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Joint equalization-despreading method for OQAM-CDMA systems

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ABSTRACT

Due to using no guard interval, Orthogonal frequency division multiplexing/offset quadrate amplitude modulation (OFDM/OQAM) systems are more bandwidth efficient than cyclic pre-OFDM (CP-OFDM). In OFDM/OQAM the real-valued symbols are transmitted and at the receiver a real-taking operation is performed to eliminate the pure imaginary intrinsic interference. Another way to eliminate the intrinsic interference of the OFDM/OQAM system is applying code-division multiple access (CDMA) to spread the transmitted symbols across the frequency axis. The resulted system, which is called Complex OQAM-CDMA, has the capability of transmitting complex-valued symbols (instead of real-valued ones) and can provide a transmission rate equal to that of the primer OFDM/OQAM system. At the receiver of the conventional Complex OQAM-CDAM, an equalizer is applied to remove the channel effect and then a despreading process is performed to detect the coded symbols. This detection method suffers from a poor performance in highly dispersive channels because of using two separate stages in the detection process. In this article based on the signal to interference plus noise ratio (SINR) maximization, we propose a novel detection method for the Complex OQAM-CDMA symbols. In the developed method, the equalization and despreading processes are performed jointly to maximize SINR. Simulation results assert that the proposed detection method outperforms the conventional scheme.

1. Introduction

Multicarier schemes are of the well-known systems to overcome the communication channel dispersion in time and frequency domains. Orthogonal frequency division multiplexing (OFDM) is the prevalent multicarrier system which uses a cyclic prefix (CP) as a guard interval between two transmitted symbols [1]. using rectangular pulse shapes to transmit symbols leads to a high out-of-band radiation and makes the CP-OFDM system very sensitive to carrier frequency offsets [2] and [3]. As a result, in case of time variant channels (e.g. underwater acoustic channels) or asynchronous user schemes (e.g. uplink and cognitive radio), CP-OFDM suffers from a high level of inter-subchannel interference [4]. In contrary to CP-OFDM, filter bank multicarrier (FBMC) systems use a well-localized prototype pulse in both time and frequency domains in order to shape the symbols [4]. Since the out-of-band power of the prototype pulse is low, the sensitivity of the FBMC scheme to the channel's time and frequency dispersion is very low [4].

On the other hand, the standard CP length in CP-OFDM is equal to the channel time-dispersion; as a result, in a fading channel with a long impulse response the length of CP becomes considerable compared to the total symbol length. Consequently, the bandwidth efficiency of CP-OFDM is very low in such a channel. In contrast, in FBMC systems, because of the well-localized pulse shapes, the necessity of inserting a CP between symbols disappears [4]. The prototype filter designation is investigated in many reasearches [5–7]. The isotropic orthogonal transform algorithm (IOTA) [5] and physical dynamic access (PHYDYAS) [6] are those widespread utilized prototype pulse shapes in FBMC systems.

It is remarkable that, because of the transient intervals of the prototype filter, FBMC suffers from a long latency; as a result, it is not a suitable system for short transmission windows and burst communication. In contrary to this drawback, because of insensitivity to time-frequency dispersion and non-CP procedure, FBMC is a promising candidate for the next generation of wireless communication networks. [8–12]. Furthermore, FBMC is the recommended scheme in highly dispersive channels, such as underwater acoustic environments [13].

Among all FBMC systems, the OFDM/OQAM scheme is a well-known FBMC system which can achieve the maximum bandwidth efficiency [4]. However, OFDM/OQAM suffers from an intrinsic in-

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terference, coming from the time-frequency overlaps of the transmitted pulse shapes. To eliminate the intrinsic interference in OFDM/OQAM, an innovative tactic is used. In this tactic the real-valued symbols are transmitted and at the receiver, a real-taking operation is performed to remove the pure imaginary intrinsic interference from the real-valued desired symbols [4,14,15].

Although OFDM/OQAM can achieve the maximal bandwidth efficiency, its direct application in multi-input multi-output (MIMO) channels is not performed effectively [4]. Especially, due to tracking the real-valued symbols at the receiver and loosing the complex orthogonality, the spatial multiplexing (SM) cannot be directly applied with OFDM/OQAM [16]. An alternative method to remove the inherent interference of the OFDM/OQAM system is to spread the symbols through the frequency axis with a well-chosen code division multiple access (CDMA) code [17]. The system proposed in [17], called Complex OQAM-CDMA, has the capability to transmit the complex-valued symbols, instead of real-valued ones. Since, this method can obtain complex orthogonality, in [18] the authors have shown that SM can be applied to the MIMO OFDM/OQAM system by using CDMA technique. Also, CDMA can achieve frequency multiplexing, which leads to increase the performance of this system. In [17] it is shown that when a specific set of Walsh–Hadamard code is used, the intrinsic interference can be removed. At the receiver of the Complex OQAM-CDMA, the despreading process is performed after removing the channel effect by using a simple equalizer.

Although the separated equalization-despreading detection procedure of Complex OQAM-CDMA leads to elimination of the interference, but it is a sub-optimal method, which is completely based on the flatness of channel spectrum over overlapping adjacent subchannels. Consequentially, in highly frequency selective channels, the performance of the conventional detection method of [17] decreases dramatically. On the other hand, in the next generation of communication systems, because of high data rate demands, the multipath effect is significant and; as a result, the channel frequency selectivity becomes very high [9]. Also in some channels, such as underwater acoustic channels, even in low data rates, the frequency selectivity of channel is too high [13]. One way to compensate the channel frequency selectivity in Complex OQAM-CDMA is to increase the number of filter bank subchannels. By increasing the number of subchannels, the bandwidth of each subcarrier reduces and the channel can be approximated by a one tap flat gain. However, this procedure leads to increase the time duration of prototype filters and as said in [4], "any timing offset at the receiver (or time spreading, arising from multipath effects in channel) results in a large number of ISI terms, which may add up and lead to a significant ISI power.". Also long prototype filter increases the latency and complexity of the transmitter and receiver [4]. Furthermore, in time varying channels, such as acoustic underwater channels, the shape of a long prototype filter will be damaged through the channel, which destroys the orthognality of the transmitted pulses and leads to a high level of interference [4]. Therefore, increasing the number of subchannels in Complex OQAM-CDMA is not a sufficient way to overcome the issue of non-flatness over each subchannel. In other words, the number of suchannels is limited by the frequency spreading of channel and must be determined such that at each time instant, the channel can be assumed time invariant.

In this regard, in this article we propose a novel detection method for the Complex OQAM-CDMA symbols based on the signal to interference plus noise ratio (SINR) maximization. In many reasearches, the maximization of SINR is addressed to achieve a good performance [19–22]. Consequently, the aim of the proposed method here is maximizing the SINR which leads to a convex optimization problem. Presenting the closed form solution for the optimization problem terminates to a joint equalization-despreading scheme. According to the simulation results, the proposed scheme outperforms the conventional detection method, especially in highly frequency selective channels. It is noteworthy that the supremacy of the proposed detection method is achievable in both lower and higher noise levels that makes the proposed system more reliable in practical situations.

The remaining of this paper is organized as follows. The primer OFDM/OQAM model is introduced in the next section. Section 3 presents the Complex OQAM-CDMA system and its conventional detection method. The proposed detection scheme for the Complex OQAM-CDMA is developed in Section 4. In this section we also compare the complexity of the proposed and conventional methods. Simulation results are given in Section 5, and finally, Section 6 contains the conclusions.

2. OFDM/OQAM system model

Let's assume that $a_{m,n}$ is the real-valued symbol transmitted at the mth subchannel and nth time slot. Accordingly, the continuous-time transmitted signal ($s(t)$) in a OFDM/OQAM system with M subchannels is

$$s(t) = \sum_{m=0}^{M-1} \sum_{n} a_{m,n} g_{m,n}(t), \tag{1}$$

where $g_{m,n}(t) \triangleq g(t - nT_0)e^{j2\pi mF_0 t}\varphi_{m,n}$ is the frequency-time shifted version of the prototype pulse shape $g(t)$ multiplied by the staggered factor $\varphi_{m,n} \triangleq (-1)^{mn} j^{m+n}$. Note that in (1) the frequency and time shifting steps of $g(t)$ are F_0 and T_0, respectively. In OFDM/OQAM system, $F_0T_0 = 1/2$. Therefore, the symbol density in frequency-time lattice is $\beta \triangleq 1/F_0T_0 = 2$ for real-valued symbols, which is equal to $\beta = 1$ for complex-valued symbols [4].

The received signal, after passing through the channel with impulse response $h(t)$ and additive noise $v(t)$, can be written as

$$y(t) = \int_{-\infty}^{+\infty} h(\tau)s(t-\tau)d\tau + v(t). \tag{2}$$

Considering (1) into (2), it yields to

$$y(t) = \sum_{m=0}^{M-1} \sum_{n} a_{m,n} g^{(h)}_{m,n}(t) + v(t), \tag{3}$$

in which, $g^{(h)}_{m,n}(t)$ is the convoluted $g_{m,n}(t)$ with $h(t)$ as

$$g^{(h)}_{m,n}(t) = \int_{-\infty}^{+\infty} h(\tau) g_{m,n}(t-\tau)d\tau. \tag{4}$$

At the receiver of the OFDM/OQAM system, in order to extract the transmitted symbol at the m_0th subchannel and n_0th time slot, the matched filter $g_{m_0,n_0}(t)$ is applied to $y(t)$, so that

$$y_{m_0,n_0} = \langle y(t), g_{m_0,n_0}(t)\rangle. \tag{5}$$

In addition, in OFDM/OQAM the prototype filter $g(t)$ is designed such that the real orthogonality condition is satisfied as

$$\Re\{\langle g_{m,n}(t), g_{m_0,n_0}(t)\rangle\} = \delta_{m,m_0}\delta_{n,n_0}. \tag{6}$$

In this regard, when the channel is distortion-free and without noise ($y(t) = s(t)$), the output of the matched filter would be

$$y_{m_0,n_0} = \langle s(t), g_{m_0,n_0}(t)\rangle = a_{m_0,n_0} + j\zeta_{m_0,n_0}, \tag{7}$$

where $j\zeta_{m_0,n_0}$ is the pure imaginary interference at the m_0th subchannel and n_0th time slot coming from the intrinsic overlaps of the OFDM/OQAM transmission strategy

$$j\zeta_{m_0,n_0} = \sum_{m=-\Delta_m}^{\Delta_m} \sum_{n=-\Delta_n}^{\Delta_n} a_{m_0+m,n_0+n}\langle g_{m_0+m,n_0+n}(t), g_{m_0,n_0}(t)\rangle. \tag{8}$$

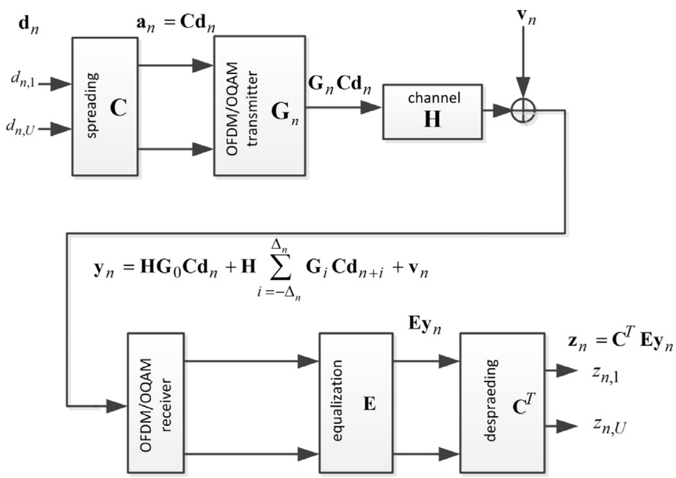

Fig. 1. Complex OQAM-CDMA transmission scheme presented in [17].

In (8), Δ_m and Δ_n are, respectively, the frequency and time interference ranges such that for either $|m| > \Delta_m$ or $|n| > \Delta_n$, the interference of the symbol a_{m_0+m,n_0+n} on the desired symbol a_{m_0,n_0}, is approximately trivial (i.e. $\langle g_{m_0+m,n_0+n}(t), g_{m_0,n_0}(t)\rangle = 0$). According to (7), the symbol at the m_0th subchannel and n_0th time slot, can be detected by taking the real part of y_{m_0,n_0} as

$$\hat{a}_{m_0,n_0} = \Re\{y_{m_0,n_0}\}. \tag{9}$$

On the other hand, if the channel is frequency selective, but its frequency spectrum can be approximately assumed, flat over the frequency-time adjacent overlapped pulses, the output of the matched filter can be written as [17]

$$y_{m_0,n_0} \approx H_{m_0}(a_{m_0,n_0} + j\zeta_{m_0,n_0}) + \nu_{m_0,n_0}, \tag{10}$$

where $H_{m_0} \triangleq \int_{-\infty}^{\infty} h(t)\exp(-j2\pi m_0 t/M)\,dt$ is the m_0th component of channel impulse response Fourier transformation and $\nu_{m_0,n_0} \triangleq \langle \nu(t), g_{m_0,n_0}(t)\rangle$ is the matched filter output noise. According to (10), before taking the real part of y_{m_0,n_0}, the effect of H_{m_0} (which is generally a complex quantity) must be eliminated. One way to get rid of the factor H_{m_0} is multiplying the y_{m_0,n_0} by $1/H_{m_0}$. In this regard, the real part of the resulted signal is considered as an estimation of the desired symbol contaminated by noise

$$\begin{aligned}\hat{a}_{m_0,n_0} &= \Re\{y_{m_0,n_0}/H_{m_0}\} = \Re\{(a_{m_0,n_0} + j\zeta_{m_0,n_0}) + \nu_{m_0,n_0}/H_{m_0}\}\\ &= a_{m_0,n_0} + \Re\{\nu_{m_0,n_0}/H_{m_0}\}.\end{aligned} \tag{11}$$

As it can be seen, in the OFDM/OQAM system, the detection process of a_{m_0,n_0} totally depends on the accuracy of the assumption of channel spectrum flatness over the interfered pulses (see (10)). Consequently, in highly frequency selective channels, where the spectrum of the channel cannot be assumed flat over adjacent subchannels, the validity of (10) is not asserted and the performance of the OFDM/OQAM system decreases, dramatically.

3. Complex OQAM-CDMA system

An alternative way to eliminate the intrinsic interference of the OFDM/OQAM system is applying a CDMA spreading code in frequency domain [17]. This technique called Complex OQAM-CDMA and has the capability of transmitting complex-valued symbols, instead of real-valued ones [17,18]. Fig. 1 illustrates the Complex OQAM-CDMA transmission scheme. Presenting the procedure of this system in a matrix formulation, suppose that $\mathbf{a}_n \triangleq [a_{0,n} \quad \cdots \quad a_{M-1,n}]^T$ is the vector of complex-valued transmitted symbols over all subchannels at the nth time slot and $\mathbf{y}_n \triangleq [y_{0,n} \quad \cdots \quad y_{M-1,n}]^T$ is the subordinate output vector at the receiver. Subsequently, if the channel is distortion-free and without noise, according to (7), it can be written that

$$\mathbf{y}_{n_0} = \mathbf{G}_0\mathbf{a}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{G}_n\mathbf{a}_{n_0+n}, \tag{12}$$

where $\mathbf{G}_n$ is a $M\times M$ matrix whose its (i, j)th entry is obtained as

$$g_n^{i,j} = \langle g_{i,n_0+n}(t), g_{j,n_0}(t)\rangle. \tag{13}$$

Note that in each row of $\mathbf{G}_n$ just $2\Delta_m + 1$ elements are non-zero (where $|i-j| \le \Delta_m$).

On the other hand, let's $\mathbf{c}_u \triangleq [c_{0,u} \ldots c_{M-1,u}]^T$ denotes the CDMA code, with length M, used by the uth user. The $\mathbf{c}_u$ code is chosen such that it satisfies the orthogonality condition $\mathbf{c}_u^T\mathbf{c}_{u_0} = \delta_{u,u_0}$. In this regard, for all users ($u = 1, \ldots, U$), the coding matrix, which is presented as $\mathbf{C} \triangleq [\mathbf{c}_1 \quad \cdots \quad \mathbf{c}_U]$, meets the orthogonality condition $\mathbf{C}^T\mathbf{C} = \mathbf{I}_U$ (where $\mathbf{I}_U$ is a $U\times U$ identity matrix). Thus, if the transmitted vector, $\mathbf{a}_n$, is the coded version of the complex-valued desired symbol vector $\mathbf{d}_n \triangleq [d_{n,1} \quad \cdots \quad d_{n,U}]^T$ at time slot n, it can be presented as

$$\mathbf{a}_n = \mathbf{C}\mathbf{d}_n. \tag{14}$$

Consequently, (12) can be rewritten as

$$\mathbf{y}_{n_0} = \mathbf{G}_0\mathbf{C}\mathbf{d}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{G}_n\mathbf{C}\mathbf{d}_{n_0+n}. \tag{15}$$

At the receiver, in order to despread the vector $\mathbf{y}_{n_0}$, it is multiplied by $\mathbf{C}^T$; thus the despread symbol vector becomes

$$\mathbf{z}_{n_0} = \mathbf{C}^T\mathbf{y}_{n_0} = \mathbf{C}^T\mathbf{G}_0\mathbf{C}\mathbf{d}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{C}^T\mathbf{G}_n\mathbf{C}\mathbf{d}_{n_0+n}, \tag{16}$$

where $\mathbf{z}_{n_0} \triangleq [z_{n_0,1} \quad \cdots \quad z_{n_0,U}]^T$ is the decoded symbol vector dedicated to all users. In [17], it is shown that when $U \le M/2$ and the CDMA code matrix $\mathbf{C}$ is a well-chosen Walsh–Hadamard code, it is derived that

$$\mathbf{C}^T\mathbf{G}_n\mathbf{C} = \begin{cases}\mathbf{I}_U, & \textit{when } n = 0\\ \mathbf{0}_U, & \textit{when } n \neq 0\end{cases}, \tag{17}$$

in which $\mathbf{0}_U$ is a zero matrix with size $U\times U$. Eventually, considering (17) into (16), leads to $\mathbf{z}_{n_0} = \mathbf{d}_{n_0}$ and the desired symbols are detected with no intrinsic interferences. In this article we will use the set of code that satisfies (17). The procedure of finding this code set is completely explained in [17]. Note that in our considerations, in order to have the maximum frequency diversity, we suppose that $U = M/2$. Therefore, the transmission rate is exactly equal to the primer OFDM/OQAM system ($\beta = 1$ for complex-valued symbols).

On the other hand, if the channel is frequency selective, (15) is replaced with

$$\mathbf{y}_{n_0} = \mathbf{G}_0^{(h)}\mathbf{C}\mathbf{d}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{G}_n^{(h)}\mathbf{C}\mathbf{d}_{n_0+n} + \mathbf{v}_{n_0}, \tag{18}$$

where $\mathbf{G}_n^{(h)}$ is an $M\times M$ matrix in which the (i, j)th element is

$$g_n^{(h)\,i,j} \triangleq \langle g_{i,n_0+n}^{(h)}(t), g_{j,n_0}(t)\rangle. \tag{19}$$

According to (4), the term $g_{m,n}^{(h)}(t)$ is the convolution of $g_{m,\,n}(t)$ and the channel impulse response $h(t)$. Also $\mathbf{v}_{n_0} \triangleq [\nu_{0,n_0} \ldots \nu_{M-1,n_0}]^T$ is the noise vector. By considering Fig. 1, at the receiver of the Complex OQAM-CDMA system, firstly, in order to eliminate the channel effect, a simple one-tap equalization is performed by multiplying $\mathbf{E}$ (the diagonal equalizer matrix) by the vector $\mathbf{y}_{n_0}$. Minimum mean square error (MMSE) or zero forcing (ZF) equalizer can be employed as a classical method. At the next step, the despreading process is applied to decode the desired symbols. As a result, the equalized-despreaded vector becomes

$$\mathbf{z}_{n_0} = \mathbf{C}^T\mathbf{E}\mathbf{y}_{n_0} = \mathbf{C}^T\mathbf{E}\mathbf{G}_0^{(h)}\mathbf{C}\mathbf{d}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{C}^T\mathbf{E}\mathbf{G}_n^{(h)}\mathbf{C}\mathbf{d}_{n_0+n} + \mathbf{C}^T\mathbf{E}\mathbf{v}_{n_0}. \tag{20}$$

Supposing that the equalizer completely removes the channel effect, the estimated symbols can be presented as $\mathbf{z}_{n_0} = \mathbf{d}_{n_0} + \mathbf{C}^T\mathbf{E}\mathbf{v}_{n_0}$. As it can be seen, by using of Walsh–Hadamard CDMA spreading scheme through the frequency axis, the intrinsic interference of the Complex OQAM-CDMA system can be perfectly removed; as a result, the desired symbols are detected with no tainting interference from the other symbols. But, note that the detection procedure depends on the removing of the interference. Thus, in case of highly frequency selective channels, for which the interference can not be removed completely, the performance of the conventional detection method decreases. In contrary, in this article we propose a detection method for Complex OQAM-CDMA which is based on the SINR maximization. As it will be shown in the next section, the proposed equalization and despreading processes are performed jointly, which improves the performance of Complex OQAM-CDMA.

4. The proposed joint equalization-deaspriding detection for complex OQAM-CDMA system

In other to develop the proposed detection scheme of Complex OQAM-CDMA symbols based on the SINR maximization, at the receiver of the system, we assume that the detection vector $\mathbf{f}_u^{(h)}$ is used to detect the transmitted symbols of the uth user. The aim of the proposed detection scheme is to develop the matrix $\mathbf{F}^{(h)} \triangleq [\mathbf{f}_1^{(h)} \quad \dots \mathbf{f}_U^{(h)}]$, of the size $M \times U$, such that SINR of $\mathbf{f}_u^{(h)^H}\mathbf{y}_n$ is maximized for $u = 1, \dots, U$. In this regard, the detected symbol of the u_0th user at the time slot n_0 can be written as

$$\begin{aligned} z_{n_0,u_0} &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{y}_{n_0} = \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\mathbf{C}\mathbf{d}_{n_0} + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_n^{(h)}\mathbf{C}\mathbf{d}_{n_0+n} + \mathbf{f}_{u_0}^{(h)^H}\mathbf{v}_{n_0} \\ &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}d_{n_0,u_0} + \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\sum_{u=1, u\neq u_0}^{U}\mathbf{c}_u d_{n_0,u} \\ &\quad + \sum_{n=-\Delta_n, n\neq 0}^{\Delta_n} \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_n^{(h)}\mathbf{C}\mathbf{d}_{n_0+n} + \mathbf{f}_{u_0}^{(h)^H}\mathbf{v}_{n_0}. \end{aligned} \tag{21}$$

As is obvious, the resulted z_{n_0,u_0} contains four segments. The first segment consists of the desired symbol, the second segment contributes the interference of the other users' symbols at the n_0th time slot, the third segment presents the interference from all users at the other time slots and finally, the fourth segment indicates the noise effect. Assuming that the transmitted symbols are i.i.d and there is an additive white Gaussian noise with variance of σ_n^2, the power of the desired signal, interference and noise are, respectively

$$\begin{aligned} P_{u_0}^{(d)} &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\mathbf{f}_{u_0}^{(h)}; \\ P_{u_0}^{(I)} &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\left(\sum_{u=1, u\neq u_0}^{U}\mathbf{c}_u\mathbf{c}_u^T\right)\mathbf{G}_0^{(h)^H}\mathbf{f}_{u_0}^{(h)} \\ &\quad + \mathbf{f}_{u_0}^{(h)^H}\left(\sum_{n=-\Delta_n, n\neq 0}^{\Delta_n}\mathbf{G}_n^{(h)}\mathbf{C}\mathbf{C}^T\mathbf{G}_n^{(h)^H}\right)\mathbf{f}_{u_0}^{(h)} \\ &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{A}_{u_0}\mathbf{f}_{u_0}^{(h)}; \\ P_{u_0}^{(n)} &= \mathbf{f}_{u_0}^{(h)^H}\mathbf{B}\mathbf{f}_{u_0}^{(h)}, \end{aligned} \tag{22}$$

where $\mathbf{A}_{u_0}$ is defined as

$$\mathbf{A}_{u_0} = \mathbf{R} - \mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}, \tag{23}$$

and $\mathbf{R} = \sum_{n=-\Delta_n}^{\Delta_n}\mathbf{G}_n^{(h)}\mathbf{C}\mathbf{C}^T\mathbf{G}_n^{(h)^H}$. Also $\mathbf{B}$ is the covariance matrix of the noise vector, which is a Hermitian matrix with the size of $M \times M$. Since the mth entry of vector $\mathbf{v}_{n_0}$ is $v_{m,n_0} = \langle v(t), g_{m,n_0}(t)\rangle$, the entry of the ith row and jth column of matrix $\mathbf{B}$ becomes $\mathbf{B}_{i,j} = \sigma_n^2\langle g_{j,n_0}(t), g_{i,n_0}(t)\rangle$. According to (22), the SINR of the u_0th user can be calculated as

$$\text{SINR}_{u_0} = \frac{P_{u_0}^{(d)}}{P_{u_0}^{(I)} + P_{u_0}^{(n)}} = \frac{\mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\mathbf{f}_{u_0}^{(h)}}{\mathbf{f}_{u_0}^{(h)^H}(\mathbf{A}_{u_0} + \mathbf{B})\mathbf{f}_{u_0}^{(h)}}. \tag{24}$$

As it is mentioned before, the aim of designing the proposed detection method is to obtain the vector $\mathbf{f}_{u_0}^{(h)}$ such that SINR_{u_0} is maximized. The SINR maximization with respect to $\mathbf{f}_{u_0}^{(h)}$ can be transformed to a convex problem, for which the global optimum solution is achievable [23]. In this regard, we can maximize the desired signal power subjected to the fact that the power of interference plus noise is constant. Accordingly, the maximization problem can be presented as

$$\begin{aligned} &\max_{\mathbf{f}_{u_0}^{(h)}} \left\{\mathbf{f}_{u_0}^{(h)^H}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\mathbf{f}_{u_0}^{(h)}\right\} \\ &\text{S.T.} \quad \mathbf{f}_{u_0}^{(h)^H}(\mathbf{A}_{u_0} + \mathbf{B})\mathbf{f}_{u_0}^{(h)} = 1. \end{aligned} \tag{25}$$

The optimization problem of (25) is convex. Since $\mathbf{A}_{u_0} + \mathbf{B}$ is a Hermitian matrix, it can be decomposed as $\mathbf{A}_{u_0} + \mathbf{B} = \Gamma_{u_0}\Gamma_{u_0}^H$ [24]. By defining $\mathbf{b}_{u_0} \triangleq \Gamma_{u_0}^H\mathbf{f}_{u_0}^{(h)}$, the optimization problem of (25) can be rewritten as

$$\begin{aligned} &\max_{\mathbf{b}_{u_0}} \left\{\mathbf{b}_{u_0}^H\Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\Gamma_{u_0}^{-H}\mathbf{b}_{u_0}\right\} \\ &\text{S.T.} \quad \mathbf{b}_{u_0}^H\mathbf{b}_{u_0} = 1. \end{aligned} \tag{26}$$

The term $\mathbf{b}_{u_0}^H\Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\Gamma_{u_0}^{-H}\mathbf{b}_{u_0}$ is maximized when $\mathbf{b}_{u_0}$ is the eigenvector associated with the maximum eigenvalue of matrix $\Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\Gamma_{u_0}^{-H}$ [24]. Note that matrix $\Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H}\Gamma_{u_0}^{-H}$ has just one non-zero eigenvalue. Thus, the optimization of (27) is solved when $\mathbf{b}_{u_0}$ is equal to the eigenvector associated with this non-zero eigenvalue and is obtained as [23]

$$\mathbf{b}_{u_0} = \Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}. \tag{27}$$

Thus, the optimized vector $\mathbf{f}_{u_0}^{(h)}$ can be calculated as

$$\begin{aligned} \mathbf{f}_{u_0}^{(h)} &= \Gamma_{u_0}^{-H}\mathbf{b}_{u_0} = \Gamma_{u_0}^{-H}\Gamma_{u_0}^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0} = (\Gamma_{u_0}\Gamma_{u_0}^H)^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0} \\ &= (\mathbf{A}_{u_0} + \mathbf{B})^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0} = \left(\mathbf{R} - \mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H} + \mathbf{B}\right)^{-1}\mathbf{G}_0^{(h)}\mathbf{c}_{u_0}. \end{aligned} \tag{28}$$

As a result, following the proposed scheme for Complex OQAM-CDMA symbol detection, the desired symbol transmitted by the u_0th user at the time slot n_0, can be estimated as $z_{n_0,u_0} = \mathbf{f}_{u_0}^{(h)^H}\mathbf{y}_{n_0}$ (as shown in (21)), in which the detection vector $\mathbf{f}_{u_0}^{(h)}$ is obtained according to (28). Fig. 2 illustrates the Complex OQAM-CDMA with the proposed detection method, in which the equalization and despreading processes are performed jointly by using of matrix $\mathbf{F}^{(h)}$ for all users (note that in the conventional method each equalization and despreading steps are performed separately).

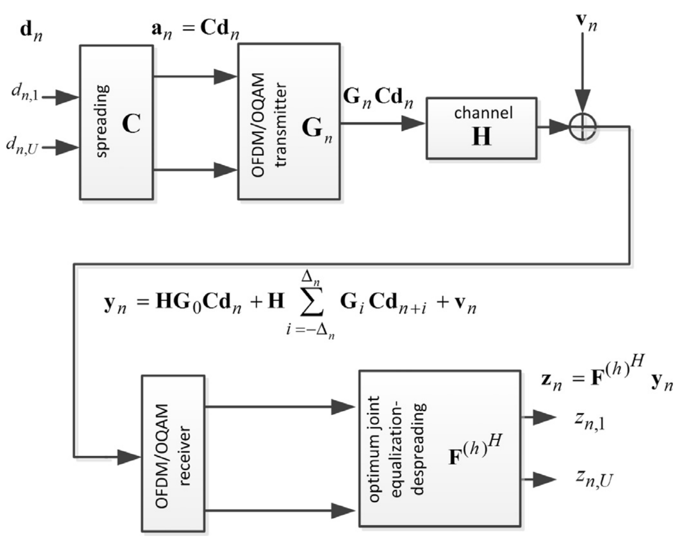

Fig. 2. Complex OQAM-CDMA with optimum detection method.

4.1. Complexity

Both proposed and conventional detection methods of Complex OQAM-CDMA include two steps: at the first step, equalization-despreading coefficients are computed (for the proposed method, these coefficients are computed in (28).); and at the next step, the derived coefficients are employed. Therefore, by considering both detection methods, there are two types of complexity. The first type of complexity is related to the computation of equalization-despreading coefficients which we call it "*Computational Complexity*"; and the second type of complexity is related to employing the computed coefficients, which is called the "*Implementation Complexity*".

In the conventional method, calculation of equalizer tap at each subchannel is performed separately. Thus, the whole *Computational Complexity* of this method is with the order $O(M)$. In contrary, in the proposed method the optimized detection vectors with length M is calculated by using (28). This equation includes some matrix multiplications and additions. Also we have a matrix inversion. The calculation complexity order of two matrix addition with the size of $M \times M$ is $O(M^2)$. The calculation complexity order of two matrix multiplication with size of $M \times M$ is $O(M^{2.37})$. Also the calculation complexity order of an arbitrary matrix inversion with size of $M \times M$ is $O(M^{2.37})$ [26], but in (28) $\mathbf{R} - \mathbf{G}_0^{(h)}\mathbf{c}_{u_0}\mathbf{c}_{u_0}^T\mathbf{G}_0^{(h)^H} + \mathbf{B}$ is a symmetric positive definite matrix and the complexity order of its matrix inversion is much lesser than $O(M^{2.37})$ [27]. Therefore, it can be concluded that the order of total *Computational Complexity* in (28) is $O(M^{2.37})$.

In case of "*Implementation Complexity*", the number of real-valued multiplications and real-valued additions needed to implement the detection methods is considered. In the conventional CDMA-OQAM proposed in [17], equalization and decoding are performed within two steps. According to this method, equalizing at each subchannel needs 4 real-valued multiplications and 3 real-valued additions. Also, for decoding at each subchannel, 2 real-valued multiplications and 1 real-valued addition are needed. Therefore, in the conventional method, equalization and decoding totally need $6M$ real-valued multiplications and $4M$ real-valued additions for all subchannels, at each time slot. In contrary, in the proposed method, since equalization and decoding are performed jointly at one step, just $4M$ real-valued multiplications and $3M$ real-valued additions are needed for all subchannels, at each time slot. Thus, in term of "*Implementation Complexity*', the proposed method has a lower complexity than the conventional method.

Another point needs attention about the detection methods' complexity is that the computation of equalization-deaspriding matrixes of both conventional and optimum methods is performed just once and the calculated coefficients remain valid until the channel impulse response does not change. Therefore, *Computational Complexity* of the proposed method becomes more tolerable when the channel is not fast time variant. On the other hand, since the conventional method is not optimized, it cannot remove the interference, perfectly which leads to significant degradation of the performance in the highly frequency selective channels. In contrary, one can say that although the proposed method has a larger complexity in the term of *Computational Complexity*, it can remove the interference perfectly, even in highly frequency channels. The simulation results show that we can achieve a very good performance in the highly frequency selective channels with the expense of more complexity by using of the proposed method. This performance never can be obtained when the conventional method is used. In other words, the conventional method loses its performance in the highly dispersive channels and cannot compensate it anyway, but the proposed method can achieve an acceptable performance with an additional computational load.

Table 1
Parameters of simulations.

Parameter	Value/type
Modulation	4-QAM
Prototype filter	IOTA
Carrier frequency	$f_c = 1\,\text{GHz}$
Frequency shift step	$F_0 = 15\,\text{KHz}$
Number of subchannels	$M = 256$
Sampling period	$t_s = 1/MF_0 = 260$ ns
Time shift step	$T_0 = Mt_s/2 = 33280\,\text{ns}$

5. Simulation results

In this section, in order to evaluate the performance of the proposed detection scheme, we compare the bit error rate (BER) of CP-OFDM-CDMA (a CP-OFDM scheme in which Walsh-Hadamard code is used), Complex OQAM-CDMA with the conventional detection method presented in [17] and Complex OQAM-CDMA with the proposed detection method. In the conventional complex OQAM-CDMA detection, a one-tap MMSE equalizer is used at each subchannel. The simulation parameters are presented in Table 1. Also for the CP-OFDM-CDMA scheme the standard length of CP is considered equal to 1/8 number of subcarriers (in here $(1/8) \times 256 = 32$) [25]. In all simulations we assume that the perfect channel state information is available at the receiver.

At first, we consider the Pedestrian-A channel where the power delay profile (PDP) is given as [28]

$$\text{Delays} = \begin{bmatrix}0 & 110 & 190 & 410\end{bmatrix} ns$$

$$\text{Powers} = \begin{bmatrix}0 & -9.7 & -19.2 & -22.8\end{bmatrix} dB$$

Fig. 3 illustrates the three systems' performances versus the Signal to Noise Ratio (SNR). As it can be seen, the performances of all three methods are approximately the same. This is because of that the delay spread of channel is too short. Therefore, the channel spectrum can be assumed flat at each subchannel. So the performance of the conventional method and the proposed method are the same. The same results can be seen in Fig. 4, where the Extended Pedestrian-A channel with the following PDP is considered as

$$\text{Delays} = \begin{bmatrix}0 & 30 & 70 & 90 & 110 & 190 & 410\end{bmatrix} ns$$

$$\text{Powers} = \begin{bmatrix}0 & -1 & -2 & -3 & -8 & -17.2 & -20.8\end{bmatrix} dB$$

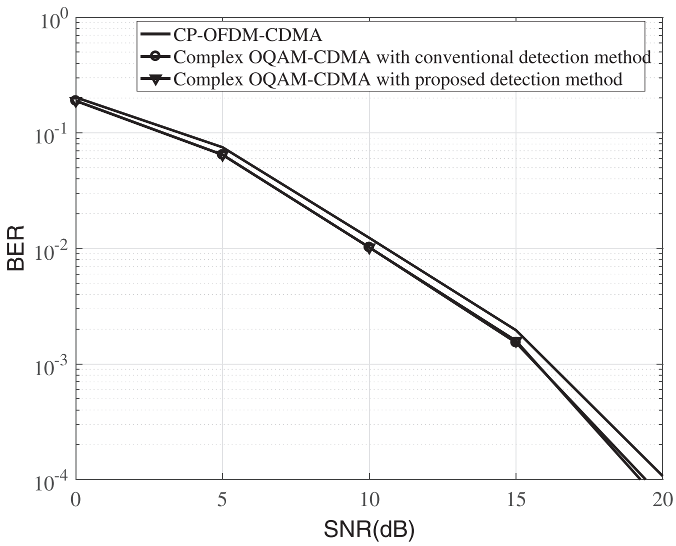

Fig. 3. Comparing the BER of CP-OFDM-CDMA with that of OQAM-CDMA with the conventional and the proposed detection methods over Pedestrian-A channel.

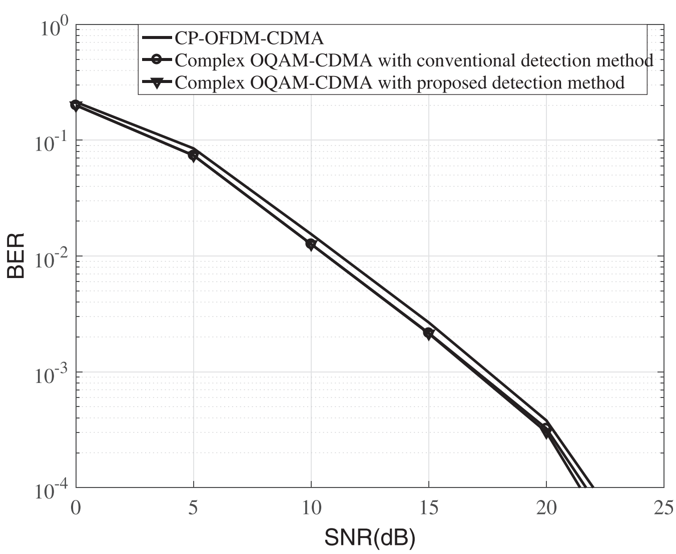

Fig. 4. Comparing the BER of CP-OFDM-CDMA with that of OQAM-CDMA with the conventional and the proposed detection methods over Extended Pedestrian-A channel.

In the following, we consider the Pedestrian-B channel, where the delay spread is very longer than that of the Pedestrian-A channel and its PDP is given as [28]

$$\text{Delays} = \begin{bmatrix} 0 & 200 & 800 & 1200 & 2300 & 3700 \end{bmatrix} ns$$
$$\text{Powers} = \begin{bmatrix} 0 & -0.9 & -4.9 & -8 & -7.8 & -23.9 \end{bmatrix} dB$$

Fig. 5 shows the BER of three methods versus SNR for the channel with the given PDP. We can see that the BER of the Complex OQAM-CDMA with the proposed detection method outperforms those of the conventional detection method and CP-OFDM-CDMA, over all the SNR range. These results assert that in highly frequency selective channels, the equalizer used in the conventional method does not perfectly remove the interference. On the other hand, since equalization and decoding in the proposed method is performed jointly, this method can remove the interference, very well.

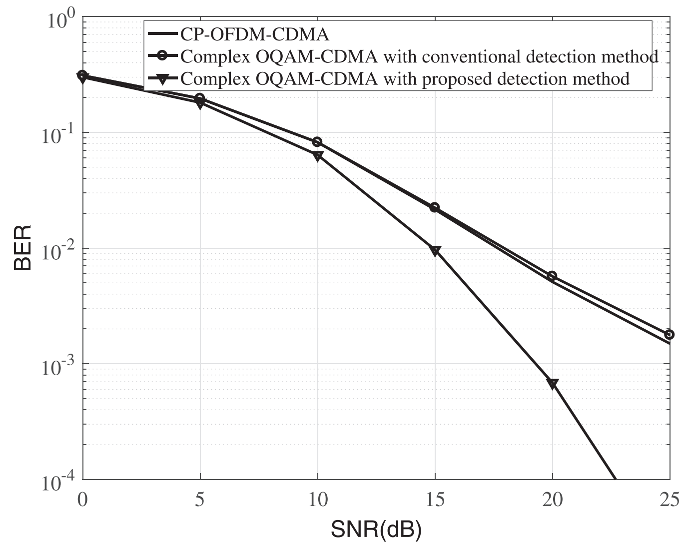

Fig. 5. Comparing the BER of CP-OFDM-CDMA with that of OQAM-CDMA with the conventional and the proposed detection methods over Pedestrian-B channel.

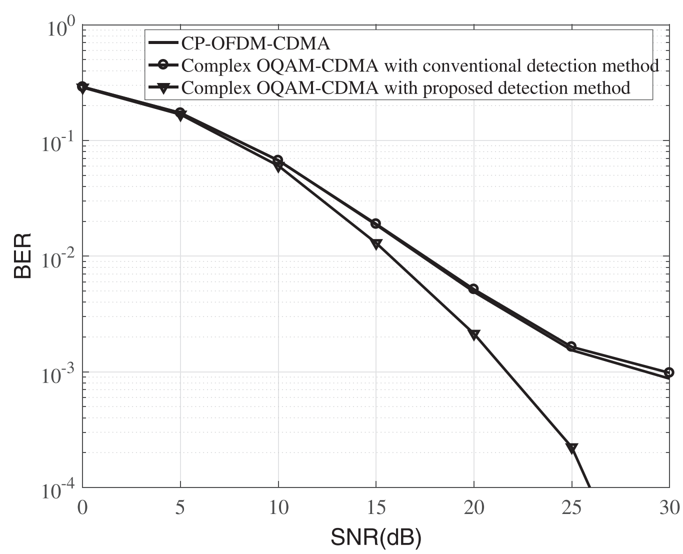

Fig. 6. Comparing the BER of CP-OFDM-CDMA with that of OQAM-CDMA with the conventional and the proposed detection methods over Vehicular-A channel.

Also Fig. 6 shows the BER versus SNR for the Vehicular-A channel model with the PDP given as [28]

$$\text{Delays} = \begin{bmatrix} 0 & 300 & 700 & 1100 & 1700 & 2500 \end{bmatrix} ns$$
$$\text{Powers} = \begin{bmatrix} 0 & -1 & -9 & -10 & -15 & -20 \end{bmatrix} dB$$

It can be seen that the performance of the proposed detection method of the Complex OQAM-CDMA symbols is considerably higher than that of the conventional detection method and also that of the CP-OFDM-CDMA system. The same results as Figs. 5 and 6 are obtained in Fig. 7, where the Extended Vehicular-A channel with the following PDP is considered as

$$\text{Delays} = \begin{bmatrix} 0 & 30 & 150 & 310 & 370 & 710 & 1090 & 1730 & 2510 \end{bmatrix} ns$$
$$\text{Powers} = \begin{bmatrix} 0 & -1.5 & -1.4 & -3.6 & -0.6 & -9.1 & -7 & -12 & -16.9 \end{bmatrix} dB$$

According to the achieved results in Figs. 3–7, one can conclude that for those channels with higher frequency selectivity, the performance of the Complex OQAM-CDMA scheme with the conventional detection method significantly decreases; in contrary, the performance of the Complex OQAM-CDMA system with the pro-

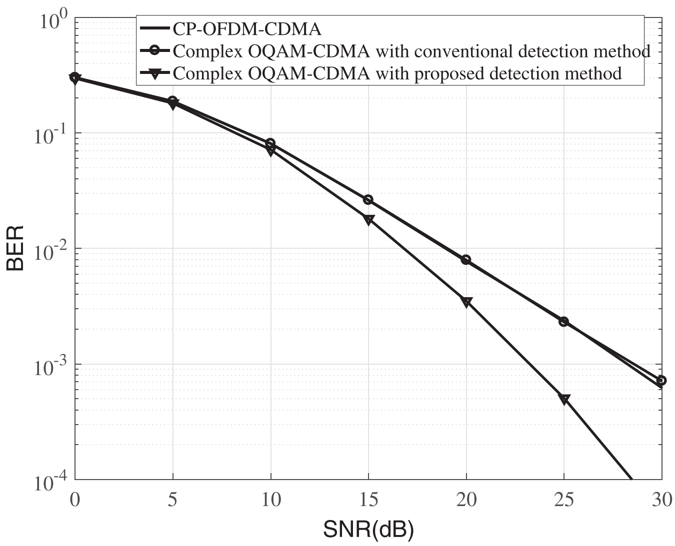

Fig. 7. Comparing the BER of CP-OFDM-CDMA with that of OQAM-CDMA with the conventional and the proposed detection methods over Extended Vehicular-A channel.

posed detector is less affected by the channel frequency selectivity. These results assert that the performance of the conventional detection method highly depends on the frequency selectivity of channel, such that the severe fluctuations in channel spectrums lead to considerable BER; on the other hand, the performance of the Complex OQAM-CDMA with the proposed detection method is much more robust against non-smooth channel spectrum and guarantees the lower BER even over high frequency selective channels.

In addition, it is noticeable that in the resulted figures the supremacy of the proposed detection method over conventional scheme is achieved through the all evaluated SNR range. In other words, in both lower and higher SNRs, the proposed detection method outperforms the conventional scheme, which makes it more reliable in practical situation with high noise levels.

6. Conclusions

In this article, based on the SINR maximization, we proposed a novel detection method for the Complex OQAM-CDMA system, which leads to a joint equalization-despreading scheme, instead of separate equalization and then despreading. The proposed detection method matrix is obtained by maximizing the desired signal power, subject to the constant interfered signal power plus the noise power. Considering no compromising assumption during finding the proposed detection method, the resulted scheme has a better BER performance in comparison with the conventional suboptimum method. According to the simulation results the proposed method outperforms the conventional scheme in frequency selective channel scenarios. This superiority is achievable in both lower and higher SNRs.

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