PILOT DESIGN FOR NON-CONTIGUOUS SPECTRUM USAGE IN OFDM-BASED COGNITIVE RADIO NETWORKS

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ABSTRACT

In this paper, challenges regarding the provision of channel state information (CSI) in non-contiguous orthogonal frequency division multiplexing (NC-OFDM) cognitive radio (CR) systems are addressed. We propose a novel scheme that utilizes convex optimization together with the cross entropy (CE) optimization to design pilot symbols that minimizes the channel estimate mean squared error (MSE) of frequency selective channels. Simulation results indicate that, the proposed pilot symbol design provides better channel estimate MSE as well as the bit error rate (BER) performance when compared with the equal powered pilot design.

***Index Terms*—** NC-OFDM, cognitive radio, cross entropy optimization, MSE, convex optimization.

1. INTRODUCTION

The field of wireless networks has witnessed tremendous growth in recent years and the scarcity in spectrum has become a serious problem since a large portion of the wireless spectrum has been allocated for specific applications. However, measurements have shown that most of the allocated spectrum is largely underutilized [1]. To alleviate this problem, many papers have proposed the concept of spectrum pooling as well as various transceiver designs for transmission across non-contiguous (NC) portions of the spectrum [1–3].

Cognitive radios (CRs) have been proposed as a technology for the opportunistic use of the underutilized spectrum. CR improves spectral efficiency by sensing the spectrum, detects the presence of the primary users (PUs) and exploits the unused spectrum without disturbing the PUs [1].

The non-contiguous orthogonal frequency division multiplexing (NC-OFDM) transceivers are prominent candidates for CR systems as they are designed to transmit information in the presence of PUs. By deactivating subcarriers utilized by the PUs interference between the PUs and the secondary users (SUs) can be mitigated [4, 5]. However, the presence of deactivated subcarriers in the active subcarrier zone may possibly lead to NC sequences of the available subcarriers for the SUs and thereby complicate the design of efficient pilot symbols for channel estimations [2–5].

Several pilot symbol designs for channel estimation have been predominantly developed for OFDM systems with and without null edge subcarriers (see [6–10], and the reference therein). Optimal pilot symbols for OFDM systems in the absence of null edge subcarriers are considered in [9, 10] where equal distant and equal powered pilot symbols were found to be optimal with respect to several performance measures. However, in [6–8], it has been demonstrated that, for OFDM systems with null edge subcarriers equal distant and equal powered pilot symbols are not necessarily optimal.

The pilot symbol designs in [6–8] are effective for OFDM systems where the spectrum is contiguous except for the DC and null edge subcarriers. For an arbitrary set of pilot subcarriers, the methods in [6–8] can be adopted for pilot power distribution. However, the schemes are not effective for pilot positions optimization in NC-OFDM systems. The cubic parameterization in [8] requires a contiguous sequence of subcarriers, while the iterative algorithm in [6] involves symmetrical remove of minimum power subcarriers. For the NC-OFDM systems, the activated subcarriers for the SUs may not be symmetrical. Non symmetrical remove is also possible but the performance is poor.

In [2, 5], pilot symbols design method for OFDM-based CR systems is proposed. The method in [2, 5] formulates the pilot design as an optimization problem that minimizes the upper bound related to the least square (LS) channel estimate mean squared error (MSE). An efficient scheme to solve the optimization problem is also proposed. The algorithm in [2,5] considers equal powered pilot symbols and obtain the optimal placement of pilot symbols for a random set of activated subcarriers. However, the optimality of equal powered pilot symbols does not necessarily hold true when there are null subcarriers [6].

In this paper, we propose a novel method that utilizes convex optimization together with the cross entropy (CE) optimization to design pilot symbols of NC-OFDM to minimize the MSE of the LS channel estimate of the frequency selective channel. We use convex optimization to find optimal power distribution to the activated subcarriers. Then, we formulate the optimal pilot placement problem as a combinatorial problem and employ the CE optimization, to select the placement of the pilot symbols. To the selected optimal pilot set, we again optimize pilot power distribution. Design examples consistent with IEEE 802.16e are provided to corroborate the superior performance of our proposed method over the equal powered pilot symbols in [2, 5].

2. SYSTEM MODEL

We consider point-to-point wireless OFDM transmissions over frequency selective fading channels. We assume that the discrete-time baseband equivalent channel has a filter impulse response (FIR) of maximum length L, and remains constant in at least one OFDM symbol, i.e., is quasi-static. The channel impulse response is denoted as $\{h_0, h_1, \ldots, h_{L-1}\}$.

Let us consider the transmission of one OFDM symbol with N number of subcarriers. At the transmitter, a symbol sequence $\{s_0, s_1, \ldots, s_{N-1}\}$ undergoes serial-to-parallel (S/P) conversion to be stacked into one OFDM symbol. Then, an N-points inverse discrete Fourier transform (IDFT) follows to produce N dimensional data, which is parallel-to-serial (P/S) converted. A cyclic prefix (CP) of length N_{cp} is appended to mitigate the multipath effects. The discrete-time baseband equivalent transmitted signals u_n can be expressed in the time-domain as

$$u_n = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k e^{j\frac{2\pi kn}{N}}, \quad n \in [0, N-1]. \tag{1}$$

Assume that N_{cp} is greater than the channel length L so that there is no inter-symbol interference (ISI) between consecutive OFDM symbols. At the receiver, we assume perfect timing and frequency synchronization. After removing CP, we apply discrete Fourier transform (DFT) to the received time-domain signal y_n for $n \in [0, N-1]$ to obtain for $k \in [0, N-1]$

$$Y_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y_n e^{-j\frac{2\pi kn}{N}} = H_k s_k + W_k, \tag{2}$$

where H_k is the channel frequency response at frequency $2\pi k/N$ given by

$$H_k = \sum_{l=0}^{L-1} h_l e^{-j\frac{2\pi kl}{N}}, \tag{3}$$

and the noise W_k is assumed to be i.i.d. circular Gaussian with zero mean and variance σ_w^2.

For OFDM-based cognitive radio (CR) systems, subcarriers occupied by the primary users (PUs) are deactivated, thus, the cognitive secondary user (SU) utilizes only the activated subcarriers for transmission of data signals and pilot symbols. Let $\mathcal{K}$ be a set of activated subcarriers, then cardinality of the set $\mathcal{K}$ can be represented as $|\mathcal{K}|$.

For channel estimation, we place N_p $(\leq |\mathcal{K}|)$ pilot symbols $\{p_1, \ldots, p_{N_p}\}$ at subcarriers $k_1, k_2, \cdots, k_{N_P} \in \mathcal{K}$ ($k_1 < k_2 < \cdots < k_{N_P}$), which are known at the receiver. We assume that $N_p \geq L$ so that the channel can be perfectly estimated if there is no noise, and we denote the index of pilot symbols as $\mathcal{K}_p = \{k_1, \ldots, k_{N_P}\}$.

Let diag$(\boldsymbol{a})$ be a diagonal matrix with the vector $\boldsymbol{a}$ on its main diagonal. Collecting the received signals having pilot symbols as $\tilde{\boldsymbol{Y}} = [Y_{k_1}, \ldots, Y_{k_{N_p}}]^T$, we obtain

$$\tilde{\boldsymbol{Y}} = \boldsymbol{D}_{H_p} \boldsymbol{p} + \tilde{\boldsymbol{W}}, \tag{4}$$

where $\boldsymbol{D}_{H_p}$ is a diagonal matrix with its nth diagonal entry being H_{k_n} such that $\boldsymbol{D}_{H_p} = \mathrm{diag}\left(H_{k_1}, \ldots, H_{k_{N_p}}\right)$, and $\boldsymbol{p}$ is a pilot vector defined as $\boldsymbol{p} = \left[p_1, \ldots, p_{N_p}\right]^T$.

From $\tilde{\boldsymbol{Y}}$, we would like to estimate channel frequency responses for equalization and decoding. Thus, H_k for $k \in \mathcal{K}$ have to be estimated from $\tilde{\boldsymbol{Y}}$. In pilot-assisted modulation (PSAM) [6–8], a few known pilot symbols are embedded in an OFDM symbol to facilitate the estimation of unknown channel.

3. LEAST SQUARE CHANNEL ESTIMATION

We define $\boldsymbol{F}$ as an $N \times N$ DFT matrix, whose $(m+1, n+1)$th entry is $e^{-j2\pi mn/N}$. We denote an $N \times L$ matrix $\boldsymbol{F}_L = [\boldsymbol{f}_0, \ldots, \boldsymbol{f}_{N-1}]^{\mathcal{H}}$ consisting of N rows and the first L columns of the DFT matrix $\boldsymbol{F}$, where $\mathcal{H}$ is the complex conjugate transpose operator. We also define an $N_p \times L$ matrix $\boldsymbol{F}_p$ having $\boldsymbol{f}_{k_n}^{\mathcal{H}}$ for $k_n \in \mathcal{K}_p$ as its nth row. Then, we can express (4) as

$$\tilde{\boldsymbol{Y}} = \boldsymbol{D}_p \boldsymbol{F}_p \boldsymbol{h} + \tilde{\boldsymbol{W}}, \tag{5}$$

where the diagonal matrix $\boldsymbol{D}_p$ and channel vector $\boldsymbol{h}$ are respectively defined as $\boldsymbol{D}_p = \mathrm{diag}\left(p_1, \ldots, p_{N_p}\right)$, and $\boldsymbol{h} = \left[h_0, \ldots, h_{L-1}\right]^T$.

Let a vector having channel responses to be estimated,i.e., H_k for $k \in \mathcal{K}$, be $\boldsymbol{H}_s = [H_{k_1}, \ldots, H_{k_{|\mathcal{K}|}}]^T$. Similar to $\boldsymbol{F}_p$, we define a $|\mathcal{K}| \times L$ matrix $\boldsymbol{F}_s$ having $\boldsymbol{f}_{k_n}^{\mathcal{H}}$ for $k_n \in \mathcal{K}$ as its nth row, where $k_n < k_{n'}$ if $n < n'$. Then, we obtain

$$\boldsymbol{H}_s = \boldsymbol{F}_s \boldsymbol{h}. \tag{6}$$

Since (5) is linear, the Least Squares (LS) estimate $\hat{\boldsymbol{H}}_s$ of $\boldsymbol{H}_s$ is given by

$$\hat{\boldsymbol{H}}_s = \boldsymbol{F}_s \left(\boldsymbol{D}_p \boldsymbol{F}_p\right)^{\dagger} \tilde{\boldsymbol{Y}}, \tag{7}$$

where $(\cdot)^{\dagger}$ stands for the pseudo-inverse of a matrix. The LS estimate does not require any prior knowledge of the channel statistics and is thus widely applicable.

If we define the estimation error vector $\boldsymbol{E}_s = \hat{\boldsymbol{H}}_s - \boldsymbol{H}_s$, then the correlation matrix $\boldsymbol{R}_e$ of $\boldsymbol{E}_s$ can be expressed as

$$\boldsymbol{R}_e = E\{\boldsymbol{E}_s \boldsymbol{E}_s^{\mathcal{H}}\} = \boldsymbol{F}_s \left[\frac{1}{\sigma_w^2} \boldsymbol{F}_p^{\mathcal{H}} \boldsymbol{\Lambda}_p \boldsymbol{F}_p\right]^{-1} \boldsymbol{F}_s^{\mathcal{H}}, \tag{8}$$

where $E\{\cdot\}$ stands for the expectation operator and $\boldsymbol{\Lambda}_p$ is a diagonal matrix given by

$$\boldsymbol{\Lambda}_p = \boldsymbol{D}_p^{\mathcal{H}} \boldsymbol{D}_p = \operatorname{diag}\left(\lambda_1, \ldots, \lambda_{N_p}\right), \tag{9}$$

with $\lambda_n = |p_{k_n}|^2$ for $k_n \in \mathcal{K}_p$.

For a traditional OFDM symbol without null subcarriers, $\boldsymbol{F}_s^{\mathcal{H}} \boldsymbol{F}_s = c\boldsymbol{I}$ for a non-zero constant c, then $E\{||\boldsymbol{E}_s||^2\} = cE\{||\hat{\boldsymbol{h}} - \boldsymbol{h}||^2\}$, where $||\cdot||$ denotes the Euclidean norm. But, this is not always possible if there are null subcarriers in the OFDM symbol [6, 8]. Thus, for NC-OFDM equal powered pilot symbols in [2, 5] are not necessarily optimal.

Now, our objective is to find the optimal pilot symbols that minimizes the channel estimate MSE at the activated subcarriers, which is defined as

$$\eta = \left(\sum_{k \in \mathcal{K}} E\{|\hat{H}_k - H_k|^2\} \right)^{\frac{1}{2}} = (\operatorname{trace} \boldsymbol{R}_e)^{\frac{1}{2}}. \tag{10}$$

Note that, $\eta^2 = E\{||\hat{\boldsymbol{H}}_s - \boldsymbol{H}_s||^2\} \neq cE\{||\hat{\boldsymbol{h}} - \boldsymbol{h}||^2\}$ if $\boldsymbol{F}_s^{\mathcal{H}} \boldsymbol{F}_s \neq c\boldsymbol{I}$.

4. PILOT DESIGN AND CHANNEL ESTIMATION

In this section, for a given $\mathcal{K}$, we use convex optimization to optimally distribute power to these subcarriers. Then, we propose a cross entropy (CE) based algorithm to determine pilot set $\mathcal{K}_p$ with minimum channel estimate MSE. Once the near-optimal pilot set is obtained, we again employ convex optimization to distributed power to the selected pilot set.

Note that, an alternative approach is to select a pilot set with CE and then optimize the power distribution. However, the complexity is higher as it requires optimization of power to every set generated by the CE algorithm. In our numerical simulations, no significant difference in MSE performance between the two approaches, thus, we adopt the first approach to simplify the designs.

4.1. Pilot Power Distribution with SDP

For any fixed power to be utilized for channel estimation, we normalize the sum of pilot power such that

$$\sum_{k \in \mathcal{K}_p} |p_k|^2 = \sum_{k=1}^{N_p} \lambda_k = 1. \tag{11}$$

Our problem is to determine the optimal power distribution $\boldsymbol{\lambda} = [\lambda_1, \ldots, \lambda_{N_p}]^T$, which can be obtained by minimizing the η^2 in (10) with respect to $\boldsymbol{\lambda}$ under the constraints that

$$[1, \ldots, 1]\, \boldsymbol{\lambda} = 1, \quad \boldsymbol{\lambda} \succeq 0, \tag{12}$$

where $\boldsymbol{a} \succeq 0$ (or $\boldsymbol{a} \succ 0$) for a vector signifies that all entries of $\boldsymbol{a}$ are equal to or greater than 0 (or strictly greater than 0).

By re-expressing the nth row of $\boldsymbol{F}_p$ as $\tilde{\boldsymbol{f}}_n^{\mathcal{H}}$, our MSE minimization problem can be stated as [6]

$$\min_{\boldsymbol{\lambda}} \operatorname{trace} \left[\sigma_w^2 \left(\sum_{n=1}^{N_P} \lambda_n \tilde{\boldsymbol{f}}_n \tilde{\boldsymbol{f}}_n^{\mathcal{H}} \right)^{-1} \boldsymbol{R} \right] \tag{13}$$
$$\text{subject to} \quad [1, \ldots, 1]\, \boldsymbol{\lambda} \leq 1, \quad \boldsymbol{\lambda} \succeq 0,$$

where $\boldsymbol{R} = \boldsymbol{F}_s^{\mathcal{H}} \boldsymbol{F}_s$. This problem can be transformed into a semidefinite programming (SDP) form as demonstrated in [6] and the global optimal solution can be obtained by using convex optimization packages. Note that, for an optimal pilot set obtained by our proposed CE algorithm, the methods in [7, 8] can also be adopted for optimal power distribution by solving the optimization problem in (13).

4.2. Pilot subcarrier selection with CE Optimization

To determine the optimal set $\mathcal{K}_p$, i.e., the optimal location of N_p pilot symbols, we have to enumerate all possible sets, then optimize the pilot symbols for each set and compare them. This design approach becomes intractable as $|\mathcal{K}|$ gets larger.

Motivated by the effectiveness of the cross entropy (CE) method for finding near-optimal solutions in huge search spaces, this paper adopts the CE based method to search for the optimal position of the pilot symbol that minimizes the LS channel estimate MSE. Assume $\mathcal{K}_p$ to be a selected set from $\mathcal{K}$. Our optimal pilot sequence design can be formulated as a combinatorial optimization problem as

$$\mathcal{K}_p^{\star} = \arg \min_{\mathcal{K}_p^m \in \boldsymbol{\Omega}} \mathcal{C}_{sel}(\mathcal{K}_p^m), \tag{14}$$

where

$$\mathcal{C}_{sel}(\mathcal{K}_p^m) = \operatorname{trace} \left[\boldsymbol{F}_s \left(\boldsymbol{F}_p^{\mathcal{H}} \Lambda_p \boldsymbol{F}_p \right)^{-1} \boldsymbol{F}_s^{\mathcal{H}} \right], \tag{15}$$

represents the channel estimate MSE of the pilot set $\mathcal{K}_p^m$, and $\mathcal{K}_p^{\star}$ is the global optimal set of the objective function. The set $\mathcal{K}_p^m$ is given by

$$\mathcal{K}_p^m = \mathcal{K}(\{I_k\}_{k=1}^{|\mathcal{K}|} = 1), \quad I_k \in \{0, 1\}, \quad m = 1, \ldots, M, \tag{16}$$

where the indicator function I_k shows whether a subcarrier at the kth position is selected. The set of all $M = \binom{|\mathcal{K}|}{|\mathcal{K}_p|}$ possible subsets is denoted by $\boldsymbol{\Omega} = \{\mathcal{K}_p^1, \ldots, \mathcal{K}_p^M\}$, where $\binom{a}{b}$ denotes the possible combination set.

Applying the CE to solve (14), the first step is transforming the deterministic optimization problem (14) into a family of stochastic sampling problems [11]. Since the considered problem is on a discrete case, a family of Bernoulli probability density functions associated with the pilot symbol selection vector, $\boldsymbol{\omega} = \left[\omega_1, \omega_2, \ldots, \omega_{|\mathcal{K}|}\right]$, $\omega_k \in \{0, 1\}$, is given by

$$f(\boldsymbol{\omega}, \boldsymbol{p}) = \prod_{k=1}^{|\mathcal{K}|} p_k^{1_k(\boldsymbol{\omega})} (1 - p_k^{1 - 1_k(\boldsymbol{\omega})}) \tag{17}$$

where $\boldsymbol{p} = [p_1, p_2, \ldots p_{|\mathcal{K}|}]$ is a probability vector whose p_k entry indicates the probability of selecting the kth subcarrier, and the indicator function $1_k(\boldsymbol{\omega}) \in \{0, 1\}$ indicates whether the kth element of ω_k(the kth tone) is selected. If ω_k is selected, then $1_k(\omega_k) = 1$. Each element of $\mathcal{K}_p^m$ is modeled as an independent Bernoulli random variable with probability mass function $p(\omega_k = 1) = p_k$, and $p(\omega_k = 0) = 1 - p_k$, for $k = 1, \ldots |\mathcal{K}|$.

The CE method aims to find an optimal distribution $\boldsymbol{p}^\star$ that generates an optimal solution $\boldsymbol{\omega}^\star$ with minimum channel estimate MSE. However, $\boldsymbol{\omega}^\star$ occurs with a very small probability. In this case, (14) is associated with the problem of estimating the probability $Pr[\mathcal{C}_{sel}(\boldsymbol{\omega}) \leq \gamma]$ for a given threshold γ. To estimate the rare event, CE iteratively updates the probability vector $\boldsymbol{p}$ so that most samples generated by $f(\boldsymbol{\omega}; \boldsymbol{p})$ satisfy $\mathcal{C}_{sel} \leq \gamma$. By iteratively improving γ , $f(\boldsymbol{\omega}; \boldsymbol{p})$ eventually converges to an optimum probability density function $f(\boldsymbol{\omega}; \boldsymbol{p}^\star)$ and optimal $\boldsymbol{\omega}^\star$ can be obtained from $\boldsymbol{p}^\star$ by $f(\boldsymbol{\omega}; \boldsymbol{p}^\star)$.

A standard CE procedure for solving combinatorial problems contains two stages [11].

1. Adaptive updating of γ^t: For a given p^{t-1} generate $\mathcal{U}$ random samples $\{\boldsymbol{\omega}^{(t,u)}\}_{u=1}^{\mathcal{U}}$ from $f(.; \boldsymbol{p}^{t-1})$, where t denotes the iteration index of CE. Then, calculate the channel estimate MSE according to (14) to obtain a set of performance values $\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)})\}_{u=1}^{\mathcal{U}}$ and rank them in ascending order so that $\mathcal{C}_{sel}^1 \leq \ldots \leq \mathcal{C}_{sel}^{\mathcal{U}}$. Finally, assign
$$\gamma^t = \mathcal{C}_{sel}^{\lceil \rho \mathcal{U} \rceil} \tag{18}$$
where ρ denotes the fraction of the best samples and $\lceil \cdot \rceil$ is the ceiling operation. For description on how to select a suitable value of $\mathcal{U}$ see [11].

2. Adaptive updating of $\boldsymbol{p}^t$: For a given $\gamma^{(t)}$ and $\boldsymbol{p}^{(t-1)}$, use the same samples $\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)})\}_{u=1}^{\mathcal{U}}$ to update the parameter $\boldsymbol{p}^{(t)} = \left[p_0^{(t)}, p_1^{(t)}, \ldots, p_{|\mathcal{K}|}^{(t)}\right]$ via
$$p_k^{(t)} = \frac{\sum_{u=1}^{\mathcal{U}} 1_{\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)}) \leq \gamma^{(t)}\}} 1_k(\boldsymbol{\omega}^{(t,u)})}{\sum_{u=1}^{\mathcal{U}} 1_{\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)}) \leq \gamma^{(t)}\}}}, \tag{19}$$
where $\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)}) \leq \gamma^{(t)}\}$ is a variable defined by
$$1_{\{\mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)}) \leq \gamma^{(t)}\}} = \begin{cases} 1, \text{if } \mathcal{C}_{sel}(\boldsymbol{\omega}^{(t,u)}) \leq \gamma^{(t)} \\ 0, \text{otherwise.} \end{cases} \tag{20}$$

Note that in order to prevent fast convergence to a local optimum, parameter $\boldsymbol{p}^{t-1}$ is not updated to $\boldsymbol{p}^t$ directly; a smoothing factor, α , $0 \leq \alpha \leq 1$ was suggested by [11] to update (20) into
$$\boldsymbol{p}^t = \alpha \times p^t + (1 - \alpha) \times \boldsymbol{p}^{t-1}. \tag{21}$$
When $\alpha = 1$ the original updating formulation is achieved.

The detailed explanation of the CE algorithm for solving the combinatorial optimization problems can be found in [11].

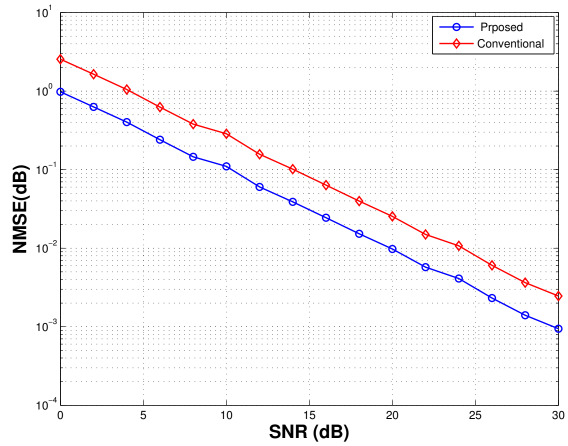

Fig. 1. Performance of the channel estimate MSE for different SNRs,$L = N_p = 8$

5. SIMULATION RESULTS

We demonstrate the effectiveness of our proposed pilot design through computer simulations. An OFDM transmission frame with $N = 256$ is considered. Out of 256 subcarriers, 200 are data carrying subcarriers used for data and pilots. Of the remaining 56 subcarriers, 28 are null in the lower frequency guard band while 27 are nulled in the upper frequency guard band and one is the central DC null subcarrier [12, p.429]. Of the 200 subcarriers used for signal transmission, a set $\mathcal{K}$ of activated subcarriers used by the cognitive secondary user is generated randomly as in [4]. The performance is measured in terms of the MSE and the bit error rate (BER) for a zero forcing (ZF) equalizer and $\mathcal{K} = 100$.

Fig. 1 shows the MSE of the channel estimator against SNR for the proposed scheme and the pilot symbols in [2, 5] which we refer to it as a conventional method for $L = N_p = 8$. From the result it is clear that, the proposed design outperforms the conventional method by a constant gap for different SNRs. This may be due to the fact that the conventional design consider equal powered pilots and optimizes only the position. To ensure better MSE performance, both pilot position and power distribution need to be careful considered. This result demonstrates the effectiveness of our proposed scheme that optimizes both the position and power of the pilot symbols.

Next we make comparison of the BER performance of the proposed pilot symbols, the conventional design in [2, 5] and the ideal case, i.e. known channel state information (CSI) for $L = N_p = 8$ and $L = N_p = 16$. Fig. 3 and 2 show BER performance of three schemes. The results show improved BER performance of the proposed design over the conventional design. This further substantiates the importance of optimizing both the pilot power and pilot positions in the NC-OFDM based systems.

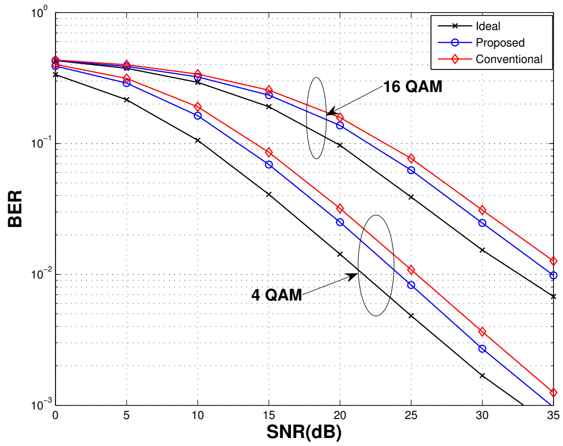

Fig. 2. Comparison of the BER performances for QAM signals for $L = N_p = 8$

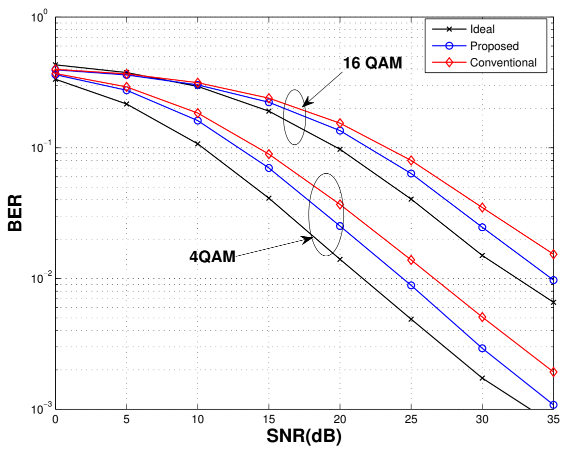

Fig. 3. Comparison of the BER performances for QAM signals for $L = N_p = 16$

6. CONCLUSION

In this paper we have presented a new pilot symbol design for channel estimation in NC-OFDM cognitive radio systems. We have demonstrated that, for NC-OFDM based systems, to obtain better performance both pilot power as well as pilot position need to be optimized. Simulation results show that, both MSE and BER performance of the proposed scheme outperforms the conventional design.

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