{aligned} \quad (14)$$

Next, define the normalised interference coefficient ρ_i as the average of the absolute value of the i th channel variation from $H_i[t]$ to $H_i[t+1]$:

$$\rho_i = \sqrt{\frac{E[|H_i[t] - H_i[t+1]|]^2}{\sigma_{H_i}^2}} \quad (15)$$

The coefficient that multiplies the interfering symbol S_j in the estimate of S_i then becomes:

$$E[H_j[t](H_i[t] - H_i[t+1])^*] = \sigma_{H_i}\rho_i E[H_j[t]] \quad (16)$$

Since $E[H_j[t]]$ is simply the average value of a Rayleigh random variable, we finally have [20]

$$E[H_j[t](H_i[t] - H_i[t+1])^*] = \sigma_{H_i}\rho_i\sigma_{H_j}\Gamma\left(\frac{3}{2}\right) \quad (17)$$

The coefficients ρ_i are clearly dependent on the speed with which the channel varies from symbol to symbol (fading rate), which can be measured by the autocorrelation of the channel. In order to simplify the notation it is hereafter assumed that the diversity branches are balanced, i.e., $\sigma_{H_1}^2 = \sigma_{H_2}^2 = \sigma_H^2$, so that the subscript i can be dropped from α_i and ρ_i .

Expanding (15) and using the approximation given in (14), the following is obtained for Rayleigh fading channels with variance σ_H^2 and the normalised autocorrelation function given by (6):

$$\rho = \sqrt{2(1-\alpha)} \quad (18)$$

Although it was assumed that the channel $H_i[t]$ varies slightly to $H_i[t+1]$, it is found that (18) accurately models the relationship between the interference coefficient ρ and the autocorrelation of the channel even for higher fading rates. Figure 2 illustrates this fact comparing results

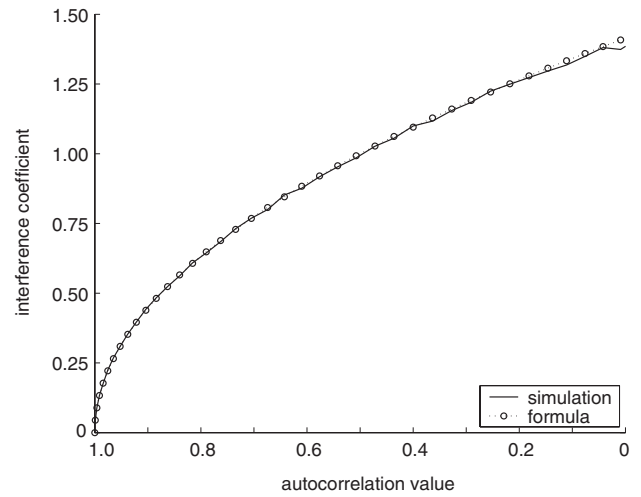


Fig. 2 Variation of the normalised interference coefficient against the channel normalised autocorrelation

obtained with (18) to those obtained directly from (15) through computer simulations.

It is noticeable from Fig. 2 that ρ grows quickly even for small deviations from the ideal autocorrelation $\alpha=1$ (corresponding to a perfect block-fading channel). This implies that in a channel that is slightly non-block-fading, significant intrasymbol interference is caused by the linear combiner described through equations (10) and (11).

Indeed, the performance of the STBC in a non-quasi-static fading channel with balanced diversity branches is driven by the following average signal-to-interference-plus-noise ratio (SINR)

$$\xi = \frac{(1 + J_0(2\pi\Delta t f_D))\sigma_H^2\sigma_S^2}{2\sigma_W^2 + \sqrt{2(1 - J_0(2\pi\Delta t f_D))}\Gamma(\frac{3}{2})\sigma_H^2\sigma_S^2} \quad (19)$$

where σ_S^2 is the average energy of the transmit symbols (energy of the signal constellation).

While the diversity order achieved by the code is given by $(1 + J_0(2\pi\Delta t f_D)) \leq 2$, the denominator term $\sqrt{2(1 - J_0(2\pi\Delta t f_D))}\Gamma(1.5)$ in the denominator accounts for the orthogonality loss caused by the inadequacy of the decoder facing a non-block-fading channel. In noisy scenarios ($\sigma_W^2 \geq \sigma_S^2$), or in channels with absolutely no variation within blocks ($\rho = 0$) (19) reduces to

$$\xi = \frac{\sigma_H^2\sigma_S^2}{\sigma_W^2} \quad (20)$$

Equation (20) indicates that, in noisy non-block-fading channels, the Alamouti space-time diversity scheme with the above decoder performs approximately as well as in a block-fading channel.

In typical OFDM systems however, the channel is non-quasi-static ($\rho \neq 0$) and the case of practical interest for high-data-rate wireless mobile communication is that of a low-noise scenario ($\sigma_W^2 < \sigma_S^2$). In this case, a floor on the SINR inversely proportional to the term $\sqrt{2(1 - J_0(2\pi\Delta t f_D))}\Gamma(1.5)$ is reached. This indicates that there is a signal-to-noise-ratio beyond which no improvement is observed on the bit error rate performance of the space-time coded OFDM system, as observed in [8–11]. Identical results are obtained if the pair of channel states $(H_1[t+1]^*, H_2[t+1])$ and $(H_2[t+1]^*, -H_1[t+1])$ are respectively used to obtain estimates for S_1 and S_2 .

Taking another look at (10) and (11) we see that if the fading rate is too fast that $E[H_i[t]H_i[t+1]^*] = 0$, ρ reduces to $\sigma_H\sqrt{2}$ and $E[|H_i[t]|^2 + H_j[t]H_j[t+1]^*]$ to σ_H^2 . Consequently, the diversity order deteriorates to 1 (no diversity gain), and the interference increases to a maximum level of $\sqrt{2}\Gamma(1.5)$. In this case, the SINR becomes:

$$\xi = \frac{\sigma_H^2\sigma_S^2}{2\sigma_W^2 + \sqrt{2}\Gamma(\frac{3}{2})\sigma_H^2\sigma_S^2} \quad (21)$$

This scenario corresponds to the Alamouti scheme in a fast fading environment. We emphasise that this analysis is purely illustrative since, if the fading rate is that high, channel estimation itself would represent a difficult problem, not to mention that channel variation within the transmission of a symbol (which is a cause of co-channel interference in OFDM systems) could not be neglected. Equation (21), however, is helpful to illustrate why linear combining is no longer effective if STBCs are used in fast fading channels. In fact, it is known that in fast-fading channels other space-time coding techniques, which nevertheless require more complex maximum likelihood decoding, outperform STBCs [4].

Fast-fading channels, which are related to very high mobile speeds and low carrier frequencies, are not typical in OFDM systems, but can be artificially created by the use of interleaving. However, in typical urban scenarios with small Doppler frequencies, very long interleavers would be required in order for fast-fading space-time codes to yield significant gain, at the cost of higher complexity and sacrificing real-time decoding.

In contrast, the technique presented in this paper is aimed at enabling real-time, low-complexity (symbol-by-symbol) maximum likelihood decoding of orthogonal STB coded OFDM systems in typical non-quasi-static (yet not fast) fading channels. The analysis and application of space-time coding schemes to OFDM systems in fast-fading scenarios are therefore out of the scope of this work.

3.2 Maximum diversity combiner

Alternatively to the conventional combiner shown in the previous subsection, $(H_1[t]^*, H_2[t+1])$ and $(H_2[t]^*, -H_1[t+1])$ could be respectively used to obtain estimates for S_1 and S_2 , ensuring that the maximum diversity order is achieved. In this case we have:

$$\begin{aligned} \hat{S}_1 &= H_1[t]^*Y[t] + H_2[t+1]Y[t+1]^* \\ &= (|H_1[t]|^2 + |H_2[t+1]|^2)S_1 + (H_2[t]H_1[t]^* \\ &\quad - H_2[t+1]H_1[t+1]^*)S_2 + W[t]H_1[t]^* \\ &\quad + W[t+1]^*H_2[t+1] \end{aligned} \quad (22)$$

$$\begin{aligned} \hat{S}_2 &= H_2[t]^*Y[t] - H_1[t+1]Y[t+1]^* \\ &= (|H_1[t+1]|^2 + |H_2[t]|^2)S_2 + (H_1[t]H_2[t]^* \\ &\quad - H_1[t+1]H_2[t+1]^*)S_1 + W[t]H_2[t]^* \\ &\quad - W[t+1]^*H_1[t+1] \end{aligned} \quad (23)$$

If $H_i[t]$ and $H_j[t]$ are uncorrelated, zero-mean, complex Gaussian processes with variances $\sigma_{H_i}^2$ and $\sigma_{H_j}^2$, respectively, the process $H_i[t]H_j[t]^*$ is also a zero-mean complex Gaussian process with variance $\sigma_{H_i}^2\sigma_{H_j}^2$. Therefore, we have:

$$\begin{aligned} E[|H_i[t]H_j[t]^* - H_i[t+1]H_j[t+1]^*|] \\ = \sigma_{H_i}\sigma_{H_j}\sqrt{2(1 - J_0(2\pi\Delta t f_D))} \end{aligned} \quad (24)$$

Again, for balanced diversity branches, the average SINR resulting from the use of this combiner is given by:

$$\xi = \frac{2\sigma_H^2\sigma_S^2}{2\sigma_W^2 + \sqrt{2(1 - J_0(2\pi\Delta t f_D))}\sigma_H^2\sigma_S^2} \quad (25)$$

Equation (25) means that although full diversity is achieved, this decoder is also not orthogonal in the presence of a non-block-fading channel. In addition, with this combiner, an even larger amount of intrasymbol interference is observed since the factor $\Gamma(1.5) \approx 0.886$ is no longer in the denominator of the SINR formula. Consequently, higher error floors on the bit-error-rate curves will also be observed at the low noise region.

3.3 Orthogonal combiner

From the results obtained with the analysis shown in Sections 3.1 and 3.2, we infer that pursuing a combiner that delivers maximum diversity is not necessarily the right approach in a non-block-fading channel.

Let us combine $Y[t]$ and $Y[t+1]$ using the pairs $(H_1[t+1]^*, H_2[t])$ and $(H_2[t+1]^*, -H_1[t])$, to obtain \hat{S}_1 and \hat{S}_2 , respectively. In this case we have:

$$\begin{aligned} \hat{S}_1 &= (H_1[t]H_1[t+1]^* + H_2[t]H_2[t+1]^*)S_1 \\ &\quad + W[t]H_1[t+1]^* + W[t+1]^*H_2[t] \end{aligned} \quad (26)$$

$$\begin{aligned} \hat{S}_2 &= (H_1[t]H_1[t+1]^* + H_2[t]H_2[t+1]^*)S_2 \\ &\quad + W[t]H_2[t+1]^* - W[t+1]^*H_1[t] \end{aligned} \quad (27)$$

Unlike all the methods above, this combiner is capable of completely decoupling the estimates \hat{S}_1 and \hat{S}_2 , regardless of the values of H_i . As before, in case the fading process is slow enough, the average signal-to-noise-ratio (SNR) obtained becomes:

$$\xi = \frac{2J_0(2\pi\Delta t f_D)\sigma_H^2\sigma_S^2}{2\sigma_W^2} \quad (28)$$

Equation (28) suggests that the full orthogonality provided with this decoder comes at the expense of reducing the diversity order achieved, which is given by $2J_0(2\pi\Delta t f_D)$, clearly less than 2 for $f_D \neq 0$. In a block-fading scenario ($\Delta t f_D \approx 0$) however, as with all other methods above, this combiner reduces to the conventional (maximum-ratio-like) linear combiner for Alamouti's code [1].

In all the linear combiners discussed here, estimates for the symbols S_1 and S_2 are first obtained and then compared to the Q symbols of the transmit constellation. Therefore, the order of complexity of the corresponding decoders is $O(2Q)$.

In contrast, with strict maximum likelihood decoding, no decoupled estimates of the transmit symbols are computed. Instead, decoding is based on the computation of the metric given below, for all possible pairs of S_1 and S_2 in the constellation, which gives a decoding complexity in the order of $O(Q^2)$.

$$\begin{aligned} D_2 &= |Y[t] - (H_1[t]S_1 + H_2[t]S_2)|^2 + |Y[t+1] \\ &\quad - (H_2[t+1]S_1^* - H_1[t+1]S_2^*)|^2 \end{aligned} \quad (29)$$

In order to illustrate the difference in terms of complexity, consider that an 8PSK constellation is used. Decoding a pair of symbols with strict maximum likelihood requires the computation of 64 metrics. In contrast, any of the linear decoders discussed in this Section requires only 16 comparisons. In other words, linear decoders maintain the complexity of the coded systems at the same level of

maximum likelihood coherent detection of uncoded systems.

3.4 Performance and complexity

In Fig. 3, the performances of the Alamouti scheme with each of the three linear combiners discussed in this section are compared to that obtained with a strict maximum likelihood decoder. The plot shows the bit-error rate at one of the 1024 carriers of an OFDM system transmitting 8PSK symbols with a nominal data rate of 10 Mbit/s in a channel with a Doppler spread of 75 Hz. Given these parameters, each carrier transmits at about 10 kbit/s and the symbol period is around 0.3 ms so that the normalised autocorrelation of the channel sampled at the symbol rate is approximately 0.9948 (see (6)).

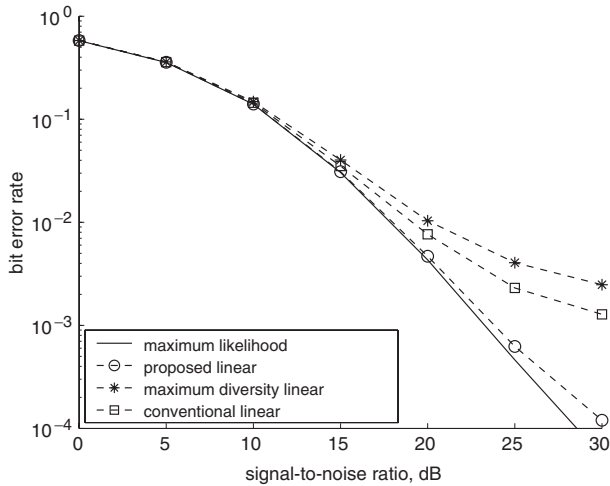


Fig. 3 Performance of Alamouti scheme (3 bits/s/Hz) in a 1024-carrier OFDM system with 8PSK modulation Doppler spread $f_D = 75$ Hz

It is noticeable that the maximum diversity combiner of Sub-section 3.2 performs worse than the conventional combiner of Sub-section 3.1. This result is in accordance with the theoretical analysis given in this Section.

The plot also shows that both the maximum diversity and the conventional (block-fading) decoders exhibit error floors at high SNRs. In contrast, it is seen that the proposed orthogonal linear combiner, which enables symbol-by-symbol maximum likelihood decoding, yields a performance nearly as good as the strict maximum likelihood decoder [3], with no error floor.

On the one hand, the complexity of a strict maximum likelihood decoder for the Alamouti scheme grows with the square of the cardinality of the modulation constellation. On the other hand, higher order modulations imply longer symbols which, for a fixed Doppler spread, result in more interference with the conventional and maximum diversity linear combiners.

This means that the gain of the orthogonal method, in terms of performance enhancement over the other linear combiners, and in terms of complexity reduction over the strict maximum likelihood decoder, is more significant for higher-order modulations.

This fact is illustrated in Fig. 4 where the performances of the Alamouti scheme in the same channel described above but for 10 Mbit/s 1024-carrier OFDM systems with several different modulations are compared.

It is seen that both with the conventional and with the maximum diversity linear combiners, the higher the order of

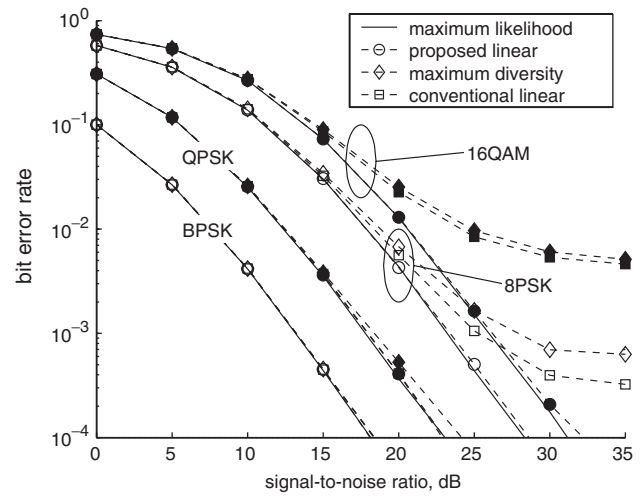


Fig. 4 Performance of Alamouti scheme (1–4 bits/s/Hz) in a 1024-carrier OFDM system with different modulation schemes Doppler spread $f_D = 75$ Hz

the modulation scheme, the higher the error floor reached and the lower the signal-to-noise ratio from where it is established. In contrast, the performance achieved with the orthogonal combiner and symbol-by-symbol maximum likelihood decoding is clearly comparable to that obtained with a strict maximum likelihood decoder.

Note that, for instance, in the case of a 16 QAM constellation, the strict maximum likelihood decoder requires as much as 256 metric-computations in order to decode a single block of two symbols, against only 32 needed with the proposed method! It is also seen that the maximum diversity linear combiner performs slightly worse than the conventional linear combiner.

All the analysis presented here suggests that if low-complexity (linear) maximum likelihood decoding of STBCs in non-quasi-static fading channels is desired, diversity order must be sacrificed in favour of orthogonality. In the following Section we will show that this is the case not only for Alamouti's code but also for Tarokh's generalised STBCs from orthogonal designs [2].

4 Decoding of orthogonal space-time block codes in non-quasi-static channels

The linear orthogonal decoder given in Section 3.3 is restricted to Alamouti's STBC [1]. Although this particular code is known to be the only full rate orthogonal STBC that achieves full diversity over any complex constellation, the result would be incomplete and less important if not applicable to the generalised STBCs from orthogonal designs developed by Tarokh [2, 3].

In fact, although these generalised orthogonal STBCs have reduced rate, it has been shown [21, 22] that they can be combined to form quasi-orthogonal constructions which, together with constellation rotations techniques [23, 24], yield codes with improved performances.

For the above reasons, we are stimulated to extend the above orthogonal linear decoder for Alamouti's scheme in non-quasi-static fading channels to Tarokh's codes under the same conditions.

4.1 Linear decoding based on a squaring method

In this Section, the squaring method proposed in [6] for linearly decoding generalised orthogonal STBCs in

block-fading is extended to the non-quasi-static fading channel. In order to maintain the notation as simple as possible, and without loss of generality, we consider one-receive antenna, N -transmit antenna schemes. The extension to the case of multiple receive antennas is straightforward.

Consider that an orthogonal STBC is applied to a carrier using N transmit antennas. Let \mathbf{G}_N and \mathbf{S} denote the code's orthogonal design matrix and a symbol vector constructed with the set of K transmit symbols (S_1, \dots, S_K) followed by their complex conjugates, as shown in (31).

$$\mathbf{G}_N = \begin{bmatrix} c_{1,1} & \cdots & c_{1,n} & \cdots & c_{1,N} \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ c_{k,1} & \cdots & c_{k,n} & \cdots & c_{k,N} \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ c_{2K,1} & \cdots & c_{2K,n} & \cdots & c_{2K,N} \end{bmatrix} \quad (30)$$

$$\mathbf{S} = [S_1 \quad \cdots \quad S_K \quad S_1^* \quad \cdots \quad S_K^*]^T \quad (31)$$

In STBCs from orthogonal designs [2], every entry $c_{n,k}$ of \mathbf{G}_N is an element of \mathbf{S} multiplied by $+1$ or -1 , such that the rows of \mathbf{G}_N are orthogonal.

Let \mathbf{H} and \mathbf{Y} denote the channel matrix and the received signal vector, respectively.

$$\mathbf{H} = \begin{bmatrix} H_1[t] & \cdots & H_1[t+k] & \cdots & H_1[t+2K-1] \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ H_n[t] & \cdots & H_n[t+k] & \cdots & H_n[t+2K-1] \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ H_N[t] & \cdots & H_N[t+k] & \cdots & H_N[t+2K-1] \end{bmatrix} \quad (32)$$

$$\mathbf{Y} = [Y[t] \quad \cdots \quad Y[t+2K-1]] \quad (33)$$

The n th row of \mathbf{H} consists of time samples of the continuous fading process corresponding to the n th transmit interval, while the k th row corresponds to the spatial samples of these fading processes across the elements of the transmit array of antennas. In other words, in the context of this paper, it is assumed that $H_n[t+k]$ varies slowly with k (along the columns of \mathbf{H}), and quickly with n (along its rows).

The received signals are simply the diagonal entries of the product $\mathbf{G}_N \cdot \mathbf{H}$ plus a noise vector \mathbf{W} .

$$\mathbf{Y} = \text{diag}[\mathbf{G}_N \cdot \mathbf{H}] + \mathbf{W} \quad (34)$$

A squared matrix, analogous to that derived in [6], suitable for linear decoding of STBCs in non-quasi-static fading channels is then obtained by finding a matrix \mathbf{R} such that:

$$\text{diag}[\mathbf{G}_N \cdot \mathbf{H}] = \mathbf{R} \cdot \mathbf{S} \quad (35)$$

Next, define a vector \mathbf{P} constructed with combinations of the received signals and the entries of \mathbf{R} :

$$\mathbf{P} = \mathbf{Y} \cdot \mathbf{R}^* \quad (36)$$

Then, an estimate \hat{S}_k for the k th transmit symbol S_k is obtained by summing the k th element of \mathbf{P} with the complex conjugate of its $(k+K)$ th element:

$$\hat{S}_k = \mathbf{P}_k + \mathbf{P}_{(k+K)}^* \quad (37)$$

If the K different symbols in each block belong to a constellation of cardinality Q , the block is decoded by computing K symbol estimates (37) and then searching the alphabet for those that hold the shortest distance to each estimate. This process involves computing the metric below, Q times for each symbol estimate:

$$|\hat{S}_k - S_q|_{q=1, \dots, Q} \quad (38)$$

Accordingly, this decoder is linear on the received signals and its complexity order is $O(QK)$. This is in contrast with a complexity order of $O(Q^K)$ for the nonlinear strict maximum likelihood decoder which is based on the computation of the metric D_K given below, for all K -tuples of possible vectors in the constellation.

$$D_K = \sum_{k=1}^{2K} |Y[t+k-1] - \text{diag}[\mathbf{G}_N \cdot \mathbf{H}]_k|^2 \quad (39)$$

Note, that unlike the scenario considered here, in block-fading channels the rows of \mathbf{H} are constant. It was shown in [12] that in this case the estimates \hat{S}_k obtained with (37) are orthogonal. In other words, this linear decoder is also a maximum likelihood decoder in block-fading channels.

Unfortunately, just as with the Alamouti scheme, the orthogonality of the symbol estimates S_k are quickly lost in the presence of a non-quasi-static fading channel, such as those seen by the symbols transmitted through each carrier of an OFDM system.

Let us illustrate this with an example. Consider Tarokh's STBC for four antennas (\mathbf{G}_4). In a non-quasi-static fading channel, \mathbf{R} becomes

$$\mathbf{R} = \begin{bmatrix} \mathbf{R}_{1,1} & 0 \\ 0 & \mathbf{R}_{2,2} \end{bmatrix} \quad (40)$$

where

$$\mathbf{R}_{1,1} = \begin{bmatrix} H_1[t] & H_2[t] & H_3[t] & H_4[t] \\ H_2[t+1] & -H_1[t+1] & H_4[t+1] & -H_3[t+1] \\ H_3[t+2] & -H_4[t+2] & -H_1[t+2] & H_2[t+2] \\ H_4[t+3] & H_3[t+3] & -H_2[t+3] & -H_1[t+3] \end{bmatrix} \quad (41)$$

$$\mathbf{R}_{2,2} = \begin{bmatrix} H_1[t+4] & H_2[t+4] & H_3[t+4] & H_4[t+4] \\ H_2[t+5] & -H_1[t+5] & H_4[t+5] & -H_3[t+5] \\ H_3[t+6] & -H_4[t+6] & -H_1[t+6] & H_2[t+6] \\ H_4[t+7] & H_3[t+7] & -H_2[t+7] & -H_1[t+7] \end{bmatrix} \quad (42)$$

The linear combination of received signals yielding the estimate for, say, the first symbol S_1 is:

$$\begin{aligned} \hat{S}_1 = & H_1[t]^* \cdot Y[t] + H_2[t+1]^* \cdot Y[t+1] \\ & + H_3[t+2]^* \cdot Y[t+2] + H_4[t+3]^* \cdot Y[t+3] \\ & + H_1[t+4]^* \cdot Y[t+4] + H_2[t+5]^* \cdot Y[t+5] \\ & + H_3[t+6]^* \cdot Y[t+6] + H_4[t+7]^* \cdot Y[t+7] \end{aligned} \quad (43)$$

Substituting (34) into (44) we have:

$$\begin{aligned} \hat{S}_1 = & \left(\sum_{k=1}^K |H_k[t+k-1]|^2 + \sum_{k=1}^K |H_k[t+k+K-1]|^2 \right) S_1 \\ & + (H_1[t]^* H_2[t] - H_1[t+5]^* H_2[t+5] \\ & + H_1[t+4] H_2[t+4]^* - H_1[t+1] H_2[t+1]^* \\ & + H_3[t+7]^* H_4[t+7] - H_3[t+2]^* H_4[t+2] \\ & + H_3[t+3] H_4[t+3]^* - H_3[t+6] H_4[t+6]^*) S_2 \\ & + (H_1[t]^* H_3[t] - H_1[t+6]^* H_3[t+6] \\ & + H_1[t+4] H_3[t+4]^* - H_1[t+2] H_3[t+2]^* \\ & + H_2[t+5] H_4[t+5]^* - H_2[t+3] H_4[t+3]^* \\ & + H_2[t+1]^* H_4[t+1] - H_2[t+7]^* H_4[t+7]) S_3 \\ & + (H_1[t]^* H_4[t] - H_1[t+7]^* H_4[t+7] \\ & + H_1[t+4] H_4[t+4]^* - H_1[t+3] H_4[t+3]^* \\ & + H_2[t+6]^* H_3[t+6] - H_2[t+1]^* H_3[t+1] \\ & + H_2[t+2] H_3[t+2]^* - H_2[t+5] H_3[t+5]^*) S_4 \end{aligned} \quad (44)$$

Comparing equations (22) and (44) it is readily seen that this extension of the linear decoder originally proposed in [6] is really an extension of the maximum diversity combiner for the Alamouti scheme discussed in Section 3.2. As the latter, this linear decoder reduces to a linear maximum likelihood decoder in block-fading channels. However, in a non-quasi-static fading channel, the symbol estimates obtained with this method are no longer orthogonal and consequently, symbol-by-symbol maximum likelihood decoding is no longer achievable.

4.2 Partially orthogonal linear combiner

In this Section, a partially orthogonal linear decoding method for orthogonal STBCs in fast-fading channels is introduced. In order to gain insight, we start by applying Li's method to the Alamouti STBC. The decoding matrices \mathbf{R} and \mathcal{R} below relate to the pair of equations (22) and (23) and to the pair of equations (26) and (27), respectively.

$$\mathbf{R} = \begin{bmatrix} H_1[t] & H_2[t] & 0 & 0 \\ 0 & 0 & H_2[t+1] & -H_1[t+1] \end{bmatrix} \quad (45)$$

$$\mathcal{R} = \begin{bmatrix} H_1[t+1] & H_2[t+1] & 0 & 0 \\ 0 & 0 & H_2[t] & -H_1[t] \end{bmatrix} \quad (46)$$

The matrix (45), in association with equations (36) and (37) yield equations (22) and (23). Analogously, the matrix (46), together with equations (36) and (37), yield equations (26) and (27). Comparing these two matrices, it is readily seen that the entries $H_1[t]$ and $H_1[t+1]$ are permuted, with the signs + and - preserved at their original positions in both matrices. The same can be seen about the values $H_2[t]$ and $H_2[t+1]$.

Given the results shown in Section 3.3, using the matrix (46) with equations (36) and (37) yields orthogonal estimates of S_1 and S_2 . This suggests that a similar strategy of permutation on the entries of (40) can be applied to derive several modified versions of \mathbf{R} , each related to the estimation of a symbol S_k , free of the interference from any other arbitrary symbol S_q , and *vice versa*.

Hereafter, we shall use the notation $\mathcal{R}_{(k,q)}$ in reference to a matrix obtained from permutations over \mathbf{R} , which yields mutually or orthogonal estimates of the pair of symbols

(S_k, S_q) . The set of permutations over \mathbf{R} so as to obtain $\mathcal{R}_{(k,q)}$ is summarised as follows.

- Select the k th column of $\mathbf{R}_{1,1}$ and the q th column of $\mathbf{R}_{2,2}$;
- Denote these C_k and C_q , respectively;
- Permute the values $H_n[t+i] \in C_k$ and $H_n[t+j] \in C_q$, for all n ;
- Select the q th column of $\mathbf{R}_{1,1}$ and the k th column of $\mathbf{R}_{2,2}$;
- Denote these C_q and C_k , respectively;
- Permute the values $H_n[t+i] \in C_q$ and $H_n[t+j] \in C_k$, for all n .

An estimate of S_k , orthogonal to S_q (denoted $\hat{S}_{k,\bar{q}}$ and read S_k not S_q) is then obtained by substituting $\mathcal{R}_{(k,q)}$ for \mathbf{R} in (36), and then using (37).

Note that in STBCs from orthogonal designs, an even number K of different symbols are encoded in each block, even if an odd number of diversity branches (antennas) is used. Consequently, the matrices $\mathbf{R}_{1,1}$ and $\mathbf{R}_{2,2}$ also have even orders. From this fact, and from the fact that the above algorithm only operates on a pair of columns in $\mathbf{R}_{1,1}$ and $\mathbf{R}_{2,2}$, it is clear that a single permuted matrix \mathcal{R} can be constructed to support the mutually orthogonal decoding of other pairs of symbols (S_u, S_w) such that $(u \neq k, q, w)$ and $(w \neq k, q, u)$. Generally, let \mathcal{K} be a set of mutually exclusive pairs (i, j) such that $i \neq j$ and no other element in \mathcal{K} contains i or j . Then, a single matrix $\mathcal{R}_{\mathcal{K}}$ can be constructed to yield the mutually orthogonal decoding of all pairs in \mathcal{K} .

As an example, let us reconsider Tarokh's \mathbf{G}_4 code. Let $\mathcal{K} = \{(1, 2), (3, 4)\}$. The matrix $\mathcal{R}_{\mathcal{K}}$ that yields the mutually orthogonal decoding of the pair (S_1, S_2) and of the pair (S_3, S_4) is given by

$$\mathbf{R} = \begin{bmatrix} \mathcal{R}_{\mathcal{K}1,1} & 0 \\ 0 & \mathcal{R}_{\mathcal{K}2,2} \end{bmatrix} \quad (47)$$

where

$$\begin{aligned} \mathcal{R}_{\mathcal{K}1,1} = & \begin{bmatrix} H_1[t+5] & H_2[t+5] & H_3[t+5] & H_4[t+5] \\ H_2[t+4] & -H_1[t+4] & H_4[t+4] & -H_3[t+4] \\ H_3[t+7] & -H_4[t+7] & -H_1[t+7] & H_2[t+7] \\ H_4[t+6] & H_3[t+6] & -H_2[t+6] & -H_1[t+6] \end{bmatrix} \end{aligned} \quad (48)$$

$$\begin{aligned} \mathcal{R}_{\mathcal{K}2,2} = & \begin{bmatrix} H_1[t+1] & H_2[t+1] & H_3[t+1] & H_4[t+1] \\ H_2[t] & -H_1[t] & H_4[t] & -H_3[t] \\ H_3[t+3] & -H_4[t+3] & -H_1[t+3] & H_2[t+3] \\ H_4[t+2] & H_3[t+2] & -H_2[t+2] & -H_1[t+2] \end{bmatrix} \end{aligned} \quad (49)$$

If the matrix (47) is used to obtain an estimate of S_1 orthogonal to S_2 ($S_{1,2}$) we have the result given in (50). Note that the term on S_2 vanishes regardless of the channel values.

From equation (50), it is also noticeable that a reduction in the diversity gain attained is expected with this decoder.

This effect is similar to that found in Section 3.3 for the Alamouti code.

$$\begin{aligned}
\hat{S}_{1,2} = & (H_1[t+5]^*H_1[t] + H_2[t+4]^*H_2[t+1] \\
& + H_3[t+7]^*H_3[t+2] + H_4[t+6]^*H_4[t+3] \\
& + H_1[t+1]H_1[t+4]^* + H_2[t]H_2[t+5]^* \\
& + H_3[t+3]H_3[t+6]^* + H_4[t+2]H_4[t+7]^*)S_1 \\
& + (H_1[t+5]^*H_2[t] - H_2[t+4]^*H_1[t+1] \\
& + H_3[t+7]^*H_4[t+2] + H_4[t+6]^*H_3[t+3] \\
& + H_1[t+1]H_2[t+4]^* - H_2[t]H_1[t+5]^* \\
& + H_3[t+3]H_4[t+6]^* + H_4[t+2]H_3[t+7]^*)S_2 \\
& + (H_1[t+5]^*H_3[t] + H_2[t+4]^*H_4[t+1] \\
& + H_3[t+7]^*H_1[t+2] - H_4[t+6]^*H_2[t+3] \\
& + H_1[t+1]H_3[t+4]^* + H_2[t]H_4[t+5]^* \\
& + H_3[t+3]H_1[t+6]^* - H_4[t+2]H_2[t+7]^*)S_3 \\
& + (H_1[t+5]^*H_4[t] - H_2[t+4]^*H_3[t+1] \\
& + H_3[t+7]^*H_2[t+2] - H_4[t+6]^*H_1[t+3] \\
& + H_1[t+1]H_4[t+4]^* - H_2[t]H_3[t+5]^* \\
& + H_3[t+3]H_2[t+6]^* - H_4[t+2]H_1[t+7]^*)S_4
\end{aligned} \tag{50}$$

It is clear that the same matrix (47) can be used to obtain $\hat{S}_{2,1}$, $\hat{S}_{3,4}$ and $\hat{S}_{4,3}$.

Analogously, other partially orthogonal symbol estimates can be obtained by constructing a different set of mutually exclusive pairs \mathcal{K} and then deriving the appropriate permuted matrix $\mathcal{R}_{\mathcal{K}}$. For instance, with $\mathcal{K} = \{(1, 3), (2, 4)\}$ we have:

$$\begin{aligned}
\mathcal{R}_{\mathcal{K}1,1} = & \\
& \begin{bmatrix} H_1[t+6] & H_2[t+6] & H_3[t+6] & H_4[t+6] \\ H_2[t+7] & -H_1[t+7] & H_4[t+7] & -H_3[t+7] \\ H_3[t+4] & -H_4[t+4] & -H_1[t+4] & H_2[t+4] \\ H_4[t+5] & H_3[t+5] & -H_2[t+5] & -H_1[t+5] \end{bmatrix}
\end{aligned} \tag{51}$$

$$\begin{aligned}
\mathcal{R}_{\mathcal{K}2,2} = & \\
& \begin{bmatrix} H_1[t+2] & H_2[t+2] & H_3[t+2] & H_4[t+2] \\ H_2[t+3] & -H_1[t+3] & H_4[t+3] & -H_3[t+3] \\ H_3[t] & -H_4[t] & -H_1[t] & H_2[t] \\ H_4[t+1] & H_3[t+1] & -H_2[t+1] & -H_1[t+1] \end{bmatrix}
\end{aligned} \tag{52}$$

Using (51) and (52) in (47), the estimates $\hat{S}_{1,3}$, $\hat{S}_{3,1}$, $\hat{S}_{2,4}$ and $\hat{S}_{4,2}$ are obtained.

Finally, we could make $\mathcal{K} = \{(1, 4), (2, 3)\}$ and construct the matrices below, which would yield the estimates $\hat{S}_{1,4}$, $\hat{S}_{4,1}$, $\hat{S}_{2,3}$ and $\hat{S}_{3,2}$.

$$\begin{aligned}
\mathcal{R}_{\mathcal{K}1,1} = & \\
& \begin{bmatrix} H_1[t+7] & H_2[t+7] & H_3[t+7] & H_4[t+7] \\ H_2[t+6] & -H_1[t+6] & H_4[t+6] & -H_3[t+6] \\ H_3[t+5] & -H_4[t+5] & -H_1[t+5] & H_2[t+5] \\ H_4[t+4] & H_3[t+4] & -H_2[t+4] & -H_1[t+4] \end{bmatrix}
\end{aligned} \tag{53}$$

$$\begin{aligned}
\mathcal{R}_{\mathcal{K}2,2} = & \\
& \begin{bmatrix} H_1[t+3] & H_2[t+3] & H_3[t+3] & H_4[t+3] \\ H_2[t+2] & -H_1[t+2] & H_4[t+2] & -H_3[t+2] \\ H_3[t+1] & -H_4[t+1] & -H_1[t+1] & H_2[t+1] \\ H_4[t] & H_3[t] & -H_2[t] & -H_1[t] \end{bmatrix}
\end{aligned} \tag{54}$$

The fact that the intrasymbol interference between any pair of arbitrary symbols can be avoided suggests that a fully orthogonal linear decoder must be possible, since the interference due to each symbol individually can be identified. This would enable symbol-by-symbol maximum likelihood decoding with a linear combiner.

In the following Sub-section it is shown that a systematic design of such combiners indeed exists.

4.3 Fully orthogonal linear combiner

It was shown in the preceding Section that in the case of the Alamouti scheme, the conventional non-orthogonal combiner quickly loses performance in the presence of a non-quasi-static fading channel, but that a simple modification on that linear combiner can be applied yielding orthogonal estimates of the transmit symbols. This enables symbol-by-symbol maximum likelihood decoding in non-quasi-static fading channels, significantly reducing the decoding complexity of maximum likelihood decoding.

In the case of the generalised orthogonal STBCs, the complexity issue is even more severe. For instance, if 16 QAM is used with the code \mathbf{G}_4 , maximum likelihood decoding requires the computation of the metric given in (39), 65 536 times in order to decode a single block of four symbols!

While it is clear that (optimal) strict maximum likelihood decoding can be prohibitively complex in some cases, it is also clear that the partially orthogonal decoder derived above for generalised orthogonal STBCs in non-quasi-static channels cannot fully eliminate the error floors at high SNR scenarios. Although low-complexity is highly desired, a sub-optimal linear decoder is only acceptable if performance is not significantly sacrificed. This can be achieved if symbol-by-symbol maximum likelihood detection can be performed with a linear combiner.

In this Section, the partially orthogonal decoder of Section 4.2 is extended into a fully orthogonal linear combiner for symbol-by-symbol maximum likelihood decoding of STBCs in non-quasi-static fading channels. The proposed decoder is obtained by successively applying the partial interference cancellation scheme described above as follows:

Step 1. Choose an arbitrary symbol S_q and construct all matrices $\mathcal{R}_{(k,q)}$.

Step 2. Use the partially orthogonal method of Section 4.2 to obtain the estimates $\hat{S}_{k,\bar{q}}$, for all $k \neq q$.

Step 3. Construct a vector $\mathbf{Y}_{\bar{q}}$ consisting of the estimates $\hat{S}_{k \neq q, \bar{q}}$, with 0 at the q th position, augmented by their respective complex conjugates (This emulates the receive signal vector of a 'virtual' encoded block in which the symbol S_q faces a null channel at every transmission).

Step 4. Construct a new channel matrix $\mathbf{H}_{\bar{q}}$ whose entries are the coefficients multiplying each S_k in all elements of $\mathbf{Y}_{\bar{q}}$ (These coefficients are simply combinations of the entries of the original channel matrix and therefore are known).

Step 5. Return to step 1 if more than two symbols are still mixed in each $\hat{S}_{k \neq q, \bar{q}}$.

Step 6. Apply the orthogonal decoder of Section 3.3 to all the pairs of mixed symbols.

Step 7. Repeat the procedure from the beginning, sequentially choosing different symbols to be cancelled so as to obtain orthogonal estimates of all other symbols.

At every iteration of the above algorithm, one symbol is eliminated and at the end of the process, two symbol estimates, fully orthogonal to all other symbols and to one another, are obtained. Repeating the process making different choices of which symbols are eliminated successively will yield orthogonal estimates for all S_k .

Before we proceed with an example, let us emphasise that although the systematic construction of orthogonal symbol estimates is described here by this iterative algorithm, the resulting combiner need not be iterative! Indeed, the algorithm provided need only be run *algebraically*, using the set of variables $\{H_n[t], H_n[t+1], \dots, H_n[t+2K-1]\}$ for all $n = 1, \dots, N$, a single time for each of Tarokh's full diversity STBCs, i.e. those where $K=N$ e.g. $\mathbf{G}_4, \mathbf{G}_8$. Once the symbolic calculations are done, we obtain, for the corresponding STBC, a set of $2K$ scalar functions $F_k(\mathbf{H})$ on the channel matrix \mathbf{H} (explicit forms are omitted here due to space limitations), which are used to directly combine the received signals onto orthogonal estimates for all S_k , as shown below.

$$\hat{S}_k = \sum_{i=1}^K F_i(\mathbf{H})Y[t+i-1] + \sum_{i=K+1}^{2K} F_i(\mathbf{H})Y[t+i-1]^* \quad (55)$$

Algebraic formulas obtained for the full diversity codes ($\mathbf{G}_4, \mathbf{G}_8$ etc.) work for the other codes where $K > N$ because these lower-order codes are simply truncations of the latter. For instance, \mathbf{G}_3 is a truncation of \mathbf{G}_4 and $\mathbf{G}_5, \mathbf{G}_6$ and \mathbf{G}_7 are truncations of \mathbf{G}_8 [2]. Consequently, \mathcal{R} is always a $2K$ -by- $2K$ matrix, only with a number of zero entries when codes of lower diversity order are used. For example, only one calculation is required for the code \mathbf{G}_4 (whose results are also valid for \mathbf{G}_3) and another for \mathbf{G}_8 , with the results applicable to $\mathbf{G}_5, \mathbf{G}_6$ and \mathbf{G}_7 .

As with the method in [6], as well as the partially orthogonal combiner of Section 4.2, the fully orthogonal method proposed here is linear on the received signals. Consequently, symbol-by-symbol maximum likelihood decoding of STBCs from orthogonal designs in non-block-fading channels [2] is achieved with (38).

Let us proceed with an example. We continue with \mathbf{G}_4 so that previous equations can be re-used. First, the interference due to an arbitrary symbol, e.g. S_4 (step 1), is removed from the estimates of the remaining symbols, in this case, S_1, S_2 and S_3 . In order to simplify the notation, let us denote the matrix \mathcal{R} built with $\mathcal{K} = \{(1, 2), (3, 4)\}$, $\mathcal{K} = \{(1, 3), (2, 4)\}$ and $\mathcal{K} = \{(1, 4), (2, 3)\}$ respectively by $\mathcal{R}_1, \mathcal{R}_2$ and \mathcal{R}_3 . Then, we have (step 2)

$$\begin{aligned} \hat{S}_{1,\bar{4}} &= \mathcal{R}_3^{*(1)} \cdot \mathbf{Y} + \mathcal{R}_3^{(5)} \mathcal{R}_3^{(5)} \cdot \mathbf{Y}^* \\ \hat{S}_{2,\bar{4}} &= \mathcal{R}_2^{*(2)} \cdot \mathbf{Y} + \mathcal{R}_2^{(6)} \cdot \mathbf{Y}^* \\ \hat{S}_{3,\bar{4}} &= \mathcal{R}_1^{*(3)} \cdot \mathbf{Y} + \mathcal{R}_1^{(7)} \cdot \mathbf{Y}^* \end{aligned} \quad (56)$$

where $\langle i \rangle$ denotes the i th column.

Next, a new 'receive vector' is constructed using these estimates (step 3), giving:

$$\mathbf{Y}_{\bar{4}} = \left[\hat{S}_{1,\bar{4}}, \hat{S}_{2,\bar{4}}, \hat{S}_{3,\bar{4}}, 0, \hat{S}_{1,\bar{4}}^*, \hat{S}_{2,\bar{4}}^*, \hat{S}_{3,\bar{4}}^*, 0 \right] \quad (57)$$

The elements of this vector can be written as

$$\begin{aligned} \mathbf{Y}_{1,\bar{4}} &= \hat{S}_{1,\bar{4}} = H_1[t]_4 S_1 + H_2[t]_4 S_2 + H_3[t]_4 S_3 \\ \mathbf{Y}_{2,\bar{4}} &= \hat{S}_{2,\bar{4}} = H_2[t+1]_4 S_1 - H_1[t+1]_4 S_2 + H_4[t+1]_4 S_3 \\ \mathbf{Y}_{3,\bar{4}} &= \hat{S}_{3,\bar{4}} = H_3[t+2]_4 S_1 - H_4[t+2]_4 S_2 - H_1[t+2]_4 S_3 \\ \mathbf{Y}_{4,\bar{4}} &= 0 \\ \mathbf{Y}_{5,\bar{4}} &= \hat{S}_{1,\bar{4}}^* = H_1[t+4]_4 S_1^* + H_2[t+4]_4 S_2^* + H_3[t+4]_4 S_3^* \\ \mathbf{Y}_{6,\bar{4}} &= \hat{S}_{2,\bar{4}}^* = H_2[t+5]_4 S_1^* - H_1[t+5]_4 S_2^* + H_4[t+5]_4 S_3^* \\ \mathbf{Y}_{7,\bar{4}} &= \hat{S}_{3,\bar{4}}^* = H_3[t+6]_4 S_1^* - H_4[t+6]_4 S_2^* - H_1[t+6]_4 S_3^* \\ \mathbf{Y}_{8,\bar{4}} &= 0 \end{aligned} \quad (58)$$

Clearly, each of the quantities $H_n[t+k-1]_4$ above is a known function of the entries $H_n[t+k-1]$ of \mathbf{H} . For instance, $H_1[t]_4$ is given by:

$$\begin{aligned} H_1[t]_4 &= H_1[t+7]^* H_1[t] + H_2[t+6]^* H_2[t+1] \\ &\quad + H_3[t+6]^* H_3[t+2] + H_4[t+4]^* H_4[t+3] \\ &\quad + H_1[t+3] H_1[t+4]^* + H_2[t+2] H_2[t+5]^* \\ &\quad + H_3[t+1] H_3[t+6]^* + H_4[t] H_4[t+7]^* \end{aligned} \quad (59)$$

Using these quantities, a new channel matrix $\mathbf{H}_{\bar{4}}$ and the corresponding squared matrix $\mathbf{R}_{\bar{4}}$ are built (step 4).

Another symbol, say S_3 , is then chosen to be cancelled next and the calculations are repeated (steps 5 and 1). New sets $\mathcal{K} = \{(1, 3)\}$ and $\mathcal{K} = \{(2, 3)\}$ are constructed, for which new permuted matrices $\mathcal{R}_{2,\bar{4}}$ and $\mathcal{R}_{3,\bar{4}}$, are derived. The partially orthogonal decoder will then yield (step 2):

$$\begin{aligned} \mathbf{Y}_{1,\{\bar{4},\bar{3}\}} &= \hat{S}_{1,\{\bar{4},\bar{3}\}} = \mathcal{R}_{2,\bar{4}}^{*(1)} \cdot \mathbf{Y}_{\bar{4}} + \mathcal{R}_{2,\bar{4}}^{(5)} \cdot \mathbf{Y}_{\bar{4}}^* \\ &= H_1[t]_{\{\bar{4},\bar{3}\}} S_1 + H_2[t]_{\{\bar{4},\bar{3}\}} S_2 \end{aligned} \quad (60)$$

$$\begin{aligned} \mathbf{Y}_{2,\{\bar{4},\bar{3}\}} &= \hat{S}_{2,\{\bar{4},\bar{3}\}} = \mathcal{R}_{3,\bar{4}}^{*(2)} \cdot \mathbf{Y}_{\bar{4}} + \mathcal{R}_{3,\bar{4}}^{(6)} \cdot \mathbf{Y}_{\bar{4}}^* \\ &= H_2[t+1]_{\{\bar{4},\bar{3}\}} S_1^* - H_1[t+1]_{\{\bar{4},\bar{3}\}} S_2^* \end{aligned} \quad (61)$$

Once more, the quantities $H_n[t+k-1]_{\{\bar{4},\bar{3}\}}$ are functions of $H_n[t+k-1]_{\bar{4}}$ and, by extension, of $H_n[t+k-1]$. Similarly, $\mathbf{Y}_{\bar{4}}$ are functions of $H_n[t+k-1]$ and \mathbf{Y} . Consequently, $\hat{S}_{1,\{\bar{4},\bar{3}\}}$ and $\hat{S}_{2,\{\bar{4},\bar{3}\}}$ can be written as combinations of the channel values $H_n[t+k-1]$ and the received signals \mathbf{Y} such as in equation (55).

The orthogonal combiner for Alamouti's scheme presented in Section 3.3 can now be used to combine these signals onto orthogonal estimates of S_1 and S_2 , giving (step 6):

$$\hat{S}_{1,\{\bar{4},\bar{3},\bar{2}\}} = H_1[t+1]_{\{\bar{4},\bar{3}\}}^* \mathbf{Y}_{1,\{\bar{4},\bar{3}\}} + H_2[t]_{\{\bar{4},\bar{3}\}} \mathbf{Y}_{2,\{\bar{4},\bar{3}\}}^* \quad (62)$$

$$\hat{S}_{2,\{\bar{4},\bar{3},\bar{2}\}} = H_2[t+1]_{\{\bar{4},\bar{3}\}}^* \mathbf{Y}_{1,\{\bar{4},\bar{3}\}} - H_1[t]_{\{\bar{4},\bar{3}\}} \mathbf{Y}_{2,\{\bar{4},\bar{3}\}}^* \quad (63)$$

Finally, it is evident that \hat{S}_1 and \hat{S}_2 can also be written as combinations of the channel values and the received signals as in (55). A similar procedure is followed to orthogonally decode S_3 and S_4 (step 7). The orthogonal estimates obtained are finally used in (38) for maximum likelihood decoding of all S_k .

We emphasise that this iterative follow-up of the method here shown is merely illustrative. As pointed out before, every $\mathbf{Y}_{\{\bar{q}, \dots\}}$ obtained along the process is a combination of $H_n[t+k-1]_{\{\bar{q}, \dots\}}$ and S_k , and every $H_n[t+k-1]_{\{\bar{q}, \dots\}}$ is a combination of the original channel values $H_n[t+k-1]$. In a usual implementation, once the channel is estimated (55) is used to obtain orthogonal estimates of S_k as linear combinations of the received signals with the functions $\mathbf{F}_k(\mathbf{H})$ as coefficients.

Note that if the channel is quasi-stationary, the columns of \mathbf{H} are identical, and so too are the matrices $\mathcal{R}_{\mathcal{K}}$, since permutations over \mathbf{R} do not change it at all. In this case, the first K functions $\mathbf{F}_{k=1, \dots, K}(\mathbf{H})$ are simply the entries of the k th column of $\mathcal{R}_{1,1}$ (multiplied by a constant) and the remaining K functions $\mathbf{F}_{k=K+1, \dots, 2K}(\mathbf{H})$ are the entries of the k th column of $\mathcal{R}_{2,2}$ (multiplied by the same constant).

Given the arguments above, it is clear that the proposed linear combiner is really a generalisation of the maximum-ratio-like linear combiner, known for STBCs in block-fading channels [1, 6], to the more general case of non-quasi-static fading channels.

5 Simulation results

In this Section, simulation results comparing the proposed linear maximum likelihood decoder with the conventional linear decoder proposed in [6] and with the strict maximum likelihood decoder, are presented.

First, Fig. 5 compares the performances of Tarokh STBCs \mathbf{G}_3 and \mathbf{G}_6 implemented at the carriers of a 1024-carrier OFDM system transmitting BPSK symbols.

It is important to stress that the low-order (BPSK) modulation scheme (not so common in OFDM systems) is used here only so that the complexity of the strict maximum likelihood decoder is not too high, so as to enable the comparison with the proposed linear (symbol-by-symbol) maximum-likelihood method. In fact, with \mathbf{G}_6 and QPSK, the strict maximum likelihood decoder would require the computation of as much as 4096 metrics per block.

The conventional decoder assumes that the channel is block-fading and uses only the channel estimates relative to the first symbol vector ($H_i[r]$). The maximum diversity decoder is that of [6] extended to the non-quasi-static fading channel as discussed in Section 4.1, and the proposed decoder is that presented in Section 4.3.

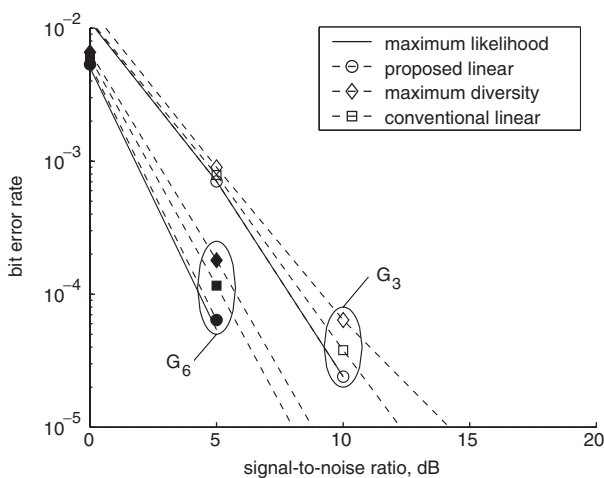


Fig. 5 Bit error rate of Tarokh \mathbf{G}_3 and \mathbf{G}_6 (0.5 bit/s/Hz) STBCs with linear and maximum likelihood decoders
Doppler frequency $f_D = 50$ Hz

The data rate of the OFDM system is 10 Mbit/s. If the loss due to the guard interval is ignored, this gives a data rate of approximately 10 kbit/s per carrier. Given that both \mathbf{G}_3 and \mathbf{G}_6 are rate 1/2 codes, this results in a transmission rate of 0.5 bit/s/Hz and a symbol period of approximately 0.05 ms.

Perfectly uncorrelated channels were assumed at different transmit antennas and no OFDM co-channel interference due to carrier offset, symbol distortion, cross-talk or other effects were considered.

At each carrier, the channel has a Doppler frequency of $f_D = 50$ Hz. This Doppler spread corresponds, for instance, to that observed in the downlink of an OFDM system operating on the 5 GHz band towards a mobile with a speed of about 11 km/h [18]. From (6), these parameters result in an autocorrelation of 0.9999 over samples taken at the symbol rate, 0.9907 over samples taken every eight symbols (\mathbf{G}_3 block size) and 0.9631 over samples taken every 16 symbols (\mathbf{G}_6 block size).

In such moderate fading channels, the error floors resulting from using the conventional linear decoder occur at levels too low to be observed in the plot. Nevertheless, it is seen that the proposed decoder outperforms the other linear decoders and that the conventional (block-fading) combiner outperforms the maximum diversity combiner. These results are similar to those observed in Fig. 4 for Alamouti scheme.

Figure 5 also shows that the performance of the linear decoder is very close to that of a strict maximum likelihood decoder. In fact, for \mathbf{G}_3 , the curves for these two decoders fully overlap.

From all the discussions of Sections 3 and 4, it is expected that the gain provided by the proposed method in comparison to other linear methods is more noticeable the more severe the channel variation within a block. Given a fixed data-rate this can occur if, for instance, more carriers are used in the OFDM system (longer symbols), the mobile moves faster (larger Doppler frequency), a higher order modulation scheme is applied (longer symbols) or codes for a larger number of antennas are implemented (longer blocks).

This can be inferred from Fig. 4 but becomes more evident in Fig. 6. The plot compares the performances of Tarokh \mathbf{G}_3 code in a channel with a Doppler frequency of $f_D = 75$ Hz (mobile speed of 16 km/h at 5 GHz) in systems employing BPSK, QPSK, and 8PSK modulation. The

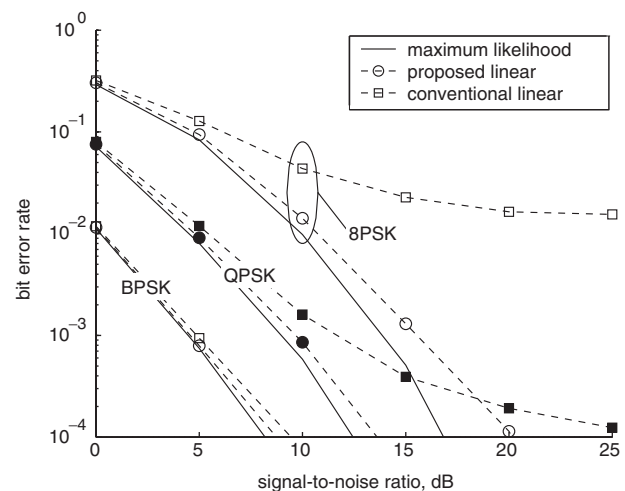


Fig. 6 Bit error rate of Tarokh \mathbf{G}_3 (0.5–1.5 bit/s/Hz) STBC with linear and maximum likelihood decoders
Doppler frequency $f_D = 75$ Hz

remaining parameters (data rate, number of carriers, etc.) are all the same as described above. Curves for the maximum diversity combiner are omitted since it was already shown that this decoder actually performs worse than the conventional linear combiner at higher SNR's.

This Figure is analogous to Fig. 4 and the general results are very similar. It is confirmed that with severe channel variation within a block, in this case caused by the increase in modulation order, a higher error floor is reached by the linear decoder. As before, it is seen that the more severe the fading rate of the channel, the lower the signal-to-noise ratio where the error floor is reached. However, Fig. 4 clearly shows that the gain of the proposed combiner with respect to the conventional linear combiner is larger at more severe scenarios, since the orthogonality guaranteed by the method enables symbol-by-symbol maximum likelihood decoding. It is also evident that the price paid in terms of loss of diversity order, compared to a strict maximum likelihood decoder, is more significant in channels with faster variations within a block. Still, it is important to emphasise that the proposed decoder is expressively simpler than the strict maximum likelihood decoder. For instance, for G_3 with 8PSK, a strict maximum likelihood decoder needs to compute and compare 512 metrics per block, against only 24 required by the proposed linear method.

We emphasise that the fast fading scenario is not the target of the technique presented and is rather unlikely to be faced by high data rate OFDM systems in an urban environment. In fact, higher fading rates are associated to faster mobility, which in turn imply that the mobile must be in an area with less obstacles and at a longer distance from the base. Consequently, the typical delay spreads channels with fast mobility are likely to be small, so that the fading process in the frequency domain is less severe, which in turn imply that the STBC could rather be implemented in the alternative space-frequency arrange. This issue has been investigated in [11] and [8].

However, an important result shown in Fig. 4 is that error floors are not reached with the proposed linear decoder even at higher fading rates. This is expected since the proposed combiner ensures the orthogonality of symbol estimates at the receiver, regardless of the channel values across a block (provided that the channel is known). This can be further appreciated from the results of an illustrative simulation where the channel is considered to be constant over a symbol and vary randomly from symbol to symbol. The results are shown in Fig. 7.

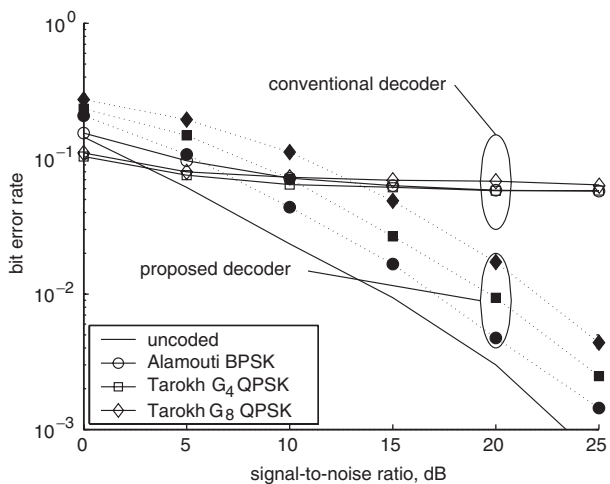


Fig. 7 Bit error rate of Alamouti and Tarokh G_4 and G_8 STBCs with linear decoders in an illustrative fast-fading scenario

As explained in Section 3.1, this case is purely illustrative not only because it cannot occur naturally (a truly fast-fading channel would cause severe co-channel interference in a OFDM system), but also because accurate channel estimation could not be achieved. Indeed, if the fast-fading scenario is artificially caused by interleaving, it is clear that other types of space-time codes specifically designed for such channels, e.g. [5], are preferred. In this case, however, not only complexity is significantly increased, but realtime decoding can no longer be performed.

The illustrative simulation in a 'fast-fading' scenario has, nevertheless, the merit of demonstrating that the proposed combiner *defacto* ensures the orthogonality of symbol estimates at the receiver and translates the variation of the channel across an encoding block onto diversity loss.

The conventional decoder and the proposed maximum likelihood linear decoder were applied to Alamouti's and to Tarokh G_4 and G_8 codes in a fully random fading channel. Data-rate, number of carriers, symbol period and central carrier frequency are irrelevant information in this case, since f_D is made infinite (the channel matrix H is completely random). Results with BPSK and QPSK modulations are shown together so that all schemes run with the same transmission rate of 1 bit/s/Hz.

It is seen that while the conventional linear decoder shows a completely flat performance, the proposed linear decoder still exhibits a bit error rate performance that follows the same inclination of an uncoded (interference-free) system. While the fact that an error floor is never reached demonstrates that the proposed linear combiner yields symbol estimates which are fully orthogonal, the fact that the inclination of all curves are the same as that of an uncoded system shows that STBCs from orthogonal design with symbol-by-symbol maximum likelihood decoding do not provide any diversity gain. In other words, it can be said that the all diversity gain is traded for orthogonality in the presence of an infinitely fast-fading scenario.

In addition, the fact that the bit error rate curves translate horizontally as the order of the code (number of antennas) increases, also indicates that the proposed linear maximum likelihood decoder increases noise in the process of decoupling the symbols estimates. This noisy characteristic, which in the slower fading rates of practical interest is small enough not to be observed, is indeed typical of zero-forcing techniques such as the one presented.

6 Conclusions

A new, low-complexity linear maximum likelihood decoder for STBCs from orthogonal designs in non-quasi-static fading channels has been proposed. Non-quasi-static fading is faced by OFDM systems (which are often required to mitigate inter-symbol interference caused by frequency selectivity) in mobile channels even if these systems transmit at high-data-rates.

The proposed linear decoder is capable of obtaining orthogonal estimates for all simultaneously transmitted symbols regardless of the rate of variation of the channel within each block, provided that channel estimates are available throughout the block. It was shown that this is achieved mainly at the cost of diversity gain and some increase in noise, both relatively small for the scenarios of interest (moderate mobility).

The proposed decoder can be seen as an extension of the linear decoder proposed in [6] to the more general

non-quasi-static channel. In fact, the proposed linear decoder also has complexity of the order $O(QK)$ and reduces to the latter in the presence of block-fading channels.

Since orthogonal estimates of transmit symbols are obtained with the proposed combiner even in the presence of channel variations within encoding blocks, symbol-by-symbol maximum likelihood is achievable with significantly lower complexity compared to a strict maximum likelihood decoder. Given its low complexity and good performance, the proposed linear maximum likelihood decoder is suitable for implementation in low-cost, small, portable terminals, enabling OFDM systems to employ space-time transmit diversity on the downlink to mobile units, which is known to be a bottleneck of high-data-rate mobile wireless systems. Moreover, since the method presented reduces to the optimal linear technique known for the block-fading case, the proposed decoder can be used to support seamless indoor-to-outdoor transition in high-data rate wireless network services ('hot-spots').

Simulation results were presented which demonstrate the performance improvement offered by the proposed decoder, compared to other known linear decoders with similar complexity. It was also shown that in typical fading scenarios, the performance achieved with the proposed linear decoder is close to that of a strict maximum likelihood decoder. Therefore, the proposed decoder offers a low-cost alternative to nonlinear maximum likelihood decoding which might be impractical especially if QAM constellations (common in OFDM systems) and codes with higher diversity orders are employed.

We hope this work will inspire the pursuit of similar solutions for other STBCs already proposed [21–24], as well as others still to be discovered.

Although actual performance is dependent on the accuracy of channel estimation, as is the case of all coherent decoding schemes (including strict maximum likelihood), the correlation between time- and frequency-domain fading processes observed in OFDM channels can be exploited to yield low-complexity, accurate channel estimation for OFDM systems [14, 15]. Indeed, the work presented here can be combined with these low-complexity channel estimation techniques for OFDM systems, so as to yield iterative decoding/channel estimation algorithms which exploit the knowledge of channel states not only at pilot symbols, but at decoded symbols. Further investigations on this matter are currently being carried out by the authors.

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8 References

- 1 Alamouti, S. M.: 'A simple transmit diversity technique for wireless communications', *IEEE J. Sel. Areas Commun.*, 1998, **11**, (8), pp. 1451–1458
- 2 Tarokh, V., Jafarkhani, H., and Calderbank, A.R.: 'Space-time block codes from orthogonal designs', *IEEE Trans. Inf. Theory*, 1999, **45**, (5), pp. 1456–1467
- 3 Tarokh, V., Jafarkhani, H., and Calderbank, A.R.: 'Space-time block coding for wireless communications: Performance results', *IEEE J. Sel. Areas Commun.*, 1999, **17**, (2), pp. 451–460
- 4 Tarokh, V., Seshadri, N., and Calderbank, A.R.: 'Space-time codes for high data rate wireless communications: Performance criteria in the presence of channel estimation errors, mobility and multiple paths', *IEEE Trans. Commun.*, 1998, **47**, (2), pp. 199–207
- 5 Firmanto, W., Vucevic, B., Yuan, J., and Chen, Z.: 'Space-time turbo trellis coded modulation for wireless data communications', *EURASIP J. Appl. Signal Process.*, 2002, **2002**, pp. 459–470
- 6 Li, X., Luo, T., Yue, G., and Yin, C.: 'A squaring method to simplify the decoding of orthogonal space-time block codes', *IEEE Trans. Commun.*, 2001, **49**, (10), pp. 1700–1703
- 7 Cimini, L.J.: 'Analysis and simulation of a digital mobile channel using orthogonal frequency division multiplexing', *IEEE Trans. Commun.*, 1985, **33**, (7), pp. 665–675
- 8 Cheng, J., Wang, H., and Cheng, S.: 'Space-time block coded transmit diversity for OFDM systems in mobile channels'. Proc. IEEE Personal, Indoor and Mobile Radio Communications (PIMRC'02), Lisbon, Portugal, 15–18 Sept. 2002
- 9 Li, Y.G., Chuang, J.C., and Sollenberger, N.R.: 'Transmitter diversity for OFDM systems and its impact on high-rate data wireless networks', *IEEE J. Sel. Areas Commun.*, 1999, **17**, (7), pp. 1233–1243
- 10 Liu, Z., Giannakis, G.B., Barbarossa, S., and Scaglione, A.: 'Transmit antennae space-time block coding for generalized OFDM in the presence of unknown multipath', *IEEE J. Sel. Areas Commun.*, 2001, **19**, (7), pp. 1352–1364
- 11 Lee, K.F., and Williams, D.B.: 'A space-frequency transmitter diversity technique for OFDM systems'. Proc. IEEE Global Telecommunications Conf. (Globecom'03), San Francisco, USA, 27 Nov.–1 Dec. 2003, pp. 1473–1477
- 12 Liu, Z., Ma, X., and Giannakis, G.B.: 'Space-time coding and Kalman filtering for time-selective fading channels', *IEEE Trans. Commun.*, 2002, **50**, (2), pp. 183–186
- 13 Edfors, O., Sandell, M., van de Beek, J.J., Wilson, S.K., and Borjesson, P.O.: 'OFDM channel estimation by singular value decomposition', *IEEE Trans. Commun.*, 1998, **46**, (7), pp. 931–939
- 14 Chang, M.-X., and Su, Y.T.: 'Model-based channel estimation for OFDM signals in Rayleigh fading', *IEEE Trans. Commun.*, 2002, **50**, (4), pp. 540–544
- 15 Deneire, L., Vandenameele, P., van der Perre, L., Gyselinckx, B., and Engels, M.: 'Space-time diversity systems based on constellation precoding', *IEEE Trans. Commun.*, 2003, **51**, (2), pp. 135–140
- 16 Tarokh, V., and Jafarkhani, H.: 'A differential detection scheme for transmit diversity', *IEEE J. Sel. Areas Commun.*, 2000, **18**, (7), pp. 1169–1174
- 17 Jafarkhani, H., and Tarokh, V.: 'Multiple transmit antenna differential detection from generalized orthogonal designs', *IEEE Trans. Inf. Theory*, 2001, **47**, (6), pp. 2626–2631
- 18 Lee, J.S., and Miller, L.E.: 'CDMA Systems Engineering Handbook' (Artech House, Boston, London, 1998)
- 19 Jakes, W.C.: 'Microwave mobile communications' (Wiley, New York, NY, USA, 1974)
- 20 Proakis, J.G.: 'Digital communications' (Mc-Graw-Hill, New York, NY, USA, 2000, 4th edn.)
- 21 Jafarkhani, H.: 'A quasi-orthogonal space-time block code', *IEEE Trans. Commun.*, 2001, **49**, (1), pp. 1–4
- 22 Ming, C., Tao, W., and Kangsheng, C.: 'A simple design method of orthogonal/quasi-orthogonal space-time block codes'. Proc. IEEE 14th Int. Symp. on Personal, Indoor and Mobile Radio Communications (PIMRC'03), Beijing, China, 7–10 Sept. 2003
- 23 Tirkkonen, O., and Hottinen, A.: 'Square-matrix embeddable space-time block codes for complex signal constellations', *IEEE Trans. Inf. Theory*, 2002, **48**, (2), pp. 384–395
- 24 Hassanpour, N., and Jafarkhani, H.: 'A class of full diversity space-time codes'. Proc. IEEE Global Telecommunications Conf. (Globecom'03), San Francisco, CA, USA, Dec. 2003