Out-of-Band Radiation Reduction in OFDM-based Cognitive Radio Systems

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***Abstract*—Orthogonal frequency division multiplexing (OFDM) suffers from significant out-of-band radiation due to high sidelobes of modulated subcarriers. In a cognitive radio (CR) system this effect results in high interference between secondary and primary users (PUs). Existing methods for reducing this interference have the drawbacks of significant data transmission rate reduction or high complexity. In this paper a simple and adaptive iterative method is proposed that treats on OFDM signal in both time and frequency domains. Simulation results show that when this algorithm is combined with differential modulation, a high value of sidelobe suppression is achieved while the transmission rate reduction is almost insignificant.**

Keywords-Cognitive radio; opportunistic spectrum access; OFDM; sidelobe supression.

I. Introduction

Due to the increasing request for wireless applications with high data rate and scarcity of spectrum, many studies are considering methods to achieve higher efficiency in spectrum usage. Cognitive Radio (CR), developed by Miltola [1], is a promising way to tackle this problem. The CR design is an innovative radio design philosophy that involves smartly sensing the swaths of spectrum and then determining the transmission characteristics of a group of secondary users (SUs) based on interaction with its environment with objectives of highly reliably communication and efficient utilization of these available spectrum resources.

Orthogonal frequency division multiplexing (OFDM) has been proposed as an appropriate transmission technique for CR systems due to its great flexibility in dynamically allocating unused spectrum among CR users, and the ease of analysis of the PU's spectral activity [2]. However, modulated OFDM subcarriers suffer from high sidelobes, which result in an interaction between the licensed system and the OFDM based rental system due to the non-orthogonality of their respective transmit signals [3]. Thus disabling a set of OFDM subcarriers to create a spectrum null may not be sufficient to avoid interference to PUs.

Well-known techniques for sidelobe suppression are windowing the transmission signal in time domain [4] and the insertion of guard bands [5]. With the former method, the symbol duration is prolonged, where as with the latter the scarce spectral resources are wasted. Both methods reduce the system throughput and, in general, do not achieve sufficient sidelobe suppression for overlay systems. Other proposed algorithms include the use of interference cancellation carriers (CC) [6], [7] and subcarrier weighting [8]. While CC technique can significantly suppress OFDM sidelobes, it results in an increase in the system peak-to-average-power ratio (PAPR) and the performance is sensitive to the cyclic prefix (CP) size. On the other hand, subcarrier weighting method causes an increase in the system bit error rate (BER), and the interference reduction is not as significant as it is with the CC method. Two optimal methods are proposed in [9] and [10] that both suffer high complexity which makes them impractical and moreover in [9] transmission rate reduction is significant and [10] needs to know the channel coefficient between SU transmitter and PU receiver which is not a practical assumption. On the other hand, in the optimal scheme proposed in [10], the gain values achieved are data-dependent because even if the modulated data is of constant amplitude, as in the case of QPSK modulation, the phase of data symbols will affect the total interference when their interferences are adding up.

In this paper a simple and adaptive iterative method is proposed in which the OFDM subcarriers out-of-band sidelobes can be suppressed to the desired level. When this algorithm is accompanied with differential modulation, the system throughput will maintain almost unchanged and BER is significantly less than other mentioned methods.

The rest of this paper is organized as follows. In section II the system model is described. In section III proposed method formulation and its performance is discussed. In section IV we show that this algorithm does not affect the signal PAPR. In section V two suboptimal schemes of [10] are introduced. Simulation results are presented in section VI. Finally, conclusion is given in section VII.

II. System Model

We consider a CR system employing OFDM as the signaling scheme. It is assumed that the CR is aware of the surrounding environment and the radio channel characteristics.

After scanning the channel, the occupied bands are sensed by the CR system and the unoccupied bands are considered for possible transmission on each side of PU bands. The opportunistic spectrum access algorithms are derived to achieve the highest possible spectral efficiency while keeping the interference to PUs in an acceptable level.

The system model is shown in Fig. 1. An OFDM system with N subcarriers is considered. The input encoded bits are symbol-mapped onto N complex data symbols d_k, k=1, …, N from a phase-shift keying (PSK) or quadrature amplitude modulation (QAM) signal constellation $X^{(m)} = [d_1, ..., d_N]$ and is then fed to an inverse fast Fourier transform (IFFT) block

$$x^{(m)} = \frac{1}{N} F_{N,N}^{*} X^{(m)} \qquad (1)$$

Where $F_{N,N}$ is the N point Fourier transform matrix of a vector of length N or

$$F_{n_1,n_2} = \exp\left(\frac{-j2\pi n_1 n_2}{N}\right) \qquad (2)$$

The cognitive engine passes required information regarding PU operating bands to both the subcarrier mapper and sidelobe suppression block. This information is used to disable subcarriers operating in PU bands and to suppress the interference caused by OFDM sidelobes to the PUs as explained in the following section.

III. THE PROPOSED ADAPTIVE ITERATIVE ALGORITHM

Let's assume the CR system detects a PU signal spanning over K subcarriers from k_1 to $k_1 + K - 1$, where $K\Delta f$ is the primary signal bandwidth, and Δf is the frequency subcarrier spacing. The above subcarrier set is disabled to avoid interference with the PU so $X^{(m)} = [d_1 ... d_{k_1 - 1}\, 0 ... 0\, d_{k_1 + K} ... d_N]$ and $x^{(m)}$ is its inverse Fourier transform.

The adaptive iterative proposed algorithm reduces the interference exerted to PU to the desired level by choosing the number of iterations and it consists of five steps as follows:

Step 1: As in [7] and [9] we perform $\gamma \times N$ length FFT on the N-length time-domain vector $x^{(m)_0} = x^{(m)}$ to achieve γN-

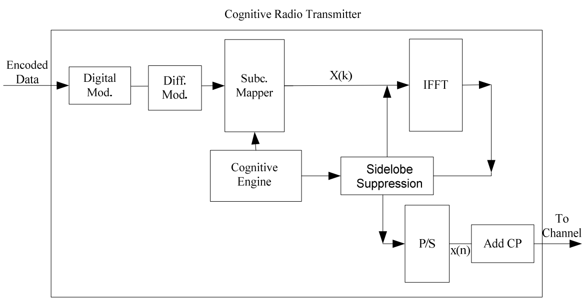

Fig. 1. *System Model*

length frequency-domain vector $X_\gamma^{(m)_0}$ which contains interference values spanning over $K\gamma$ indexes from $(k_1 - 1)\gamma + 1$ to $(k_1 + K - 1)\gamma$ ($k\gamma$ indexes in this region are located on the nulls). The subscripts 0 and γ stand for being in 0_{th} iteration and having the length of $\gamma \times N$ respectively.

Step 2: In the second step $X_\gamma^{(m)_0}$ elements over the indexes pertaining to the interference region are set to zero.

Step 3: In this step $\gamma \times N$ length IFFT is performed on the new $X_\gamma^{(m)_0}$ and vector $x_\gamma^{(m)_1}$ is achieved.

Step 4: In this step vector $x^{(m)_1}$ is defined as the first N elements of $x_\gamma^{(m)_1}$. A new iteration begins by replacing $x^{(m)_0}$ in step1 with $x^{(m)_1}$.

In *Fig. 2* the power spectral density (PSD) of the secondary users' OFDM transmitted signal for different number of iterations, L, is shown. It can be deduced from *Fig. 2* that by each increasing of the number of iterations to 10 times, the interference power level is reduced by almost 10dB. Therefore L can be chosen appropriately depending on the tolerable interference threshold of the detected PUs.

From the above steps the relation between signal vectors in two consecutive iterations can be derived. In the frequency domain we have

$$\begin{aligned} X_\gamma^{(m)_i}(k) &= fft\left\{ifft\left\{X_\gamma^{(m)_{i-1}}(k) \times G(k)\right\} \times w(n)\right\} \\ &= \left(X_\gamma^{(m)_{i-1}}(k) \times G(k)\right) * W \qquad 1 \le k, n \le \gamma \times N \end{aligned} \qquad (3)$$

Thus in the time domain we have

$$x_\gamma^{(m)_i}(n) = \left(x_\gamma^{(m)_{i-1}}(n) * g(n)\right) \times w(n) \qquad 1 \le k, n \le \gamma \times N \qquad (4)$$

where

$$w(n) = \begin{cases} 1 & 1 \le n \le N \\ 0 & n > N \end{cases} \qquad (5)$$

and

$$\begin{aligned} g(n) &= ifft\left\{1 - \Pi\left(\left(n - \left(k_1 + (K-1)/2\right)\gamma\right) / K\gamma\right)\right\} \\ &= \delta(n) - \frac{K}{N} \operatorname{sinc}\left(\frac{K}{N} \times n\right) \times e^{-j\left(k_1 + (K-1)/2\right)\gamma n} \end{aligned} \qquad (6)$$

In which $\Pi\left(\left(k - \left(k_1 + (K-1)/2\right)\gamma\right) / K\gamma\right)$ is a rectangular window of length Kγ and is centered at $\left(k_1 + (K-1)/2\right)\gamma$.

so the relation between time-domain signals in two consecutive iterations is

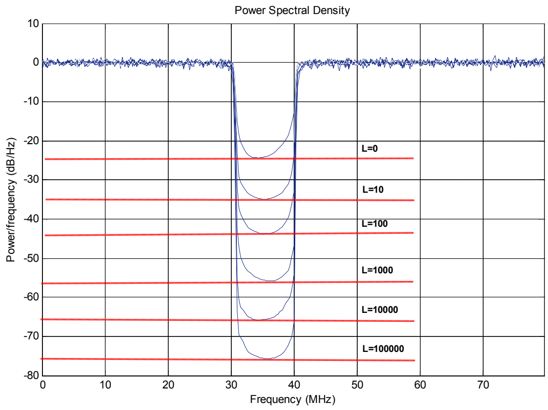

Fig. 2. Power spectral density of the secondary users OFDM signal for different number of iterations

$$x_\gamma^{(m)_i}(n)=\begin{cases} x_\gamma^{(m)_{i-1}}(n)-x_\gamma^{(m)_{i-1}}(n)*\frac{K}{N}\sin c\left(\frac{K}{N}\times n\right)\times e^{-j\left(k_1+(K-1)/2\right)\gamma n} \\ \qquad 1\le n\le N \\ 0 \qquad n>N \end{cases} \tag{7}$$

In other words

$$x_\gamma^{(m)_i}(n)=x_\gamma^{(m)_{i-1}}(n)$$
$$-\frac{K}{N}\times e^{-j\left(k_1+(K-1)/2\right)\gamma n}\sum_{j=1}^{N}\left(x_\gamma^{(m)_{i-1}}(j)\sin c\left(\tfrac{K}{N}(n-j)\right)\times e^{j\left(k_1+(K-1)/2\right)\gamma j}\right) \tag{8}$$
$$1\le n\le N$$

The second term in (8) defines the distortion inserted to the time domain signal in the ith iteration. Thus as the number of iterations increases, total exerted distortion is increased. In Fig. 3 the effect of this distortion on the transmitted bits for a large number of iterations (L=100000) is shown and is compared with the original bits (symbols are BPSK modulated). To overcome this change, we consider the fact that the inserted distortion performs like a piecewise offset added to data symbols. In other words, the amount of distortion exerted to consecutive symbols is almost the same. In Fig. 4 the distortion difference between consecutive symbols for above signal is shown. As it can be seen, the amount of distortion difference except for 3 symbols on each side of PU band is insignificant. From this it can be inferred that by using a differential modulation block after the conventional modulation we can shield most of the transmitted symbols form the distortion. For example for the BPSK signal, final transmitted symbols bi will be

$$b_i=\begin{cases} d_i & i=1 \\ 1 & d_i=d_{i-1}=1 \quad or \quad d_i=d_{i-1}=-1 \\ -1 & d_i=-d_{i-1}=1 \quad or \quad d_i=-d_{i-1}=-1 \end{cases} \tag{9}$$

The same scenario can be performed for other types of PSK or QAM modulations.

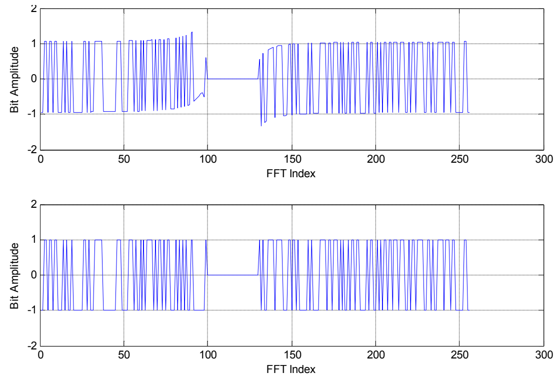

Fig. 3. Distorted transmitted symbols after L=100000 iterations compared to their original values for a BPSK signal

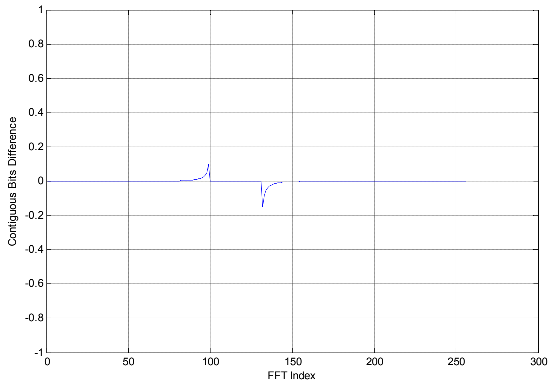

Fig. 4. Consecutive symbols distortion difference for L=100000 iterations

IV. PROPOSED ALGORITHM IN THE SENSE OF PAPR

To study the effect of proposed algorithm on the signal's PAPR value, first let check this parameter for the second term or the distortion term of (8). For PAPR value of this term we have

$$PAPR_{dt_i}=$$
$$\frac{\left|\sum_{j=1}^{N}\left(x_\gamma^{(m)_{i-1}}(j)\sin c\left(\frac{K}{N}(N/2-j)\right)e^{j\left(k_1+(K-1)/2\right)\gamma j}\right)\right|^2}{\frac{1}{N}\sum_{n=1}^{N}\left|\sum_{j=1}^{N}\left(x_\gamma^{(m)_{i-1}}(j)\sin c\left(\frac{K}{N}(-n+j)\right)e^{j\left(k_1+(K-1)/2\right)\gamma j}\right)\right|^2}\le$$
$$\frac{N\sum_{j=1}^{N}\left|\sin c\left(\frac{K}{N}(-j+N/2)\right)\right|^2\times\sum_{j=1}^{N}\left|x_\gamma^{(m)_{i-1}}(j)\right|^2}{\sum_{n=1}^{N}\sum_{j=1}^{N}\left(\left|\sin c\left(\frac{K}{N}(-j+n)\right)\right|^2\right)\times\sum_{j=1}^{N}\left|x_\gamma^{(m)_{i-1}}(j)\right|^2}= \tag{10}$$
$$\frac{N\sum_{j=-N/2+1}^{N/2}\left|\sin c\left(\frac{K}{N}j\right)\right|^2}{\sum_{n=1}^{N}\sum_{j=1}^{N}\left|\sin c\left(\frac{K}{N}(-j+n)\right)\right|^2}$$

where

$$dt_i = -x_\gamma^{(m)_{i-1}}(n) * \frac{K}{N}\operatorname{sinc}\left(\frac{K}{N}\times n\right)\times e^{-j\left(k_1+(K-1)/2\right)\gamma n} \qquad (11)$$

$$n \le N$$

if we evaluate the phrase on the left of the above inequality, which is independent of the transmitted signal, for our simulation parameters N=256 and K=32 we get the value 1.01 or 0.071dB which is too small compared to an OFDM signal that is known to suffer from high PAPR (between 6 to 12dB). From this we can conclude that the PAPR parameter of (8) is not affected by the distortion term and it is almost unchanged for different number of iterations. The diagram in ***Fig. 5*** confirms this conclusion.

V. SUBOPTIMAL SCHEMES A AND B

In this section two suboptimal schemes of [10] are introduced and in section VI they are compared to the proposed scheme. In both of these suboptimal schemes power allocated to subcarriers is inversely proportional to their spectral distance of the PU bands in stepwise manners. In other words less power is assigned to the subcarriers that are near to PU bands. The total transmit power is determined such that the total interference introduced by all the subcarriers is equal to the interference threshold.

A. Scheme A

In this scheme power is distributed such that for the subcarriers that are adjacent to the PU bands the allocated power is P, then the power is increased by P as we move away from the PU bands. Hence to subcarriers that are adjacent to the PU bands the power allocated is P, to those that are right next to them the allocated power is 2P, and so on. In *Fig.* ***6*** the transmitted signal's PSD when power is allocated using scheme A is shown. It is considered that two active PUs are present.

B. Scheme B

In this scheme, the step size of the ladder is taken to be inversely proportional to the total interference that the i[th] subcarrier introduces to all of the PU bands. In other words the power allocated to the i[th] subcarrier using scheme B would be

$$p_i^B = \frac{I_{th}}{M\sum_{j=1}^{J} K_i^j} \qquad (12)$$

where I_{th} is the interference threshold for the PUs, M is the number of subcarriers allocated to SUs and K_i^j is the interference introduced by the i[th] subcarrier to the j[th] PU band. In *Fig.* ***7*** the transmitted signal's PSD when power is allocated using scheme B is shown.

VI. NUMERICAL RESULTE

In this section, simulation results for performance evaluation of the proposed method are presented and compared with the results of the schemes A and B of [10]. We consider an OFDM-based CR system with N=256 and CP=16. The up sampling factor γ is considered 16. Data subcarriers are modulated with a QPSK signal and they are then differential modulated as in (9) for both in-phase and quadrature parts of the signal. All results shown are averaged over 1000 OFDM symbols.

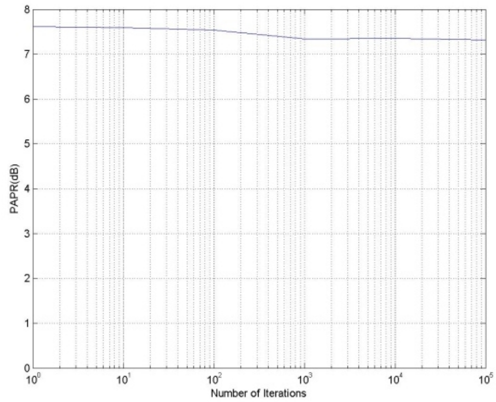

Fig. 5. Transmitted signal's PAPR value for different number of iterations

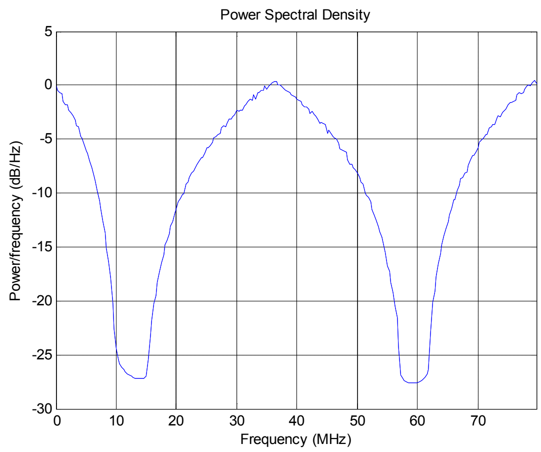

Fig. 6. Transmitted signal's PSD using the power allocation of scheme A in the presence of two PUs

the subcarrier spacing is considered $\Delta f = 0.3125MHz$. It is assumed that two PUs are present at 13.125 and 60 MHz each with the bandwidth of 5MHz (equal to 16 sub-channels). The normalized PSD of the signal at the output of the sidelobe suppression block is shown in *Fig.* ***8***.

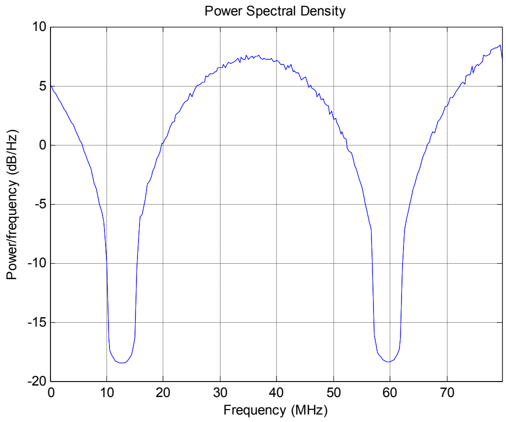

Fig. 7. Transmitted signal's PSD using the power allocation of scheme B in the presence of two PUs

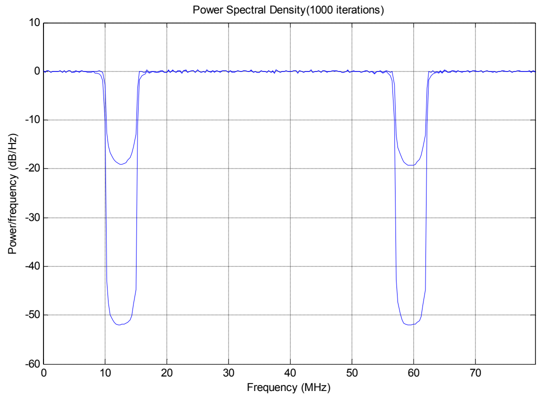

Fig. 8. normalized PSD of the transmitted signal at the output of sidelobe suppression block for L=1000

In *Fig.* ***9*** the CR system throughput rates for the proposed iterative method and the schemes A and B when the Signal to Interference plus Noise power Ratio (SINR) is equal to 13dB are shown. In the proposed algorithm we simply changed the number of iteration for changing the interference level while In schemes A and B, since the power ratio of subcarriers is always the same, we had to change the total power to change the interference level for different points on the diagrams.

As it can be seen in *Fig.* ***9*** When compared to schemes A and B which have the best performance in [10] suboptimal schemes, system throughput of the proposed algorithm is significantly higher. We did not allocate any subcarriers for sidelobe suppression nor we spread the time domain OFDM symbols and nonetheless *Fig.* ***8*** shows that the interference level is reduced appropriately.

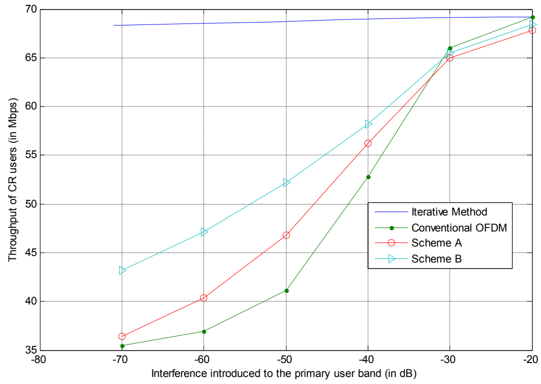

Fig. 9. The transmitted data rates of the CR users

VII. CONCLUSION

In this paper we have developed a simple iterative method to suppress OFDM sidelobes in PU bands and shape the transmitted OFDM signal of the SUs. The proposed iterative method reduces the interference power to the desired level by choosing an appropriate number of iterations. Simulation results show that when this algorithm is accompanied with differential modulation it keeps the system throughput almost unchanged while the similar schemes usually cost in significant reduction in system throughput. We could also show that this algorithm does not change the PAPR value of the transmitted OFDM signal.

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