Time Synchronization Issues for Quasi-Orthogonal Space-Time Block Codes

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***Abstract*—This paper deals with some of the problems arising in multiple-input multiple-output (MIMO) antenna systems, when the receiver's clock is not synchronized to the modulator. Considering that the spatial diversity is exploited through quasi-orthogonal space-time block codes (QOSTBC), we examine the influence of the remaining timing offset on the performance of the system, after coarse timing acquisition has been carried out at the receiver. Specifically, by utilizing the statistical properties of the channel paths and simulating the time error as a uniformly distributed random variable, we show that the average trace of the diversity gain matrix (DGM) undergoes a decrease with respect to the ideal case of zero intersymbol interference (ISI).**

***Index Terms*—Space-time block coding, quasi-orthogonal designs, time error detector, wireless communications.**

I. INTRODUCTION

Space-Time Block Coding (STBC) has been extensively used as a means of exploiting spatial diversity in multiple transmit antenna systems. These codes provide an effective solution for combating fading in wireless channels, leading thus to capacity increase without necessarily sacrificing bandwidth recourses. Due to their relative simplicity, STBC schemes are well suited to single-satellite diversity operations, such as polarization diversity from a single satellite by using orthogonal polarization components. However, in major part of research works related to STBC, it is assumed that the symbol timing at the receiver is perfectly known, an assumption that reflects the ideal case. In practice, timing error must be estimated and compensated for very accurately in order to avoid performance degradation of the wireless link. This problem is conventionally approached either by a feedback loop employing Timing Error Detector (TED) or by feedforward direct estimation of the timing error and its compensation by interpolation.

The optimum sample selection algorithm presented in [1] was the first work to address the problem of data-aided feedforward symbol timing estimation in multi-antenna systems. However, the algorithm requires extra transmission of training sequences, resulting thus in the reduction of the transmission rate. Furthermore, in order to achieve a reasonable performance, a large oversampling factor is needed. An improvement of this algorithm was presented in [2], where accurate estimates can be achieved even for small oversampling ratios (e.g. $Q = 4$) and was later enhanced further in [3] by utilizing the information about the transmitted pulse. Finally, in a similar paper [4], the squaring algorithm was used as another means of achieving timing recovery in space-time coding systems.

A special category of STBC, Orthogonal STBC (OSTBC), has proved to be an excellent coding technique for achieving the maximum diversity gain with simple linear decoding at the receiver. Research works that have addressed the issue of timing epoch tracking, for the specific cases of two, three and four transmit antenna orthogonal codes, for both complex and real modulation constellations at the baseband are [5]−[7]. The techniques presented there, are based on the use of maximum-likelihood detection (MLD) variables for estimating the timing error by examining the difference in threshold crossings.

Quasi-OSTBC (QOSTBC) constitutes a generalization of OSTBC with promising contribution to modern wireless applications. The significance of these codes stems from the fact that they have extended the family of OSTBC used in MIMO systems, achieving higher transmission rates when complex symbol transmission is concerned. The reduced orthogonality of their structure leads to increased decoding complexity, rendering the corresponding performance analysis a challenging task. To the best of our knowledge, just a single work [8], deals with the time tracking issue for the case of QOSTBC. However, in [8], a low complexity TED is presented under the consideration of a specific type of QOSTBC.

The organization of this paper is as follows. Section II presents the system model used. In Section III a simple decoupling scheme for QOSTBC derived in [9] is employed in order to extract the equivalent linear processed received signal in terms of the residual timing error, which occurs due to non-optimal sampling of the matched filter output. Simulation results and discussion regarding the influence of the timing offset on the performance of the system are presented in Section IV, examining various cases of space-time designs, number of receive antennas and values of the roll-off factor of the combined transmit-receive pulse shaping filter. Finally, Section V concludes the paper.

II. SYSTEM MODEL

We consider a communication system that employs STBC with N_T antennas at the transmitter and N_R antennas at the receiver. At time n the transmitter maps N_s information symbols $s_1(n), s_2(n), \cdots, s_{N_s}(n)$ into the constellation vector $\mathbf{x}_c(n) = [x_1(n)\ x_2(n) \cdots x_{N_s}(n)]$, whose elements are then encoded by the QOSTBC encoder as N_T code symbols in each time slot. The transmission of a single block of symbols is completed in N time slots, resulting thus in a code rate of $R = N_s/N$. The corresponding baseband signal model at the i_{th} transmitter branch can be written as

$$x_i(t) = \sum_n x_i(n)\,\delta(t - nT_s), \tag{1}$$

where δ is the Dirac delta and T_s the symbol period which is considered to be equal to the duration of each time slot. Data at the output of the QOSTBC encoder is passed through the transmit pulse shaping filter with impulse response $p_T(t)$, given by a square-root-raised-cosine (SRRC) pulse. The pulse-shaped signal $s_i(t) = x_i(t) * p_T(t)$ transmitted from the i_{th} antenna can be written as

$$s_i(t) = \sum_n x_i(n)\,p_T(t - nT_s). \tag{2}$$

During propagation, the signal undergoes corruption due to multipath fading and noise addition. The received signal at the j_{th} antenna can be modeled as

$$\begin{aligned} y_j(t) &= \sum_{i=1}^{N_T} h_{ij}(t)\,s_i(t-\tau) + \eta_j(t) \\ &= \sum_{i=1}^{N_T} h_{ij}(t) \sum_n x_i(n)\,p_T(t - nT_s - \tau) + \eta_j(t) \end{aligned} \tag{3}$$

where τ denotes the equal for all branches propagation time delay of the signal from the transmitter to the receiver, $h_{ij}(t)$ is the complex channel gain from the i_{th} transmit to the j_{th} receive antenna and $\eta_j(t)$ represents a zero mean complex Gaussian random variable with variance $N_o/2$ per real dimension. Since we assume a quasi-static frequency flat fading Rayleigh channel, the corresponding path gains are considered independent identically distributed samples of a Rayleigh random process remaining constant within a block, that is $h_{ij}(t) \approx h_{ij}$.

The matched filter output at the j_{th} receiver branch can be given by

$$\begin{aligned} r_j(t) &= y_j(t) * p_R(t) \\ &= \sum_{i=1}^{N_T} h_{ij} \sum_n x_i(n)\,p(t - nT_s - \tau) + n_j(t), \end{aligned} \tag{4}$$

where $p_R(t)$ is a SRRC pulse shape with frequency response $P_R(f) = P_T^*(f)$, $p(t)$ is the overall impulse response of the combined transmit-receive filter with spectrum $P(f) = P_T(f)\,P_R(f) = |P_T(f)|^2$ defining thus a Nyquist raised-cosine (RC) pulse shape with bandwidth $(1+\alpha)/2T_s$, α being the roll-off factor with $0 \le \alpha \le 1$ and $n_j(t) = \eta_j(t) * p_R(t)$ denotes the zero mean colored noise at the output of the matched filter.

As mentioned above, a synchronous digital communication system has to employ a symbol synchronizer at the receiver in order to achieve coherent detection. Thus, assuming that the output of the timing recovery loop gives an estimation $\hat{\tau}$ of the exact instant within the symbol interval T_s for effective matched filter output sampling against ISI, the timing error ϵ can be written as $\epsilon = \hat{\tau} - \tau$. We consider that ϵ remains constant within a block, being at the same time equal for all branches. Therefore, the $r_j(t)$ samples at time instants $t_m = mT_s + \hat{\tau}$, $m \in \mathbb{N}^*$ can be modeled as

$$r_j(m) = \sum_{i=1}^{N_T} h_{ij} \sum_{n=1}^{m} x_i(n)\,p_\epsilon(m-n) + n_j(m), \tag{5}$$

where it can easily be shown [1] that the noise samples $n_j(m)$ are uncorrelated since they have been taken at the symbol rate $1/T_s$. In the last equation we used the notations $r_j(t_m) \triangleq r_j(m)$, $n_j(t_m) \triangleq n_j(m)$ and $p(qT_s + \epsilon) = p_\epsilon(q)$ with $q \in \mathbb{N}^*$.

An alternative expression for (5) can be

$$r_j(m) = p_\epsilon(0) \sum_{i=1}^{N_T} h_{ij} x_i(m) + v_j(m) + n_j(m), \tag{6}$$

where it is obvious that the term $v_j(m)$ represents the intersymbol interference (ISI) from other transmitted symbols, given by

$$v_j(m) = \sum_{i=1}^{N_T} h_{ij} \sum_{n=1}^{m-1} x_i(n)\,p_\epsilon(m-n). \tag{7}$$

It can easily be drawn that if $\epsilon = 0$, then (6) takes the form

$$r_j(m) = \sum_{i=1}^{N_T} h_{ij} x_i(m) + n_j(m), \tag{8}$$

which is the well known STBC formula that gives the received signal at the j_{th} antenna in the ISI-free case. This stems from the fact that $p(t)$ function satisfies the *Nyquist condition for zero ISI*, that is $p(qT_s) = \delta_{q0}$, δ being the Kronecker delta.

III. Decoupling and Synchronization Issues

In this section we extract the equivalent system model for the received signal after linear processing, in the non-ideal case of interference caused from other transmitted symbols. In our analysis we assume that the symbol synchronizer achieves just a coarse timing phase recovery $\hat{\tau}$ within each block, leading thus to a residual timing offset ϵ at the receiver, equal to the difference $\hat{\tau} - \tau$. We will examine the influence of this quantity on the decoupling of orthogonal and quasi-orthogonal STBC under the assumption that the transmitter has no channel state information (CSI), whereas the receiver has perfect CSI.

A. QOSTBC Decoupler

Equation (5) can be written alternatively as

$$r_j(m) = \sum_{i=1}^{N_T} h_{ij} x_i^{\epsilon}(m) + n_j(m), \tag{9}$$

where $x_i^{\epsilon}(m) = \sum_{n=1}^{m} x_i(n) p_{\epsilon}(m-n)$. We can easily notice that each matched filter output sample given by (9), conveys information of all the code symbols transmitted so far, considering of course an appropriately large value for the roll-off factor α of the combined pulse shape $p(t)$. Hence, m can be expressed in terms of the current time slot k of the l_{th} block, that is $m = (l-1)N + k$, where $l \in \mathbb{N}^*$ and $k = 1, 2, \ldots, N$. In this way, by introducing the notations $r_j(m) \triangleq r_j^{(l)}(k)$, $n_j(m) \triangleq n_j^{(l)}(k)$ and $x_i^{\epsilon}(m) \triangleq x_{i,\epsilon}^{(l)}(k)$, (9) can be written in the following form

$$r_j^{(l)}(k) = \sum_{i=1}^{N_T} h_{ij} x_{i,\epsilon}^{(l)}(k) + n_j^{(l)}(k), \tag{10}$$

where $x_{i,\epsilon}^{(l)}(k) = \sum_{l'=1}^{l} \sum_{q=1}^{N} x_i^{(l')}(q) p_{\epsilon}((l-l')N + k - q)$ with $x_i^{(l')}(q)$ the transmitted code symbol at the q_{th} time slot of the l'_{th} block. For notational convenience, we consider the specific block $l = 1$ and we drop the block index superscript for the remainder of the derivation. Therefore, (10) takes the simpler form

$$r_j(k) = \sum_{i=1}^{N_T} h_{ij} x_i^{\epsilon}(k) + n_j(k), \tag{11}$$

with $x_i^{\epsilon}(k) = \sum_{q=1}^{N} x_i(q) p_{\epsilon}(k-q)$. By adopting matrix notation, the subsystem concerning the received signals at the j_{th} antenna can be described by the $N \times 1$ column vector

$$\mathbf{r}_j = \mathbf{X}_{\epsilon}\mathbf{h}_j + \mathbf{n}_j, \tag{12}$$

with $\mathbf{X}_{\epsilon} = \mathbf{P}_{\epsilon}\mathbf{X}$, where $\mathbf{X} \in \mathbb{C}^{N \times N_T}$ is the transmitted code matrix and $\mathbf{P}_{\epsilon} \in \mathbb{R}^{N \times N}$ is a Toeplitz matrix with entries the shifted pulse shapes $p_{kq}^{\epsilon} = p_{\epsilon}(k-q)$, given the timing error ϵ.

Like in [9], we focus on STBC where at each time-slot all N_s symbols are transmitted simultaneously via $N_T = N_s$ different transmit antennas. Hence, all the rows of the code matrix $\mathbf{X}$ can be considered as generalized permutations of either $\mathbf{x}_c = [x_1\ x_2 \cdots\ x_{N_s}]$ or $\mathbf{x}_c^*$ entries, since many complex orthogonal or quasi-orthogonal STBC have one or more of their rows conjugated including both signed and unsigned entries of $\mathbf{x}_c$. Considering that these row indices define a set S, we can recall from [9] the definition of $\Lambda_S[\cdot]$ operator, whose action on a matrix simply conjugates only the rows included in the set S. Therefore, we can proceed with the following matrix manipulations

$$\begin{aligned} \Lambda_S[\mathbf{r}_j] &= \Lambda_S[\mathbf{X}_{\epsilon}\mathbf{h}_j + \mathbf{n}_j] \\ &= \mathbf{P}_{\epsilon}\Lambda_S[\mathbf{X}\mathbf{h}_j] + \Lambda_S[\mathbf{n}_j] \\ &= \mathbf{P}_{\epsilon}\dot{\mathbf{H}}_j\mathbf{x}_c^T + \Lambda_S[\mathbf{n}_j] \\ &= \dot{\mathbf{H}}_{j,\epsilon}\mathbf{x}_c^T + \Lambda_S[\mathbf{n}_j], \end{aligned} \tag{13}$$

where $\dot{\mathbf{H}}_{j,\epsilon} = \mathbf{P}_{\epsilon}\Lambda_S\left[diag\left(\mathbf{h}_j^T\right)_N\right]\ddot{\mathbf{G}}^T$ defines the $N \times N_T$ equivalent channel matrix concerning the j_{th} receiving antenna in the presence of ISI, $diag(\cdot)_N$ denotes a $N \times N$ block diagonal matrix and $\ddot{\mathbf{G}}$ is the $N_T \times NN_T$ appended matrix of all the GPM related to the construction pattern of the code matrix $\mathbf{X}$, as described in [9].

B. Effect of the Timing Error on the Decoupling Procedure

In order to obtain a simple implemented ML decoder, we should first decouple the transmitted symbols appearing on the right side of (13). However, we cannot proceed by calculating the conjugate transpose of $\dot{\mathbf{H}}_{j,\epsilon}$, that is $\dot{\mathbf{H}}_{j,\epsilon}^H = \ddot{\mathbf{G}}\Lambda_S\left[diag\left(\mathbf{h}_j^*\right)_N\right]\mathbf{P}_{\epsilon}^T$, since we are not aware of the timing error quantity ϵ that is being involved in the construction of $\mathbf{P}_{\epsilon}$. For this reason, we will proceed by considering that ϵ takes uniformly distributed values in the interval $[-0.5,\ 0.5]$. Hence, multiplication on both sides of (13) by the term $\dot{\mathbf{H}}_{j,\epsilon}^H$, leads to the following equivalent linear decoupling scheme

$$\bar{\mathbf{r}}_j = \bar{\mathbf{G}}_{j,\epsilon}\mathbf{x}_c^T + \bar{\mathbf{n}}_j, \tag{14}$$

with

$$\begin{aligned} \bar{\mathbf{r}}_j &= \ddot{\mathbf{G}}\Lambda_S\left[diag\left(\mathbf{h}_j^*\right)_N\right]\mathbf{P}_{\epsilon}^T\Lambda_S(\mathbf{r}_j) \\ \bar{\mathbf{G}}_{j,\epsilon} &= \ddot{\mathbf{G}}\Lambda_S\left[diag\left(\mathbf{h}_j^*\right)_N\right]\mathbf{P}_{\epsilon}^T\mathbf{P}_{\epsilon}\Lambda_S\left[diag\left(\mathbf{h}_j^T\right)_N\right]\ddot{\mathbf{G}}^T, \end{aligned}$$

where $\bar{\mathbf{r}}_j$ is the $N_T \times 1$ linear processed received signal vector at the j_{th} antenna, $\bar{\mathbf{G}}_{j,\epsilon}$ defines the corresponding $N_T \times N_T$ diversity gain matrix (DGM) and $\bar{\mathbf{n}}_j$ represents the $N_T \times 1$ equivalent noise vector.

To proceed further, we have to take into consideration the rest of receiving branches as well, that is $\bar{\mathbf{R}} = \sum_{j=1}^{N_R}\bar{\mathbf{r}}_j = \bar{\mathbf{G}}\mathbf{x}_c^T + \bar{\mathbf{N}}$. Thus, the closed form expression describing the whole system model after predecoder becomes

$$\bar{\mathbf{R}} = \ddot{\mathbf{G}}\sum_{j=1}^{N_R}\Lambda_S\left[diag\left(\mathbf{h}_j^*\right)_N\right]\mathbf{P}_{\epsilon}^T\Lambda_S[\mathbf{r}_j], \tag{15}$$

with the diversity gain matrix given by

$$\bar{\mathbf{G}}_{\epsilon} = \ddot{\mathbf{G}}\sum_{j=1}^{N_R}\left(\Lambda_S\left[diag\left(\mathbf{h}_j^*\right)_N\right]\acute{\mathbf{P}}_{\epsilon}\Lambda_S\left[diag\left(\mathbf{h}_j^T\right)_N\right]\right)\ddot{\mathbf{G}}^T, \tag{16}$$

where $\acute{\mathbf{P}}_{\epsilon} = \mathbf{P}_{\epsilon}^T\mathbf{P}_{\epsilon}$ is a symmetric $N \times N$ matrix with entries $\acute{p}_{kq}^{\epsilon} = \sum_{c=1}^{N} p_{\epsilon}(c-k)p_{\epsilon}(c-q)$.

From (16), one can expect that $\acute{\mathbf{P}}_{\epsilon}$ will spoil the sparse nature of $\bar{\mathbf{G}}_{\epsilon}$, by generating extra non-zero entries off the main diagonal of the latter. Furthermore, since for non-trivial values of the timing error ϵ it holds that $\acute{p}_{kq}^{\epsilon} < 1$, it would be reasonable to expect a slight reduction in the entries of the equivalent channel matrix $\dot{\mathbf{H}}_{\epsilon} = \sum_{j=1}^{N_R} \dot{\mathbf{H}}_{j,\epsilon} = \mathbf{P}_{\epsilon}\Lambda_S \left[diag\left(\sum_{j=1}^{N_R} \mathbf{h}_j^T\right)_N\right] \ddot{\mathbf{G}}^T$. This reflects to a relative decrease in the average squared Frobenius norm of $\dot{\mathbf{H}}_{\epsilon}$ or equivalently in the average trace of $\bar{\mathbf{G}}_{\epsilon}$. At this point, we have to mention that the diagonal elements of $\bar{\mathbf{G}}_{\epsilon}$ define the maximum gain that can be achieved (equal to $\sum_{i=1}^{N_T}\sum_{j=1}^{N_R} |h_{ij}|^2$ for the ISI–free case and employment of orthogonal codes), being thus a measure of the system performance.

IV. Simulation Results and Discussion

In this section, we calculate the average trace of $\bar{\mathbf{G}}_{\epsilon}$ using Monte Carlo simulations. The residual timing offset has been generated to be uniformly distributed in the interval $[-0.5,\ 0.5]$. The channel gains h_{ij} have been generated as zero mean complex Gaussian random variables with variance $\sigma_N^2 = \Omega/2$ per real dimension, Ω being the average fading power. The combined transmit-receive pulse shape $p(t)$ has been considered to be a Nyquist raised-cosine given by

$$p(t) = \sin c\,(t/T_s)\,\frac{\cos(\alpha\pi t/T_s)}{1-(2\alpha t/T_s)^2}.$$

The simulations have been performed for the three STBC found in [10], [11] and [12] correspondingly, by the following structures

$$\mathbf{X}_{AL} = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix},\ \mathbf{X}_{ABBA} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ x_3 & x_4 & x_1 & x_2 \\ -x_4^* & x_3^* & -x_2^* & x_1^* \end{bmatrix},$$

$$\mathbf{X}_{J} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix}.$$

Figs. 1–3 show the trace of $\bar{\mathbf{G}}_{\epsilon}$ versus the average fading power Ω (dB) for the three different codes presented above. The simulations have been implemented by averaging $tr\left(\bar{\mathbf{G}}_{\epsilon}\right)$ over $10^5/N_T$ estimates (number of blocks for 10^5 transmitted symbols since we examine codes where $N_s = N_T$) and for two different values of the roll-off factor ($\alpha = 0.35$ and 0.9). The displayed results concern the scenaria of one and three antennas at the receiver.

Clearly, the contribution of the timing error is negative on the performance of the system. For all the cases, in the presence of ISI ($\epsilon \neq 0$) the line representing the average trace of the diversity gain matrix for increasing values of the average fading power, shifts down to indicate a decrease in the performance with respect to the ideal case of zero ISI ($\epsilon = 0$). Assuming that the amount of this shift is equal to d, the expectation of $\acute{\mathbf{P}}_{\epsilon}$ could be generally expressed as a one-to-one function $f : \mathbb{R}^{N \times N} \longrightarrow \mathbb{R}^{N \times N}$ applied on a $N \times N$ matrix D whose entries depend on d, that is

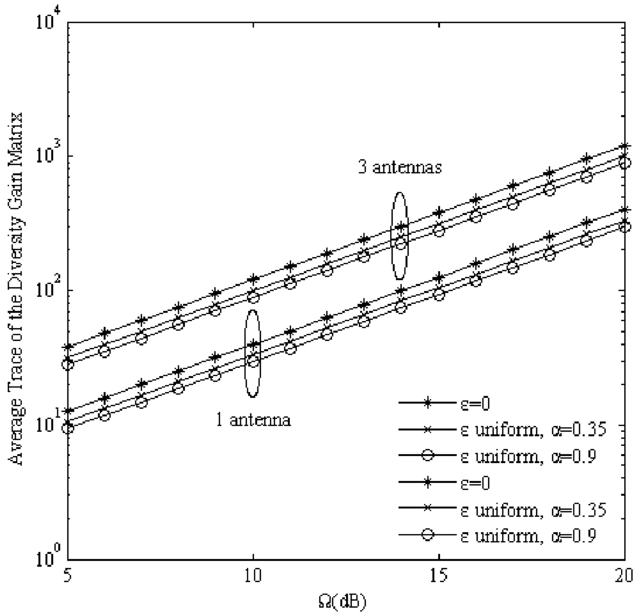

Fig. 1. Average trace of DGM for the Alamouti OSTBC.

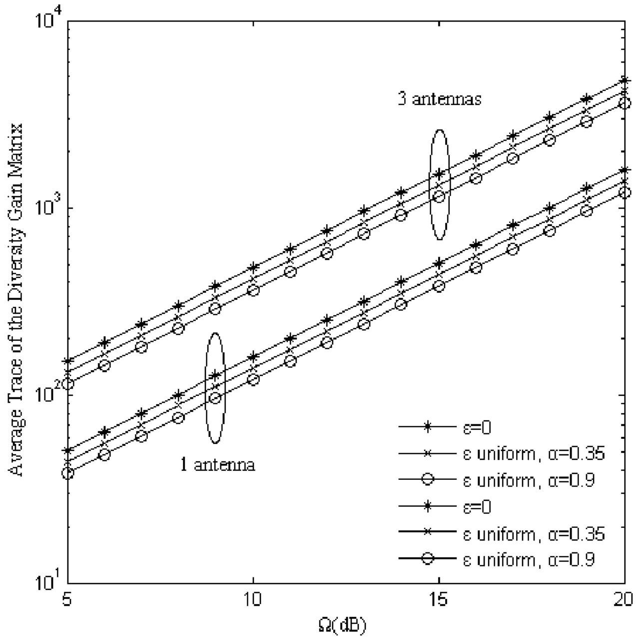

Fig. 2. Average trace of DGM for the Tirkkonen QOSTBC.

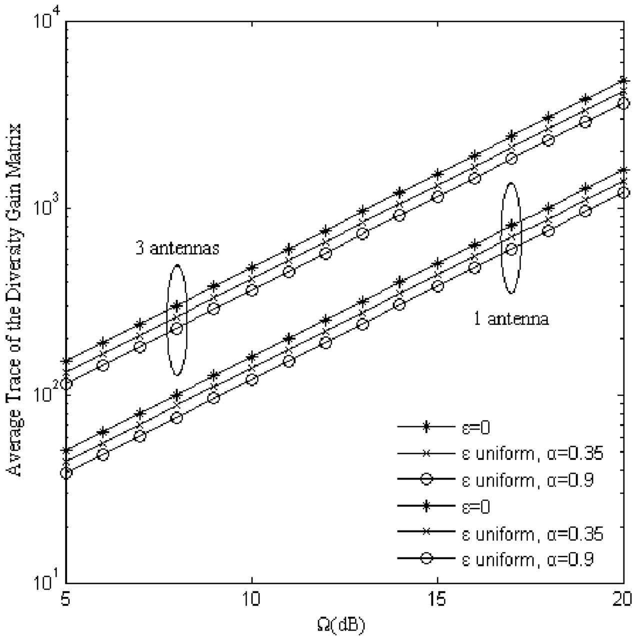

Fig. 3. Average trace of DGM for the Jafarkhani QOSTBC.

$$E\left[\acute{\mathbf{P}}_{\epsilon}\right] = f\left(D\right) \tag{17}$$

or in element-wise form $E\left[\acute{p}^{\epsilon}_{kq}\right] = g_{kq}\left(d\right)$, where $g_{kq}|_{k,q=1,2,\ldots,N}$ denotes a family of particular one-dimensional real functions. Finally, from the same figures we can observe that the greater the roll-off factor, the greater the distance d becomes.

V. Conclusion

An equivalent linear scheme for a communication system employing a general quasi-orthogonal design has been presented, considering that there is a remaining time offset ϵ between the clocks of the transmitter and the receiver. We verified that the non-trivial value of ϵ destroys the block diagonality of the DGM, prevented therefore us from extracting a closed form expression for the decoupling of the transmitted symbols. This agrees with our intuition, since the specific time drift causes interference from other transmitted symbols, spoiling thus the inherent distribution of the symbols inside the concerned block. This finding, along with the statistical behavior of ϵ, as deduced from (17), has stimulated us to study the design of a TED based on a decision-directed algorithm, planning to include it in our forthcoming research work.

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