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Optimal Precoding in Conjunction with Windowing for Reduction in OOB Power of OFDM Based CR Systems

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Abstract In this paper we propose the design of an orthogonal precoder, in conjunction with optimal windowing, for reduction in out of band power in primary users' (PUs) frequency bands in OFDM based cognitive radio systems. First we derive the power spectral density expression to calculate the interference power in the PUs frequency bands. We then construct an interference minimization problem which is called a generalized Hermitian eigenvalue problem with an orthogonality constraint. The resulting precoding, in conjunction with optimal windowing, reduces the interference power when compared to only precoding and only windowing, at the cost of spectral efficiency.

Keywords Orthogonal frequency division multiplexing · Cognitive radio · Out of band power · Power spectral density · Windowing

1 Introduction

OFDM offers several advantages as compared to single carrier systems. It provides high spectral efficiency and also has the inherent advantage of converting frequency selective fading channels into at fading channels. In the context of cognitive radio systems, OFDM systems can adapt the transmit signal frequency band by simply deactivating the subcarriers in the frequency bands used by primary users. OFDM has been adopted as the modulation technique in many broadband wireless standards such as WLAN, IEE802.11a-n, IEEE802.16, WiMAX, LTE and 3GPP.

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Despite its several advantages, OFDM has some unique limitations like high out of band (OOB) power, sensitivity to frequency offset errors and peak to average power ratio. The high amount of OOB power result in serious interference to the systems operating in adjacent bands. The reduction of OOB power is of great importance in OFDM based CR systems [1]. For example, in cognitive radio, due to the high OOB power of the OFDM signal it is often not sufficient to de-activate sub-carriers which lie in the primary users' (PUs) band and hence further waveform shaping techniques are required.

Several methods have been proposed in literature for sidelobes suppression [2–7].

The OOB power performance is evaluated using the power spectral density analysis (PSD). In [8], power spectral density analysis of the analog representation and the DFT based transmitter of the OFDM system is discussed in details. The spectral analysis of OFDM systems directly using the DFT based representation based approach are more useful than using the analog waveform based approach in practical systems. An analytical expression for PSD in the DFT based transmitter model having precoding and rectangular windowing was derived in [9]. The analytical PSD expression of uncoded OFDM systems with arbitrary windows is given in [10].

Precoding method to reduce the OOB power in the DFT based transmitter model have been proposed [9, 11–13]. The concept of precoding is to introduce the correlation among the subcarriers and design in such in such a manner that sidelobes of subcarriers cancel each other. In these method they have considered the rectangular window.

Use of windowing for minimum spectral leakage in a DFT based OFDM transmitter has been proposed in [14–17]. In [14, 15], the window is designed for minimum spectral leakage. In [16], a windowed OFDM system for suppression of sidelobes is proposed but it results in considerable reduction in the bandwidth efficiency of the system. In [17], a transmit-windowing-based approach has been developed to reduce spectral leakage by allowing for a controlled amount of inter-carrier interference (ICI).

In this work we propose the design of an orthogonal precoder, in conjunction with windowing, to minimize the OOB power in PUs frequency bands in OFDM based CR systems. Precoding techniques are applied before the IDFT operation in the frequency domain and introduce correlation among the subcarriers. On the other hand windowing techniques are applied in the time domain after the IDFT operation. In our work we combine these two time and frequency domain approaches and formulate an objective function for minimum interference power in PUs' frequency bands.

We start with the derivation of the analytical expression for PSD with precoding and arbitrary windowing to calculate the interference power in the primary users' frequency bands. The window is considered to have the same length as the OFDM symbol length and cyclic prefix property as in [14]. We then construct an interference minimization problem which is called a generalized Hermitian eigenvalue problem with orthogonality constraint and obtain the optimal orthogonal precoder that minimizes the OOB power in PUs' frequency bands.

The paper is organized as follows. Section 2 describes the system model. In Sect. 3 the PSD expression is derived. while in Sect. 5 we present the design of the optimal precoder. Section 6 provides the OOB performance evaluation and Sect. 7 concludes the paper.

2 System Model

A system model of an OFDM transmitter having precoding and windowing is shown in Fig. 1. The data symbols to be transmitted are first passed through a serial-to-parallel converter to form $N \times 1$ input vectors $\mathbf{c}^l$, where l represents the transmitted symbol index.

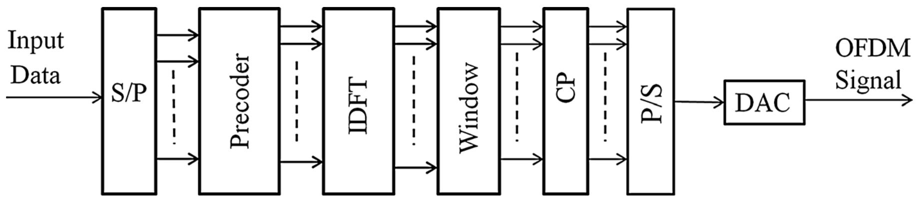

Fig. 1 Block diagram of windowed OFDM transmitter with precoding

These data vectors are pre-multiplied with an $M \times N$ precoding matrix $\mathbf{G}$, where $M > N$. The precoder output is multiplied by an $M \times M$ IDFT matrix. The output vector is multiplied by the window vector $\tilde{\mathbf{w}}$, followed by addition of a cyclic prefix of length L. Since we consider the window length to be equal to the transmitted OFDM symbol length, the window vector $\tilde{\mathbf{w}}$ is $M \times 1$. The output analog signal $y(t)$ is obtained by passing the discrete-time baseband signal $y^l[n]$ through a digital-to-analog converter (DAC), which consists of a sampler (with sampling time T_s) and an interpolation filter h(t). This system model is a general case of system model described in [9] where rectangular window is assumed.

Mathematically, the precoded mth element of the lth OFDM symbol can be expressed as

$$d^l[m] = \sum_{n=0}^{N-1} g_{m,n} c^l[n], \quad m = 0, 1, \ldots, M-1. \tag{1}$$

where $c^l[n]$ is the nth element of the lth data symbol $\mathbf{c}^l$, $g_{m,n}$ is the (m, n)th element of the precoding matrix $\mathbf{G}$.

The precoded output after the inverse DFT block can be written as

$$a^l[k] = \sum_{n=0}^{M-1} d^l[n] e^{(j2\pi(k-L)n/M)}, \quad k = 0, 1, \ldots, M-1. \tag{2}$$

The precoded $\mathbf{a}^l$ block is passed through an arbitrary window $\mathbf{w}_{arb}$ and can be expressed as

$$x^l[k] = \sum_{i=-\infty}^{\infty} \sum_{n=0}^{M-1} d^l[n] e^{(j2\pi(k-L)n/M)} w_{arb}[n - iM], \quad k = 0, 1, \ldots, M-1. \tag{3}$$

Assuming the window length to be the same as the OFDM symbol length M i.e.,

$$\tilde{w}[n] = 0, \quad n \notin \{0, 1, \ldots, M-1\}.$$

the output after windowing is obtained as

$$z^l[k] = \tilde{w}[k] x^l[k], \quad k = 0, 1, \ldots, M-1. \tag{4}$$

A cyclic prefix of length L is added to $\mathbf{z}^l$. This can be interpreted as if both the inverse DFT output $\mathbf{x}^l$ and the window $\tilde{\mathbf{w}}$ are cyclically extended. The inverse DFT output is of the form $\mathbf{x}^l = \left[x^l[M-L], x^l[M-L+1], \ldots, x^l[0], x^l[1], \ldots, x^l[M-1]\right]^T$. Similarly, the last L samples of window $\tilde{\mathbf{w}}$ are also added as prefix to form the vector

$$\mathbf{w} = [\tilde{w}[M-L], \ldots, \tilde{w}[M-1], \tilde{w}[0], \tilde{w}[1], \ldots, \tilde{w}[M-1]]^T$$

The output of the cyclic extension block is

$$y^l[k] = \mathbf{W}[k,k] \sum_{n=0}^{M-1} d^l[n] e^{(j2\pi(k-L)n/M)}, \quad k = 0, 1, \ldots, K-1. \tag{5}$$

where $K = M + L$ and $\mathbf{W} = diag(\mathbf{w})$ is $K \times K$ diagonal matrix obtained from vector $\mathbf{w}$.

After parallel-to-serial conversion, $y^l[k]$ is passed through the D/A operation. Using the representation given in [9], he resulting OFDM signal $y(t)$ can be written as

$$y(t) = \sum_{l=-\infty}^{\infty} \sum_{k=0}^{K-1} y^l[k] h(t - (lK + k)T_s) \tag{6}$$

In next section, we derive the analytical expression of the PSD as a function of precoding matrix and windowing.

3 OFDM Transmit Signal Spectrum as a Function of Precoder and Window

To obtain the analytical expression for PSD of the transmitted signal y(t), we proceed in the same manner as given in [9].

The PSD of the output $y(t)$ of the interpolation filter is given as

$$\begin{aligned} P(j\Omega) &= \mathbb{E}\left[|\mathcal{F}\{y(t)\}|^2 \right] \\ &= |H(j\Omega)|^2 \cdot \mathbb{E}\left[\left| Y^l(j\Omega) \right|^2 \right] \end{aligned} \tag{7}$$

where $\mathcal{F}$ denotes the Fourier transform operator, $H(j\Omega)$ is the frequency response of the interpolation filter and $Y^l(j\Omega)$ is the frequency response of the lth transmitted OFDM symbol which can be expressed as

$$Y^l(j\Omega) = \sum_{m=0}^{M-1} d_n^l \sum_{k=0}^{K-1} \mathbf{W}[k,k] e^{(j2\pi(k-L)n/N)} e^{(-jk\Omega T_s)} \tag{8}$$

From (1) and (8), $Y^l(j\Omega)$ can be expressed as

$$Y^l(j\Omega) = \sum_{m=0}^{N-1} c_m^l \sum_{n=0}^{M-1} g_{m,n} \sum_{k=0}^{K-1} \mathbf{W}[k,k] e^{(j2\pi(k-L)n/N)} e^{(-jk\Omega T_s)} \tag{9}$$

We can also write (9) in matrix form as

$$Y^l(j\Omega) = \mathbf{e}(j\Omega)\mathbf{W}\mathbf{F}^H\mathbf{D}\mathbf{G}\mathbf{c}^l$$

where $\mathbf{e}(j\Omega)$ is a $1 \times K$ vector with kth entry $e^{(-j\Omega k T_s)}$, $\mathbf{F}$ is an $M \times K$ matrix with $[F]_{m,k} = e^{(-j2\pi mk/M)}$ and $\mathbf{D}$ is an $M \times M$ diagonal matrix with kth entry $e^{(-j2\pi L(k-1)/N)}$. Here $\mathbf{c}^l$ represents the $N \times 1$ data vector on the lth OFDM symbol that is taken from an M-ary quadrature amplitude modulated (M-QAM) constellation with the assumption that $\mathbb{E}[\mathbf{c}^l] =$

0 and covariance matrix $\mathrm{Cov}(\mathbf{c}^{\mathrm{l}}) = \sigma^2 \mathbf{I}_{\mathrm{N}}$ where $\mathbf{I}_N$ is an $N \times N$ identity matrix. Then $\mathbb{E}\left[\left|Y^l(j\Omega)\right|^2\right]$ simplifies to

$$\mathbb{E}\left[\left|Y^l(j\Omega)\right|^2\right] = \sigma^2 \mathrm{tr}\{\mathbf{G}^{\mathrm{H}}\mathbf{D}^{\mathrm{H}}\mathbf{F}\mathbf{W}\mathbf{e}^{\mathrm{H}}(\mathrm{j}\Omega)\mathbf{e}(\mathrm{j}\Omega)\mathbf{W}\mathbf{F}^{\mathrm{H}}\mathbf{D}\mathbf{G}\} \tag{10}$$

where tr(.) represents the trace of a matrix.

From (7) and (10), the expression for PSD can be written as

$$P(j\Omega) = |H(j\Omega)|^2 \cdot \sigma^2 tr\{\mathbf{G}^H\mathbf{D}^H\mathbf{F}\mathbf{W}\mathbf{e}^H(j\Omega)\mathbf{e}(j\Omega)\mathbf{W}\mathbf{F}^H\mathbf{D}\mathbf{G}\} \tag{11}$$

For the case of rectangular windowing, **W** will be an identity matrix of size $M \times M$ and then the PSD expression becomes

$$P_{rect}(j\Omega) = |H(j\Omega)|^2 \cdot \sigma^2 tr\{\mathbf{G}^H\mathbf{D}^H\mathbf{F}\mathbf{e}^H(j\Omega)\mathbf{e}(j\Omega)\mathbf{F}^H\mathbf{D}\mathbf{G}\}$$

which is the same as given in [9, Eq. (7)]. Thus the expression given in [9] is a special case of the expression obtained in (11).

From (7) it can be seen that the power spectral density expression of the transmitted OFDM signal is a product of the two terms $|H(j\Omega)|^2$ and $Y^l(j\Omega)$ which can be interpreted in the following manner as given in [9]. The interpolation filter is a low-pass filter characterised by its cut-off frequency, which is π/T_s in this case. Within the OFDM band, $|H(j\Omega)|^2$ is constant. The out of band radiation of the transmitted OFDM signal outside the OFDM spectrum depends on the design of $H(j\Omega)$ and not on the precoding matrix. The second term $Y^l(j\Omega)$ is interpreted as the sum of spectra of individual M subcarriers and is periodic with period $2\pi/T_s$ and hence the PSD of the transmit signal within $2\pi/T_s$ (i.e. within the OFDM band) depends on $Y^l(j\Omega)$.

The objective in this work is to minimize the OOB interference in the PUs' bands that lies within the OFDM sub-band in OFDM based CR systems. The subcarriers that lies in PUs' bands are deactivated but can still cause interference to PUs because of high OOB power. Therefore the focus is on $Y^l(j\Omega)$.

From (9), $Y^l(j\Omega)$ is a function of the precoding matrix **G** and windowing **W**. Precoding introduces correlation among the subcarriers and it can be designed to cancel the sidelobes of other subcarriers and thus reduce the OOB power in the desired PUs' frequency bands. From (10), the OOB power within the OFDM sub-band can be reduced by jointly optimizing the windowing and precoding. In this work we use sequential design to minimize the OOB interference power in the PU band. In the first step we obtain the window having the minimum spectral leakage using the result in [14]. In the second step we formulate the optimization problem based on the precoding matrix.

4 Review of the Optimal Window for Minimum OOB Power Criteria

The design of the optimal window itself having cyclic prefix property for minimum OOB power leakage proposed in [14] is revisited in this section.

Let $\mathbf{w}$ be the $K \times 1$ column vector and $\tilde{\mathbf{w}}$ be the $M \times 1$ column vector containing the window elements. The cyclic-prefix property can be written as

$$\mathbf{w} = \mathbf{U}\tilde{\mathbf{w}},$$

where

$$U = \begin{bmatrix} \mathbf{0}_{L\times M-L} & & I_{L\times L} \\ & I_{M\times M} & \end{bmatrix}$$

The Fourier transform of the window is

$$\mathbf{w}(e^{j\omega}) = \mathbf{e}(e^{j\omega})\mathbf{U}\tilde{\mathbf{w}}$$

where, $\mathbf{e}(e^{j\omega}) = (1, e^{-j\omega}, e^{-j2\omega}, \ldots, e^{-j(K-1)\omega})$

It follows that the squared magnitude response of the window is

$$|\mathbf{w}(e^{j\omega})|^2 = \tilde{\mathbf{w}}^H\mathbf{U}^T\mathbf{E}(e^{j\omega})\mathbf{U}\tilde{\mathbf{w}} \tag{12}$$

where, $\mathbf{E}(e^{j\omega}) = \mathbf{e}^H(e^{j\omega})\mathbf{e}(e^{j\omega})$.

The stop band energy, that is, the out of band leakage power of the window is given as

$$S = \tilde{\mathbf{w}}^\dagger\mathbf{U}^T\int_{\omega_s}^{2\pi\omega_s}\mathbf{E}(e^{j\omega})\frac{d\omega}{2\pi}\mathbf{U}\tilde{\mathbf{w}} = \tilde{\mathbf{w}}^\dagger\mathbf{U}^T\mathbf{Q}\mathbf{U}\tilde{\mathbf{w}} \tag{13}$$

where $\mathbf{Q}$ is an $N \times N$ real, symmetric and positive semi-definite matrix.

We can see that the minimization of stop band energy S becomes the minimization of $\tilde{\mathbf{w}}^\dagger\mathbf{U}^T\mathbf{Q}\mathbf{U}\tilde{\mathbf{w}}$. Since, the matrix $\mathbf{U}^T\mathbf{Q}\mathbf{U}$ is positive semi-definite, the objective function $\tilde{\mathbf{w}}^\dagger\mathbf{U}^T\mathbf{Q}\mathbf{U}\tilde{\mathbf{w}}$ can be minimized by choosing $\tilde{\mathbf{w}}$ to be the eigenvector corresponding to the smallest eigenvalue of $\mathbf{U}^T\mathbf{Q}\mathbf{U}$. $\mathbf{U}^T\mathbf{Q}\mathbf{U}$ is real. Hence, the optimal window $\mathbf{W} = diag(\tilde{\mathbf{w}})$ has real coefficients.

5 Design of Optimal Precoder

To obtain the precoding matrix if we only minimize the power in the PU band, the spectrum of the active data subcarrier band close to the primary user band also changes. Hence the design problem is to minimize the *ratio* of the power in the frequency band where subcarriers are deactivated to the power in the frequency band where data subcarriers are present. This ratio is referred to as the contrast energy ratio (CER) [9].

$$CER = \frac{\int_{f\in\Omega_{NC}} \mathrm{E}\left[|\mathrm{Y}^{\mathrm{l}}(\mathrm{j}\Omega)|^2\right]\mathrm{d}\Omega/2\pi}{\int_{f\in\Omega_{DC}} \mathrm{E}\left[|Y^{l}(j\Omega)|^2\right]d\Omega/2\pi} \tag{14}$$

Here Ω_{NC} represents the frequency sub-bands within the OFDM bands where subcarriers are deactivated and Ω_{DC} represents the frequency sub-bands where data subcarriers are present.

From (10), the numerator and denominator terms in the CER expression (14) reduces to $tr(\mathbf{G}^H\mathbf{D}^H\mathbf{F}\mathbf{W}\mathbf{P}_1\mathbf{W}\mathbf{F}^H\mathbf{D}\mathbf{G})$ and $tr(\mathbf{G}^H\mathbf{D}^H\mathbf{F}\mathbf{W}\mathbf{P}_2\mathbf{W}\mathbf{F}^H\mathbf{D}\mathbf{G})$ respectively. Here $\mathbf{P}_1$ and $\mathbf{P}_2$ are obtained by solving the following integrals

$$\mathbf{P}_1 = \int_{f\in\Omega_{NC}} \mathbf{e}^H(j\Omega)\mathbf{e}(j\Omega)\, d\Omega/2\pi$$

$$\mathbf{P}_2 = \int_{f\in\Omega_{DC}} \mathbf{e}^H(j\Omega)\mathbf{e}(j\Omega)\, d\Omega/2\pi$$

The precoding matrix is optimized based on minimizing the CER. Now the CER minimization problem can be formulated as

$$\begin{array}{ll} \underset{\mathbf{G}}{\text{minimize}} & \left(\frac{\text{tr}(\mathbf{G}^{\text{H}}\mathbf{A}\mathbf{G})}{\text{tr}(\mathbf{G}^{\text{H}}\mathbf{B}\mathbf{G})}\right) \\ \text{subject to} & \mathbf{G}^H\mathbf{G} = \mathbf{I} \end{array} \tag{15}$$

where $\mathbf{A} = \mathbf{D}^H\mathbf{F}\mathbf{W}\mathbf{P}_1\mathbf{W}\mathbf{F}^H\mathbf{D}$ and $\mathbf{B} = \mathbf{D}^H\mathbf{F}\mathbf{W}\mathbf{P}_2\mathbf{W}\mathbf{F}^H\mathbf{D}$ are Hermitian matrices since $\mathbf{P}_1$ and $\mathbf{P}_2$ are Hermitian. Also, $\mathbf{W} = \mathbf{W}^H$ because $\mathbf{W}$ is diagonal matrix with real entries.

Note that in the absence of the condition $\mathbf{G}^H\mathbf{G} = \mathbf{I}$, if a matrix $\mathbf{G}$ is a solution to the problem then so is any nonzero scalar multiple a of $\mathbf{G}$. This can be seen from

$$\left(\frac{\text{tr}((\text{a}\mathbf{G}^{\text{H}})\mathbf{A}(\text{a}\mathbf{G}))}{\text{tr}((\text{a}\mathbf{G}^{\text{H}})\mathbf{B}(\text{a}\mathbf{G}))}\right) = \left(\frac{a^2\text{tr}(\mathbf{G}^{\text{H}}\mathbf{A}\mathbf{G})}{a^2\text{tr}(\mathbf{G}^{\text{H}}\mathbf{B}\mathbf{G})}\right) = \left(\frac{\text{tr}(\mathbf{G}^{\text{H}}\mathbf{A}\mathbf{G})}{\text{tr}(\mathbf{G}^{\text{H}}\mathbf{B}\mathbf{G})}\right) \tag{16}$$

We also note that orthogonal precoding matrices do not change the transmitted power or the peak to average power ratio and also help in decoding at the receiver [9].

The formulated objective function in (15) is a tr minimization problem with orthogonality constraint. This problem can be treated as a generalized Hermitian eigenvalue problem and optimal $\mathbf{G}$ provides the basis for the invariant subspace of eigenvectors corresponding to the M smallest eigenvalues of $B^{-1}A$ [9].

6 Numerical Results and Discussion

In this section we calculate the precoding matrix $\mathbf{G}$ and examine its ability to suppress sidelobes by plotting the PSD. We obtained the precoding matrix using the conventional eigenvalue solver approach. Thus we find a matrix, say $\mathbf{E}$, from the N eigenvectors corresponding to the smallest N eigenvalues of $\mathbf{B}^{-1}\mathbf{A}$. Since $\mathbf{B}^{-1}\mathbf{A}$ may or may not be Hermitian, the eigenvectors may not be orthonormal. To satisfy the orthogonality constraint, we calculate the optimal $\mathbf{G}$ that provides an orthonormal basis for the range of $\mathbf{E}$ i.e. the columns of $\mathbf{G}$ span the same space as the columns of $\mathbf{E}$.

The input data symbols are taken from a 4-QAM signal constellation with the length of data vector $N = 52$, dimension of precoding matrix 64×52 i.e. $M = 64$ and cyclic prefix of length $L = 8$. The interpolation filter was taken as a rectangular pulse and the normalized PSD is plotted using (11). Figure 2 shows the time domain response of the optimal window obtained for N = 64 and cyclic prefix of length 8. In Fig. 3 we plot the frequency response of this window.

In Fig. 4 we plot the normalized PSD for the cases of optimal windowing, optimal precoding along with a rectangular window, optimal precoding along with optimal windowing and only rectangular windowing for $N = 128$, $L = 8$ and $M = 52$. The PU band has been assumed to have a width of 10 subcarriers from subcarrier count 27–36. In Fig. 5 we assume that everything is same as before except that the PU band has a

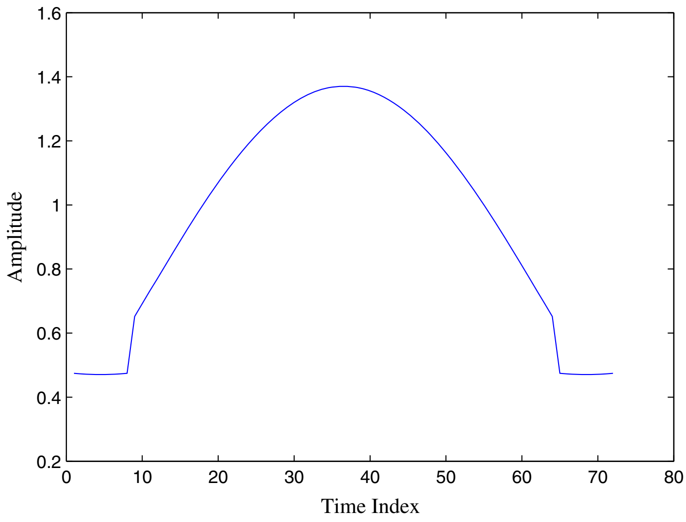

Fig. 2 Time response of optimal window of length 64

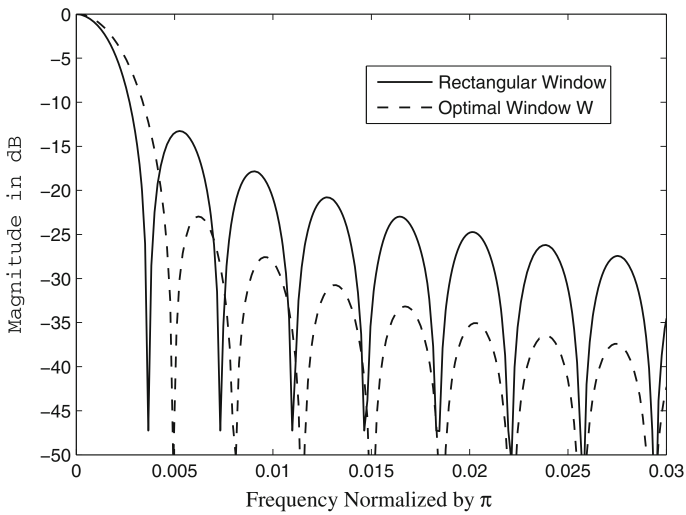

Fig. 3 Frequency response of optimal window of length 64

width of 20 subcarriers from subcarrier count 23–42. In both the cases considered, the proposed system provides 5 dB less OOB interference power compared to optimal precoding along with a rectangular window, 10 dB improvement compared to optimal windowing and 14 dB improvement compared to only rectangular windowing with no precoding.

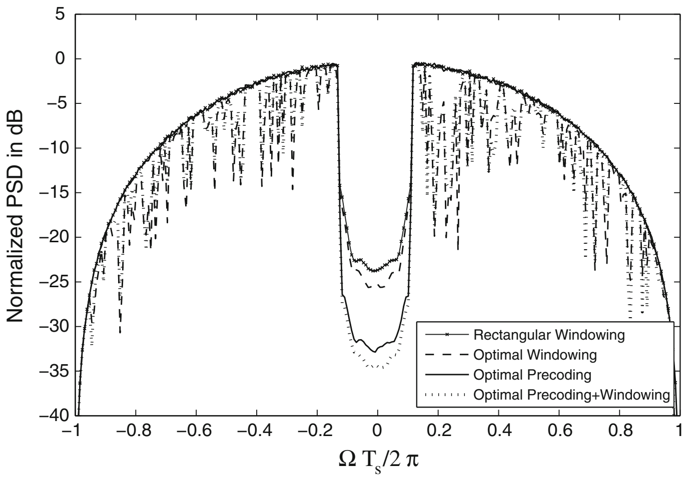

Fig. 4 PSD comparison with 10 deactivated subcarriers

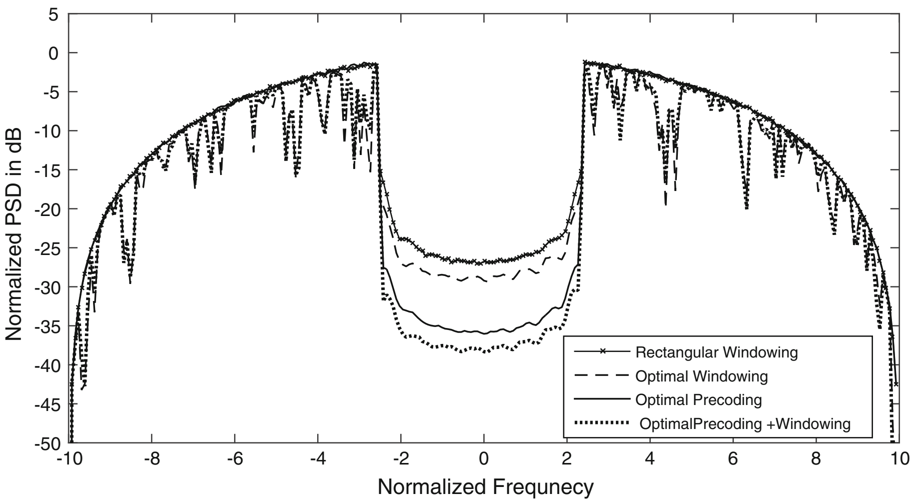

Fig. 5 PSD comparison with 20 deactivated subcarriers

7 Conclusions

We considered optimal precoding in conjunction with optimal windowing for OOB power reduction in OFDM based CR systems. An optimal precoding matrix is obtained by minimizing the CER. The minimization of CER is formulated as a generalized Hermitian eigenvalue problem with an orthogonality constraint. The combined techniques leads to reduction in OOB power compared to only precoding and only windowing methods.

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