Exact BER Analysis for Square Orthogonal Space-Time Block Codes on Arbitrary Fading Channels with Imperfect Channel Estimation

Lennert Jacobs and Marc Moeneclaey
Ghent University, TELIN Department, DIGCOM Group
Sint-Pietersnieuwstraat 41, B-9000 Gent, Belgium
{Lennert.Jacobs,Marc.Moeneclaey}@telin.ugent.be

Abstract—In this contribution, we examine the effect of imperfect channel estimation on the bit error rate (BER) performance of square orthogonal space-time block codes (OSTBCs). The propagation channels from the $N_t$ transmit antennas to each of the $N_r$ receive antennas are assumed to be affected by (possibly correlated) flat block fading with an arbitrary distribution. The transmitted symbols belong to a PAM or QAM constellation.

The resulting average BER can easily be written as an expectation over $4N_s N_r$ real-valued random variables, but the computing time needed for its numerical evaluation increases exponentially with $N_t$ and $N_r$. However, we point out that expressing the BER in terms of the distribution of the Frobenius norm of the channel matrix, allows to reduce the BER expression for any number of receive antennas to an expectation over only $2N_s + 2$ random variables (with $N_s$ denoting the number of information symbols in the OSTBC matrix). Due to the high diversity order of OSTBCs, the numerical evaluation of this expectation is much less time-consuming than a straightforward computer simulation. Moreover, the presented BER expression is useful not only when the fading distribution is given in closed form, but also when only experimental data (e.g. a histogram) on the fading are available.

I. INTRODUCTION

In order to improve the error performance of wireless communication systems, several diversity techniques have been proposed to provide the receiver with independent replicas of the transmitted signal. Spatial diversity can be exploited by using multiple transmit and/or receive antennas. Combined with proper space-time coding, a maximum diversity order of $N_t N_r$ (with $N_t$ and $N_r$ denoting the number of transmit and receive antennas, respectively) can be achieved. Orthogonal space-time block codes (OSTBCs) [1]–[3] achieve this full spatial diversity, and their maximum-likelihood (ML) detection algorithm reduces to symbol-by-symbol detection based only on linear processing at the receiver. Therefore, these codes are a very appealing transmit diversity technique. Under the assumption of perfect channel knowledge (PCK), the bit error rate (BER) performance of OSTBCs has been studied extensively in e.g. [4]–[7].

In realistic wireless systems, however, the channel state information (CSI) is not a priori available and the receiver has to estimate the channel response, which inevitably results in a performance degradation. Usually, the system performance is investigated under the assumption of Rayleigh fading: in case of minimum mean-square error (MMSE) channel estimation, an analytical expression for the BER of OSTBCs with pulse amplitude modulation (PAM) or quadrature amplitude modulation (QAM) constellations, was derived in [8]; for orthogonal space-time block coded systems employing $M$-ary phase-shift keying ($M$-PSK) modulation, analytical BER expressions as well as the tight Chernoff bound were given in [9]; By means of an eigenvalue approach, asymptotical pairwise error probability (PEP) expressions were derived in [10] for quite general STBCs with coherent and noncoherent receivers; an exact closed-form expression for the PEP of both orthogonal and non-orthogonal space-time codes in the case of least-squares channel estimation was obtained in [11] by means of characteristic functions. However, from the PEP only an upper bound on the BER can be derived, which in a fading environment does not converge to the true BER at high signal-to-noise ratio (SNR). Expressions for the exact decoding error probability (DEP) of OSTBCs were presented in [12], for the case of fading channels with an arbitrary distribution, but the analysis is only applicable to PSK constellations. Although the impact of imperfect channel estimation on the performance of OSTBCs has been examined extensively, exact BER expressions for OSTBCs with PAM and QAM constellations on generalized fading channels with imperfect channel estimation (ICE) are still lacking in the open literature.

In this contribution, we provide an exact BER analysis for square OSTBCs with PAM or QAM constellations on arbitrary flat-fading channels with ICE. We assume a linear pilot-based channel estimation method, with the well-known least-squares estimation and linear MMSE estimation as special cases. The resulting BER expression involves an expectation over at most $2N_s + 2$ variables (with $N_s$ denoting the number of information symbols in the OSTBC matrix), that allows numerical evaluation with a computing time that is independent of the number of receive antennas. The presented BER expression can be used when the fading distribution of the Frobenius norm of the channel matrix is available in closed form or as an experimentally obtained histogram, and allows for a faster evaluation than by means of straightforward computer simulation, for practical values of the operating BER.
Throughout this paper, the superscripts $T$ and $H$ represent the vector (matrix) transpose and conjugate transpose, respectively. $\mathbb{R}\{x\}$, $\mathbb{E}\{x\}$ denote the real part, the imaginary part and the expected value of $x$, respectively, while $\|X\|$ and $\text{tr}(X)$ refer to the Frobenius norm and the trace of $X$, respectively.

II. SIGNAL MODEL

Let us consider a multiple-input multiple-output (MIMO) wireless communication system with $N_t$ transmit antennas and $N_r$ receive antennas. The propagation channels between each transmit and receive antenna pair are affected by flat fading with an arbitrary distribution. The data transmission is organized in frames consisting of $K_p$ known pilot symbols and $K$ coded data symbols per transmit antenna; the pilot symbols enable channel estimation at the receiver. Since the complex fading gains are distributed according to an arbitrary distribution, the data matrices $X$ and $Y$ are variables with variance $\sigma^2$. Both least-squares and linear MMSE estimation satisfy (6) with $\alpha = 1$ and $\alpha = K_p E_p/(K_p E_p + N_0)$, respectively [14].

Examples of square OSTBCs satisfying (4) are the $2 \times 2$ Alamouti code [1] and the $4 \times 4$ codes given in [2, eq. 40], [3, eq. 62] and [13, eq. 41].

Assuming that $K$ is a multiple of $K_n$, the transmitted data symbol matrix consists of $K/K_n$ coded symbol matrices $C(k)$, with $k$ denoting the block index. In this way, we can write $A = \sqrt{E_s}[C(1), \ldots, C(K/K_n)]$. Considering a normalized information symbol constellation $\mathbb{E}[|s|^2] = 1$, it follows from (3) that $E_s$ denotes the average energy of the transmitted coded symbols:

$$\frac{1}{N_tK_p}\mathbb{E}\|A\|^2 = E_s. \quad (5)$$

Similarly, the average energy of the pilot symbols is $E_p$.

III. PILOT-BASED CHANNEL ESTIMATION

Using the known pilot matrix $A_p$ and the corresponding received signal matrix $R_p$, the receiver can estimate the channel matrix $H$. We assume orthogonal pilot sequences, i.e. $A_p A_p^H = K_p E_p I_{N_t}$, and consider linear channel estimates of the form

$$\hat{H} = \frac{\alpha}{K_p E_p} R_p A_p^H, \quad (6)$$

with $\alpha \in \mathbb{R}$, such that $\hat{H}$ can be decomposed into the sum of two statistically independent contributions:

$$\hat{H} = \alpha H + N. \quad (7)$$

It can be shown that the entries of $N = (\alpha/(K_p E_p)) W_p A_p^H$ are ZMCSCG random variables, whose real and imaginary parts have a variance $\sigma^2 = \alpha^2 N_0/(2 K_p E_p)$. Hence, when conditioned on the channel $H$, the channel estimate $\hat{H}$ is a complex Gaussian random matrix with mean $\alpha H$ and diagonal covariance matrix with diagonal elements $2 \alpha^2 N_0$. Both least-squares and linear MMSE estimation satisfy (6) with $\alpha = 1$ and $\alpha = K_p E_p/(K_p E_p + N_0)$, respectively [14].

An accurate channel estimate can be obtained by allocating a large total energy to pilot symbols. Increasing the total energy of the pilot symbols, however, reduces the symbol energy $E_s$ available for data transmission. Denoting by $E_b$ and $\gamma = E_p/E_s$ the energy per information bit and the ratio of the pilot energy $E_p$ versus the symbol energy $E_s$, we have

$$E_s = \frac{K}{K + \gamma K_p} \rho \log_2(M) E_b, \quad (8)$$

where $\rho = N_0/(N_t K_c)$ and $M$ denote the code rate and the number of constellation points, respectively.

IV. ML DETECTION

When the receiver knows the channel matrix $H$, ML detection is known to be the optimal detection algorithm for the data matrices $C(k)$

$$\hat{C}(k) = \arg\min_C \| R'(k) - \sqrt{E_s} H C \|^2, \quad (9)$$

where $R'(k)$ is the received signal matrix corresponding to the transmitted symbol matrix $C(k)$, and the minimization is over the valid code matrices satisfying (2). However, in realistic applications the channel state information is not known, and the receiver has to estimate the channel. We consider a
mismatched receiver that uses $\hat{H}$ in the same way an ML receiver would apply $H$ (we omit the index $k$ for notational convenience):

$$\hat{C} = \arg \min_C \left\| R' - \sqrt{E_c} \hat{H} C \right\|^2. \quad (10)$$

Taking (3) into account, the detection algorithm (10) for the information symbols $s_i$ reduces to symbol-by-symbol detection:

$$\hat{s}_i = \arg \min_{s_i} \left| u_i - s_i \right|, \quad 1 \leq i \leq N_s,$$

where the decision variables $u_i$ are given by

$$u_i = \text{tr} \left( C_i^H \hat{H}^H R' + R'^H \hat{H} C_i \right),$$

$$\lambda \sqrt{E_c} \left\| \hat{H} \right\|^2. \quad (12)$$

V. Bit Error Rate Analysis

In this contribution, we consider square $M$-QAM transmission with Gray mapping, which reduces to $\sqrt{M}$-PAM transmission for both the in-phase and quadrature information bits. Because of the rotational symmetry of the $M$-QAM constellation and assuming that the symbol vectors $s = (s_1, \ldots, s_{N_s})$ are equally likely, it can be verified that the BERs related to the in-phase and quadrature bits of $s_i$ are equal, and independent of the index $i$. Hence, the average BER of the OSTBC equals the BER of the in-phase information bits related to the transmitted symbol $s_i$, irrespective of $i$. The latter BER can be obtained by averaging the conditional BER $\text{BER}_{i,R}(H, \hat{H})$ (which is related to the in-phase information bits of $s_i$ and conditioned on the channel $H$ and the channel estimate $\hat{H}$). In this way, the BER of the OSTBC is given by

$$\text{BER} = \mathbb{E}_{H, \hat{H}} \left[ \text{BER}_{i,R}(H, \hat{H}) \right]. \quad (13)$$

Denoting by $\text{BER}_{i,R}(s, H, \hat{H})$ the conditional BER related to the in-phase information bits of $s_i$ (conditioned on $H$ and the symbol vector $s$), $\text{BER}_{i,R}(H, \hat{H})$ is given by

$$\text{BER}_{i,R}(H, \hat{H}) = \frac{1}{M^{N_s}} \sum_{s \in \Psi_{s \neq s_i}} \text{BER}_{i,R}(s, H, \hat{H}), \quad (14)$$

where $\Psi$ denotes the normalized $M$-QAM constellation. Referring to the projection (on the real axis) of the decision area of a QAM symbol $b$, for which the in-phase component $b_{\text{R}}$ is different from the in-phase component $s_i$ of the transmitted symbol $s_i$, as the decision region of $b_{\text{R}}$, we can write $\text{BER}_{i,R}(s, H, \hat{H})$ as

$$\text{BER}_{i,R}(s, H, \hat{H}) = \sum_{b_{\text{R}} \in \Psi_{b_{\text{R}}}} \frac{N(s_{i,R}, b_{\text{R}})}{\log_2 M} \text{BER}_{i,R}(s_{i,R}, b_{\text{R}}, H, \hat{H}), \quad (15)$$

where $\Psi_{b_{\text{R}}}$ is the set consisting of the real parts of the constellation points. $N(s_{i,R}, b_{\text{R}})$ represents the Hamming distance between the in-phase bits of the transmitted symbol $s_i$ and the in-phase bits of the detected symbol $b$, and $\text{BER}_{i,R}(s_{i,R}, b_{R}, H, \hat{H})$ is the probability that the real part of the decision variable $u_i$ is located inside the decision area of $b_R$ (when the transmitted symbol vector $s$, the channel $H$ and the channel estimate $\hat{H}$ are known):

$$P_{i,R}(s, b_R, H, \hat{H}) = \Pr \left[ \hat{s}_{i,R} = b_R \mid s, H, \hat{H} \right]. \quad (16)$$

Taking (1) and (2) into account, expanding the real part of the decision variable (12) yields:

$$u_{i,R} = u'_{i,R} + \frac{\Re \left\{ \text{tr} \left( C_i^H \hat{H}^H W' + W'^H \hat{H} C_i \right) \right\}}{\lambda \sqrt{E_c} \left\| \hat{H} \right\|^2},$$

$$\leq \frac{\left\| h - \hat{h} \right\|}{\left\| h \right\|}. \quad (17)$$

where $u'_{i,R}$ is a function of the transmitted symbol vector $s$, the channel matrix $H$ and the estimated channel matrix $\hat{H}$. The second term in (17) represents zero-mean Gaussian noise with variance $N_0/(2\lambda E_c \left\| \hat{H} \right\|^2)$. Let $d_1(b_R)$ and $d_2(b_R)$ denote the boundaries of the projection of the decision area of $b$ on the real axis, with $d_1(b_R) < d_2(b_R)$; we set $d_1(b_R) = -\infty$ ($d_2(b_R) = \infty$) if $b$ is a left (right) outer constellation point. In this way, (16) reduces to

$$P_{i,R}(s, b_R, H, \hat{H}) = Q_1 - Q_2,$$

where the quantities $Q_k, k \in \{1, 2\},$ are given by

$$Q_k = Q \left( \sqrt{2\lambda E_c} \left\| \hat{H} \right\| \left( d_k(b_R) - u'_{i,R} \right) \right).$$

Here, $Q(.)$ denotes the Gaussian Q-function, defined as

$$Q(x) \equiv \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp \left( -\frac{x^2}{2} \right) dx. \quad (20)$$

Hence, the BER (13) of the OSTBC can be expressed as

$$\text{BER} = \int \text{BER}_{i,R}(H, \hat{H}) p(H) p(\hat{H}) dH d\hat{H}, \quad (21)$$

where $p(\hat{H})$ is the joint Gaussian pdf of $N_i N_r$ complex-valued random variables, with mean $\alpha H$ and diagonal covariance matrix with diagonal elements $2\sigma_n^2$, and $p(H)$ represents the arbitrary joint distribution of $N_i N_r$ complex-valued fading gains. Although the general expression (21), which allows correlation between the components of $H$, is conceptually simple, it is not well suited for numerical evaluation as its evaluation requires taking the expectation over $4N_i N_r$ real-valued variables; hence, the associated computing time increases exponentially with the number of transmit and receive antennas.

In order to avoid the computational complexity corresponding to the evaluation of (21), we will manipulate (21) into an expectation over only $2N_s + 2$ variables. Therefore, we introduce the $(2N_i N_r \times 1)$ real-valued column vectors $\hat{H}$ and $H$, which consist of all elements of $\hat{H}$ and $H$, respectively:

$$\hat{h} = [\hat{h}_{1, R}^T, \hat{h}_{1, I}^T, \cdots, h_{N_r, R}^T, h_{N_r, I}^T]^T,$$

$$h = [h_{1, R}^T, h_{1, I}^T, \cdots, h_{N_r, R}^T, h_{N_r, I}^T]^T, \quad (22)$$

where $\hat{h}$ and $h$ are the estimated and true channel coefficients, respectively.
with \( \hat{h}_{i,R} + j\hat{h}_{i,I} \) and \( h_{i,R} + jh_{i,I} \) denoting the \( i \)-th column of \( \hat{H} \) and \( H \), respectively. In this way, it can be shown that \( u_{i,R}' \) reduces to

\[
u_{i,R}' = \frac{1}{\lambda \|H\|} \sum_{j=1}^{N} s_{j,R} \left( \begin{bmatrix} \hat{h}^T g_{i,R}(j) \\ -s_{j,1} (\hat{h}^T g_{i,1}(j)) \end{bmatrix} \right),
\]

where the \((2N_t N_c \times 1)\) column vectors \( g_{i,R}(j) \) and \( g_{i,1}(j) \) are given by

\[
\begin{align*}
g_{i,R}(j) &= M_{i,R}(j) h \\
g_{i,1}(j) &= M_{i,1}(j) h,
\end{align*}
\]

and the \((2N_t N_r \times 2N_t N_r)\) matrices \( M_{i,R}(j) \) and \( M_{i,1}(j) \) depend on the coefficient matrices \( C_i, C_i', C'_i \) and \( C_j' \). The term in \( s_{i,R} \) in (23) is the useful term, whereas the term in \( s_{i,1} \) represents interference from the quadrature component of \( s_i \). The terms in \( s_{j,R} \) and \( s_{j,1} (j \neq i) \) represent interference from the in-phase and quadrature components of the transmitted symbols \( s_j \), respectively. For square OSTBCs satisfying (4), it can be shown that, for any given \( i \), the following orthogonality conditions hold:

\[
\begin{align*}
&\left\{ g_{i,R}(j)^T g_{i',R}(j') = 0 \\
g_{i,R}(j)^T g_{i,R}(j') = \lambda^2 \|H\|^2 \delta_{j-j'}, \\
g_{i,1}(j)^T g_{i,1}(j') = \lambda^2 \|H\|^2 \delta_{j-j'},
\end{align*}
\]

where \( \delta_{j-j'} \) denotes the Kronecker delta. Moreover, since \( M_{i,R}(i) = I_{2N_t N_r} \), it is readily verified that \( u_{i,R}' = s_{i,R} \) in case of PCK (\( h = h_c \)).

Let us now define an orthonormal coordinate system with \( 2N_t N_r \) unit vectors \( \{ e_{i,R}(j), e_{i,1}(j), j = 1, \ldots, N_t N_r \} \), with \( e_{i,R}(j) \) and \( e_{i,1}(j) \), \( 1 \leq j \leq N_s \), directed along \( g_{i,R}(j) \) and \( g_{i,1}(j) \), respectively. The projections of \( \hat{h} \) on \( e_{i,R}(j) \) and \( e_{i,1}(j) \) are denoted \( x_{i,R}(j) \) and \( x_{i,1}(j) \), respectively, which yields

\[
\|\hat{H}\|^2 = \|\hat{h}\|^2 = \sum_{j=1}^{N} (x_{i,R}^2(j) + x_{i,1}^2(j)) + z^2,
\]

with

\[
z^2 = \sum_{i=N_s+1}^{N_t N_r} (x_{i,R}^2(j) + x_{i,1}^2(j)).
\]

Because of the specific choice of \( e_{i,R}(j) \) and \( e_{i,1}(j) \) for \( 1 \leq i \leq N_s \) and taking (25) into account, we have

\[
\begin{align*}
x_{i,R}(j) &= \hat{h}^T e_{i,R}(j) = \frac{\hat{h}^T g_{i,R}(j)}{\lambda \|H\|}, \\
x_{i,1}(j) &= \hat{h}^T e_{i,1}(j) = \frac{\hat{h}^T g_{i,1}(j)}{\lambda \|H\|}.
\end{align*}
\]

From (23) and (19) it follows that \( \text{BER}_{i,R}(H, \hat{H}) \) is a function of only \( 2N_s + 2 \) random variables: \( x_{i,R}(j) \) and \( x_{i,1}(j) (1 \leq j \leq N_s) \), \( z \) and \( \|H\| \). Hence, we denote \( \text{BER}_{i,R}(H, \hat{H}) \) by \( \text{BER}_{i,R}(x, z, \|H\|) \), where the vector \( x \) contains the \( 2N_s \) components \( x_{i,R}(j) \) and \( x_{i,1}(j), 1 \leq j \leq N_s \). Taking (7) into account, it can be shown that \( z \) and the entries of \( x \), when conditioned on \( \|H\| \), are independent variables which satisfy the following properties:

- \( x_{i,R}(i) \) is a Gaussian variable with mean \( \alpha \|H\| \) and variance \( \sigma_n^2 \).
- \( x_{i,1}(j) (j \neq i) \) and \( x_{i,1}(j) \) are zero-mean Gaussian variables with variance \( \sigma_n^2 \).
- \( z/\sigma_n \) is distributed according to the chi-distribution with \( 2N_t N_r - 2N_s \) degrees of freedom [15].

Taking the above properties into account, (21) reduces to

\[
\text{BER} = \int \text{BER}_{i,R}(x, z, u) p(x, z | \|H\| = u) p(u) dx dz du. \quad (30)
\]

where \( p(u) \) is the pdf of the Frobenius norm \( \|H\| \) of the channel matrix \( H \). It is important to note that instead of the joint distribution \( p(H) \) of the \( N_t N_r \) complex-valued fading gains, we need only the distribution of the Frobenius norm \( \|H\| \). The evaluation of the BER involves an expectation over only \( 2N_s + 2 \) random variables, which can be evaluated numerically by approximating the \( (2N_s + 2) \)-fold integral (30) by a \( (2N_s + 2) \)-fold sum, running over discretized versions of the continuous variables.

The number of random variables to be considered in the expectation (30) is reduced in the following cases:

1) When the product of the number of transmit and receive antennas equals the number of information symbols in the square OSTBC matrix \( (N_t N_r = N_s) \), the dimension of \( \hat{h} \) is \( 2N_s \), such that the only projections are \( x_{i,R}(j) \) and \( x_{i,1}(j) (1 \leq j \leq N_s) \). Hence, the BER computation involves averaging over only \( 2N_s + 1 \) variables: \( x_{i,R}(j) \) and \( x_{i,1}(j) (1 \leq j \leq N_s) \), \( z' \) and \( \|H\| \). (31)

\[
\|\hat{H}\|^2 = \sum_{i=1}^{N_s} (x_{i,R}^2(j)) + z'^2
\]

and \( z'/\sigma_n \) has a chi-distribution with \( 2N_t N_r - N_s \) degrees of freedom.

The above analysis is not only applicable to complex-valued channel gains. Also real-valued propagation channels (along with PAM constellations) can be considered. The resulting BER is nearly the same as for PAM transmission over complex-valued channels (averaging over \( N_t N_r \) variables, \( x_{i,R}(j) (1 \leq j \leq N_s) \), \( z' \) and \( \|H\| \)), except for the variable \( z'/\sigma_n \) being distributed according to a chi-distribution with \( N_t N_r - N_s \) instead of \( 2N_t N_r - N_s \) degrees of freedom. For specific cases, the number of random variables involved in the expectation can be further reduced. According to [12], only a twofold integral must be computed in the case of M-PSK constellations and arbitrary fading.
Fig. 1. BER for Alamouti’s code on complex-valued Nakagami-\(m\) fading channel, 16-QAM

VI. NUMERICAL RESULTS

To obtain our numerical results, we use Alamouti’s code [1] \((N_t = K_p = N_s = 2)\) and linear MMSE channel estimation with \(E_p = E_s\). The complex-valued MIMO propagation channel consists of \(2N_t\) i.i.d. Nakagami-\(m\) fading channels, with parameters \(m\) and \(\Omega\) [16], where we assume \(\Omega = 1\).

For both the mismatched receiver and the PCK receiver, fig. 1 displays the exact BER curves in case of 16-QAM transmission, for several values of \(m\) and \(N_t\). It is easily derived that the Frobenius norm \(\|H\|\) of the MIMO channel is distributed according to a Nakagami-\(m\) distribution with parameters \(2N_t, m\) and \(2N_t, \Omega\). The BER curves for the mismatched receiver are obtained by numerically computing the expectation (30) over \(2N_s + 2 = 6\) variables, whereas the BER of the PCK receiver involves the expectation over \(\|H\|\) only. Also shown in the figure are straightforward computer simulation results for the mismatched receiver that confirm the result obtained from numerical averaging.

VII. CONCLUSIONS AND REMARKS

In this contribution, we investigated the effect of ICE on the BER performance of square OSTBCs on arbitrary distributed (possibly correlated) flat-fading propagation channels. We assumed a generic pilot-based linear channel estimation method, which includes the well-known least-squares estimation and linear MMSE estimation as special cases. The resulting exact BER expression involves an expectation over only \(2N_s + 2\) variables, irrespective of the number of receive antennas. Moreover, we have shown that for some cases \((N_t, N_r = N_s,\) real-valued channel, \(PAM\) constellation, and combinations thereof\) the number of variables involved in the expectation can be further reduced. Generally, the evaluation of the BER expression requires a numerical integration over at most \(2N_s + 2\) variables. Due to the high diversity order of OSTBCs, straightforward simulations are extremely time-consuming, such that the numerical averaging is to be preferred for BER values of practical interest. Moreover, the BER expression is useful not only when the pdf of the Frobenius norm of the channel matrix is available in closed form but also when only experimental data (e.g. a histogram) on the fading are available.

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