Cross-band interference reduction trade-offs in SISO and MISO OFDM-based cognitive radios

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***Abstract*—Cognitive radio is a promising approach for efficient utilization of radio spectrum. Due to its high spectral efficiency and flexibility, OFDM is considered as a good signaling scheme for cognitive radios. In this paper, we investigate the problem of cross-band interference minimization in OFDM-based cognitive systems. Cross-band interference is mainly caused by high OFDM sidelobes. In the first part of our work, we propose a framework to study the trade-off between two recently proposed techniques, *adaptive symbol transition* which is performed in the time domain, and *active interference cancellation* which is performed in the frequency domain. We use the trade-off study results to maximize the useful data rate for a desired level of interference. Simulation results show that the best trade-off depends on the configuration of spectral opportunities. In the second part, a new method for interference reduction in multiple-antenna cognitive systems is developed. We show that with knowledge of the channel, the secondary transmitted sequences can be jointly optimized over multiple antennas such that the interference at the primary receiver location is better minimized. Computer simulations demonstrate an improvement of almost 10 dB compared to separate-antenna optimization.**

***Index Terms*—Cognitive radio, OFDM, interference cancellation, MISO-OFDM.**

I. Introduction

THE extensive growth of wireless applications over the past decade has caused an increasing demand for radio spectrum resources. Within the current spectrum regulatory framework, almost all of the available bands have been allocated to existing applications [1], which has resulted in shortage of spectrum. However, actual measurements have shown inefficient spectrum usage as most of licensed spectrum goes unused in a specific location or period of time [2]. Cognitive radio, introduced by J. Mitola [3], is a promising solution to the spectrum shortage problem that suggests using spectrum in an opportunistic manner. That is, cognitive radio devices should be capable of detecting unused spectrum bands and communicate without causing interference to the primary licensed users. To this end, the secondary (unlicensed) users need to use a flexible and efficient signaling scheme. Based on these criteria, some techniques have been proposed as candidates for cognitive radio signaling schemes such as Orthogonal Frequency Division Multiplexing (OFDM), wavelet-based multi-carrier modulation, filter bank multitone communication, and single-carrier frequency division multiple-access (SC-FDMA).

Among these modulation techniques, OFDM has become a popular modulation technique for wireless applications in recent years. This is due to its robustness against multipath fading, high spectral efficiency, and its capacity for dynamic spectrum use. It also has the ability to allocate different power and data rates to distinct subchannels. Thus, OFDM appears to be a good signaling technique candidate for cognitive radio, as it is easy to turn on/off subcarriers in accordance to available sensed spectrum. A more detailed scheme called *spectrum pooling* is introduced in [4].

Despite the aforementioned advantages, however, OFDM inherently suffers from out-of-band radiation due to high sidelobes of subcarriers, causing cross-band interference to other users. Hence, in OFDM-based cognitive systems, turning off the subcarriers that correspond to primary spectrum activity is not enough to mitigate interference to the primary user and other mechanisms should also be taken into consideration.

In this paper, the problem of cross-band interference reduction in OFDM-based cognitive radio systems is considered. We study the problem in two different cases.

A. Single-antenna Cognitive Transmitter

In the first part of our work, we consider the problem of interference minimization in single-antenna transmitter cognitive systems. In this case, the high sidelobes of data subcarriers of a single-antenna secondary transmitter causes interference to the primary users. To suppress the sidelobes, several methods have been investigated in the literature such as time domain windowing [5], [6], [7], subcarrier weighting (SW) [8], and multiple choice sequences (MCS) [9]. Two other novel and efficient methods addressed in the literature are active interference cancellation (AIC) [10] and adaptive symbol transition (AST) [11]. Considering power constraints, an improved version of AIC is also introduced in [12].

In the AIC method, which is performed in the frequency domain, a few subcarriers are inserted at the border of the primary bandwidth. These subcarriers, referred to as cancellation carriers, do not carry data, but are modulated by data dependent complex values such that their sidelobes cancel those of the original transmission signal. The idea is depicted in Fig. 1 [12], where two cancellation carriers are shown to reside at the edge of the primary band. To calculate the complex values of the cancellation carriers, least squares (LS) optimization is used. The main drawback of this method is the loss in throughput since some of the subcarriers no longer convey useful data.

The AST method uses the same approach as the AIC but in the time domain. In the AST method, instead of windowing the signal, each OFDM symbol is extended in the time domain with a complex valued data dependent extension which is calculated to minimize the power level in the primary band.

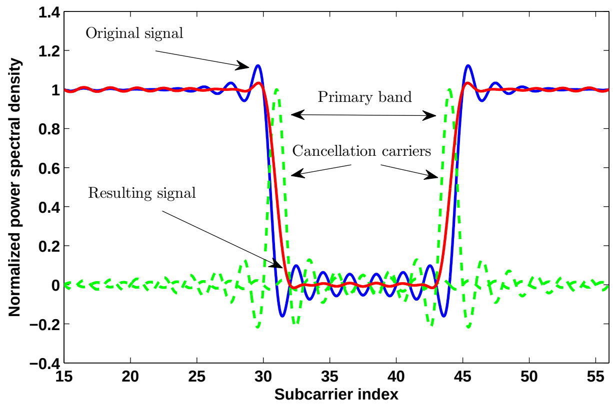

Fig. 1. Using cancellation carriers to reduce the interference power in the primary band.

The idea relies on the fact that the smoother the transition between successive OFDM symbols, the lower the sidelobe levels. The objective is to find the extension vector such that the total interference of the two OFDM symbols and the spectrum of the extension in the primary band cancel each other as much as possible. Similar to AIC, LS optimization is used to find the extension vector in AST. This method reduces interference at the cost of throughput degradation as a portion of time is not used to send useful information.

In all the aforementioned methods, interference cancellation is done without considering the effect of the channel. However, it is important to note that this only works well for low scattering environments, where the channel does not have a serious effect on the spectrum of the transmitted signal. If an OFDM signal is to be transmitted over a frequency-selective fading channel, one can expect that the interference will be better minimized using knowledge of the channel.

The use of channel state information to minimize interference has been studied in a different context for flat fading channels [13], [14], [15]. There, channel state information is used to perform dynamic power control to optimize the transmission rate to secondary user(s), subject to primary interference constraints.

As mentioned, both AIC and AST techniques have analogous complexity and effect on data throughput. The main difference between them is that AIC is performed in the frequency domain while AST is performed in the time domain. As the first part of our work, in this paper, we propose a joint time-frequency scheme in which the interference to the primary user is jointly minimized over the time domain extension and frequency domain cancellation carriers using *channel state information (CSI)*. The objective is to study the *trade-off* between these two methods to find the best trade-off point, that is, the best combination of cancellation carriers and symbol extension for a given amount of interference reduction. In other words, using the trade-off study results, the data rate can be maximized for a desired level of interference reduction.

The contributions of this part are as follows:

- A new joint time/frequency scheme considering knowledge of the channel is proposed to study the time/frequency trade-off in LS based sidelobe suppression methods.
- We show that the time/frequency trade-off between the AIC and AST methods depends on the configuration of spectral opportunities and specifically, whether there is one large primary band, or multiple smaller primary bands.
- Based on the trade-off study results, we show that at the best trade-off point, significant system complexity reduction is possible by an approximation to the least squares optimization.

B. Multiple-antenna Cognitive Transmitter

OFDM can be used in multiple-antenna cognitive systems. This combination results in a multiple-input multiple-output (MIMO) OFDM configuration which has the advantage of higher system capacity [16] and more reliability due to the diversity gain [17] in fading channels.

In multiple-antenna cognitive systems, the total interference to the primary user results from the interference powers caused by each antenna separately. To minimize the interference power, some methods have been addressed in the literature in which the existing sidelobe suppression methods for the single antenna case are extended to the multiple antenna case. Specifically, due to its high performance, AIC has been widely considered by researchers in this context. In [18], an improved AIC is used in which the cancellation carriers are inserted in the transmission symbols of each antenna and the values of cancellation carriers are optimized jointly over all antennas. [19] exploits another improved version of AIC to suppress the interference. In this method, cancellation carriers are transmitted through only one antenna and are designed to cancel the interference resulting from other subcarriers of the same antenna and all subcarriers of other antennas. However, in all these techniques, channel state information is not considered, while it is more important in multiple antenna systems to involve the effect of the channel because the received signal spectrum is the superposition of transmitted signals from each antenna passed through different fading channels.

In the second part of this paper, we consider the problem of interference minimization in multiple-antenna OFDM cognitive systems. Using channel state information, we propose a novel technique, referred to as the *joint antenna* method, to reduce the interference at the location of the primary receiver. Our system consists of a secondary transmitter with multiple antennas sending data to its own receiver, while trying to minimize interference to a primary user. In the joint antenna technique, the streams of OFDM symbols transmitted from the secondary antennas are designed such that the resultant interference *at the primary receiver* is minimized, assuming full channel state information at the secondary. Simulation results show significant improvement of the proposed method of more than 10 dB compared to optimizing over each antenna separately, and/or optimizing without considering effect of the channel.

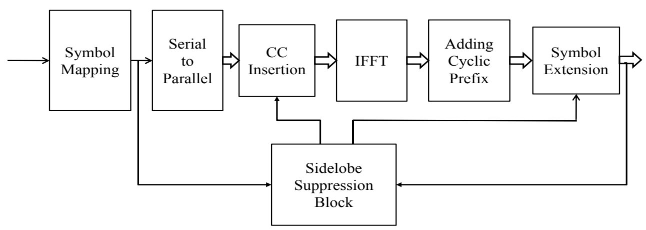

Fig. 2. Cognitive OFDM transmitter block diagram.

The contributions of this part are the following:

- A novel interference reduction technique in multiple-antenna OFDM cognitive systems is proposed.
- We study the time/frequency trade-off in the multiple-antenna case as well.
- Again, based on the trade-off study results, we propose an approximation at the best trade-off point which significantly reduces the system complexity.

The rest of this paper is organized as follows. In Section II, the cognitive OFDM system model is described. The joint time/frequency method for single-antenna case along with simulation results and discussion are presented in Section III. In Section IV we propose the new interference minimization method for multiple-antenna case. This includes the method description and simulation results. Finally, the conclusion is drawn in Section V.

II. COGNITIVE OFDM SYSTEM MODEL

We consider a cognitive radio system in which primary users are detected by a cognitive controller engine. The secondary user should avoid causing interference to the primary. It is assumed that the cognitive system employs OFDM modulation with N subcarriers. The block diagram of the transmitter is depicted in Fig. 2. The input bits are symbol-mapped using a linear modulation scheme such as PSK or QAM. The symbols are then serial to parallel converted resulting in a complex vector to modulate the active subcarriers according to the bandwidth of detected primary user(s). The output of the serial to parallel block is fed into the cancellation carriers insertion block which inserts a few cancellation tones whose amplitudes are calculated by the sidelobe suppression unit to suppress the interference to the primary user. The resulting vector $\mathbf{X} = [X_0, X_1, ..., X_{N-1}]^T$ then passes through the inverse fast Fourier transform (IFFT) module and produces the time domain vector $\hat{\mathbf{x}} = [x_0, x_1, ..., x_{N-1}]^T$ where

$$x_n = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j2\pi kn/N}. \quad (1)$$

We can rewrite (1) in matrix form as $\hat{\mathbf{x}} = \frac{1}{\sqrt{N}} W_{N,N}^{\dagger} \mathbf{X}$, where $W_{N,N}^{\dagger}$ denotes the conjugate transpose of matrix $W_{N,N}$, which is the $N \times N$ discrete Fourier transform (DFT) matrix defined as

$$W_{N,N} = \begin{bmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & w & w^2 & \dots & w^{N-1} \\ 1 & w^2 & w^4 & \dots & w^{2(N-1)} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & w^{N-1} & w^{2(N-1)} & \dots & w^{(N-1)(N-1)} \end{bmatrix},$$

and w is the primitive N^{th} root of unity $e^{-j2\pi/N}$. To avoid intersymbol interference, the cyclic prefix of the OFDM modulated sequence, i.e. the last G samples of the IFFT output, is appended at the beginning of the symbol, where G is assumed to be larger than the maximum delay spread of the channel. To include the cyclic prefix, we define the modified DFT matrix as $W_{N,N+G} = [A \quad W_{N,N}]$, where A is the submatrix of $W_{N,N}$ consisting of the last G columns of $W_{N,N}$. Hence, the time domain OFDM symbol including the cyclic prefix is expressed as

$$\mathbf{x} = \frac{1}{\sqrt{N}} W_{N,N+G}^{\dagger} \mathbf{X}. \quad (2)$$

The extension insertion unit then extends each symbol by optimal extension samples calculated by the sidelobe suppression unit to further mitigate interference to the primary user.

Finally, each OFDM symbol in the time domain is pulse shaped using a pulse shaping filter and sent by the antenna.

Remark: In order to investigate the spectrum of OFDM symbols in-between the subcarrier frequencies, we use an upsampled (by L) discrete Fourier transform (DFT) defined by the $NL \times N$ matrix

$$W_{N,N}^{(L)} = \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & w^{1/L} & \dots & w^{(N-1)/L} \\ 1 & w^{2/L} & \dots & w^{2(N-1)/L} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & w^{(NL-1)/L} & \dots & w^{(NL-1)(N-1)/L} \end{bmatrix}.$$

Hence, the upsampled spectrum of $\mathbf{X}$ is calculated as

$$\mathbf{X}^{(L)} = \frac{1}{N} W_{N,N+G}^{(L)} W_{N,N+G}^{\dagger} \mathbf{X}, \quad (3)$$

where $W_{N,N+G}^{(L)} = [A^{(L)} \quad W_{N,N}^{(L)}]$ is the modified upsampled DFT matrix in which $A^{(L)}$ is the submatrix of $W_{N,N}^{(L)}$ consisting of the last G columns of $W_{N,N}^{(L)}$.

III. SINGLE-ANTENNA COGNITIVE TRANSMITTER: JOINT TIME/FREQUENCY OPTIMIZATION

In this section, the joint time/frequency method for the single-antenna cognitive transmitter is presented. First, we describe the details of the joint method that uses least squares optimization in attempt to minimize interference to the primary user jointly over time and frequency. Using this fact, we employ the joint method to study the trade-off between time and frequency interference reduction. Simulation results and discussion are given afterwards.

A. The joint time/frequency method

As mentioned, both the AIC and AST techniques have approximately the same complexity. Also, they both result in the same approximate decrease in system data throughput, i.e., sacrificing two subcarriers has almost the same impact as extending each OFDM symbol by two samples. By applying the joint optimization, there are two degrees of freedom: the number of subcarriers used as cancellation carriers, and the size of time domain extension. Thus, for a fixed level of interference suppression, there is a tradeoff between the number of tones to be allocated as cancellation carriers and the size of the symbol extension. In other words, for an acceptable loss in data throughput, using the joint technique enables us to minimize the interference, or, for a desired level of interference reduction, data throughput is maximized by allocating the optimal number of cancellation subcarriers in the frequency domain and extension samples in the time domain.

The method is based on jointly minimizing the interference over time and frequency *at the location of primary receiver*, using knowledge of the channel. Namely, in the frequency domain, a number of cancellation carriers on each side of the primary band are used, and in the time domain, a symbol extension is added to each OFDM symbol. Considering the effect of the wireless channel, the weights of the cancellation carriers and the values of the extension are jointly optimized such that the interference to the primary user is minimized.

The secondary user employs the cognitive engine to sense the spectrum. It can also use the received signals from the primary user to accurately estimate the channel between the primary and secondary users. Alternatively, the primary user may be employing coexistence features which provide the secondary users with channel state information as this reduces impact to the primary network. The beacon in IEEE 802.22.1 is an example of coexistence features. Accordingly, here we assume that the secondary transmitter has full knowledge of the channel which is a reasonable assumption (see e.g. [20]).

According to [10], to find the optimum weights for the cancellation carriers of each OFDM symbol, only the spectrum of that symbol is considered during the calculations, while the optimum values of the time domain extensions are found by considering the spectrum of two successive OFDM symbols [11]. To resolve this, we consider a pair of OFDM symbols and their extensions as shown in Fig. 3, assuming that the first symbol has already been well-optimized over time (extension) and frequency (cancellation carriers) to have the least interference to the primary user. The objective is to compute the complex values of the cancellation carriers (denoted by the vector $\boldsymbol{\mu}$) and extension (denoted by the vector $\boldsymbol{\eta}$) of the second symbol.

First, we find the interference to the primary user caused by the OFDM symbol pair, before performing any optimization on the second symbol in the symbol pair. Without loss of generality, we assume that there is a single primary user whose bandwidth is spread over B consecutive subcarriers $[X_{t+1}, X_{t+2}, \ldots, X_{t+B}]$, which are located in the middle of the total available bandwidth of the cognitive radio system, where $B < N$. Depending on the primary bandwidth, a number of subcarriers of the OFDM system are deactivated, or equivalently, corresponding elements in $\mathbf{X}$ are forced to zero. Let $\mathbf{X}_d^{(k)}$ denote the k^{th} OFDM symbol in which tones within the primary band and the cancellation carriers are set to zero, i.e.,

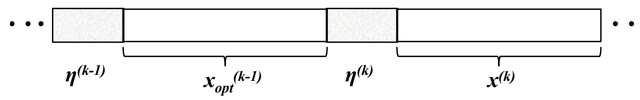

Fig. 3. A pair of OFDM symbols in the time domain in which the first symbol has been optimized.

$$\mathbf{X}_d^{(k)} = [X_0^{(k)}, \ldots, X_{t-g}^{(k)}, 0, \ldots, 0, X_{t+B+g+1}^{(k)}, \ldots, X_{N-1}^{(k)}]^T,$$

where g is the number of subcarriers used as cancellation carriers on each side of the primary band and $\mathbf{X}_{opt}^{(k)}$ denotes the k^{th} OFDM symbol in which the optimum cancellation carries are inserted from the previous round. Also, let $\mathbf{x}_d^{(k)}$ and $\mathbf{x}_{opt}^{(k)}$ be the corresponding time domain symbols, respectively. We denote the upsampled frequency response of the channel between the secondary transmitter and the primary receiver by $\mathbf{h} = [h_0, h_1, \ldots, h_{NL-1}]^T$. Thus, the upsampled spectrum of the non-optimized symbol pair is

$$\mathbf{s} = HW_{N,2(N+G+a)}^{(L)} \boldsymbol{\chi}_d \tag{4}$$

$$= [s_0, s_1, \ldots, s_{NL-1}]^T, \tag{5}$$

where

$$H = \begin{bmatrix} h_0 & 0 & \ldots & 0 \\ 0 & h_1 & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \ldots & h_{NL-1} \end{bmatrix},$$

and

$$\boldsymbol{\chi}_d = \begin{bmatrix} \boldsymbol{\eta}^{(k-1)} \\ \mathbf{x}_{opt}^{(k-1)} \\ \mathbf{0}_a \\ \mathbf{x}_d^{(k)} \end{bmatrix}$$

in which a is the length of the extension and $\mathbf{0}_a$ is the zero vector of length a. $\boldsymbol{\eta}^{(k-1)}$ denotes the optimal extension vector of the $(k-1)^{th}$ symbol calculated in the previous iteration. Hence, the interference vector is

$$\widetilde{\mathbf{s}} = \mathbf{s}^{(t+1)L,(t+B)L}, \tag{6}$$

which is a subvector of $\mathbf{s}$ containing indexed elements $(t+1)L$ through $(t+B)L$ of $\mathbf{s}$. $\|\widetilde{\mathbf{s}}\|^2$ represents the amount of interference power to the primary user and is to be minimized. To this end, the next step is to calculate the contribution of the cancellation carriers and the extension of the second symbol in the primary band.

The upsampled spectrum of the j^{th} unit-weight cancellation carrier is computed as

$$\mathbf{c}_j = \frac{1}{\sqrt{N}} W_{N,2(N+G+a)}^{(L)} \hat{\mathbf{c}}_j, \tag{7}$$

in which

$$\hat{\mathbf{c}}_j = \begin{cases} \begin{bmatrix} \mathbf{0}_{N+G+2a} \\ \frac{1}{\sqrt{N}} W^{\dagger}_{N,N+G} \mathbf{e}_N^{(t-g+j)} \end{bmatrix}, & j = 1, \ldots, g, \\ \begin{bmatrix} \mathbf{0}_{N+G+2a} \\ \frac{1}{\sqrt{N}} W^{\dagger}_{N,N+G} \mathbf{e}_N^{(t+B-g+j)} \end{bmatrix}, & j = g+1, \ldots, 2g, \end{cases}$$

where $\mathbf{e}_N^{(k)}$ is an $N \times 1$ zero vector except the k^{th} entry which is 1. Thus, the spectrum of the j^{th} unit-weight cancellation carrier in the primary band is

$$\widetilde{\mathbf{c}}_j = \mathbf{c}_j^{(t+1)L,(t+B)L}. \tag{8}$$

Similarly, setting the data symbols to zero, the upsampled spectrum of the j^{th} unit-weight sample of the extension is

$$\mathbf{z}_j = \frac{1}{\sqrt{N}} W^{(L)}_{N,2(N+G+a)} \mathbf{e}_{2(N+G+a)}^{(N+G+a-1+j)}, \; j = 1, 2, \ldots, a. \tag{9}$$

Therefore, the contribution of the extension's j^{th} unit-weight sample in the primary band is

$$\widetilde{\mathbf{z}}_j = \mathbf{z}_j^{(t+1)L,(t+B)L}. \tag{10}$$

The cancellation carriers and the extension samples are then weighted by some complex values. These values are jointly optimized such that the interference to the primary user is minimized at the primary receiver location. Letting $C = [\widetilde{\mathbf{c}}_1 \;\; \widetilde{\mathbf{c}}_2 \;\; \ldots \;\; \widetilde{\mathbf{c}}_{2g}]$ and $Z = [\widetilde{\mathbf{z}}_1 \;\; \widetilde{\mathbf{z}}_2 \;\; \ldots \;\; \widetilde{\mathbf{z}}_a]$, we have

$$\begin{aligned} (\boldsymbol{\mu}_{opt}^{(k)}, \boldsymbol{\eta}_{opt}^{(k)}) &= \arg\min_{(\boldsymbol{\mu}, \boldsymbol{\eta})} \|\widetilde{\mathbf{s}} + \widetilde{H} C \boldsymbol{\mu} + \widetilde{H} Z \boldsymbol{\eta}\|^2, \\ \text{s.t.} \quad & |\mu_i|^2 \leq \alpha, \quad i = 1, \ldots, 2g, \\ \text{and} \quad & \|\boldsymbol{\eta}\|^2 \leq p, \end{aligned} \tag{11}$$

where

$$\widetilde{H} = \begin{bmatrix} h_{(t+1)L} & 0 & \ldots & 0 \\ 0 & h_{(t+1)L+1} & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \ldots & h_{(t+B)L} \end{bmatrix},$$

and $\boldsymbol{\eta} = [\eta_1, \eta_2, \ldots, \eta_a]^T$ and $\boldsymbol{\mu} = [\mu_1, \mu_2, \ldots, \mu_{2g}]^T$ are the complex weight vectors of the extension samples and the cancellation carriers respectively. $\alpha = \mathbb{E}\{|X_i|^2\}, i = 1, \ldots, N$, is the power constraint on the cancellation subcarriers where $\mathbb{E}$ represents the expectation operation. This type of power constraint avoids creating overshoot in the resulting signal spectrum. Furthermore, according to [11], by choosing the power constraint on the symbol extension properly, the peak-to-average power ratio (PAPR) of the OFDM signals is not increased. The proper choice for the power constraint is

$$p = a \cdot \frac{E_s}{N+G}, \tag{12}$$

where E_s is the OFDM symbol energy before applying the joint method.

Now, by defining $\mathbf{r} \triangleq [\boldsymbol{\mu}^T \quad \boldsymbol{\eta}^T]^T$ and $D \triangleq [\widetilde{H}C \quad \widetilde{H}Z]$, (11) is simplified to

$$\begin{aligned} \mathbf{r}_{opt} &= \arg\min_{\mathbf{r}} \|\widetilde{\mathbf{s}} + D\mathbf{r}\|^2, \\ \text{s.t.} \quad & |r_i|^2 \leq \alpha, \quad i = 1, \ldots, 2g, \\ \text{and} \quad & \|\tilde{\mathbf{r}}\|^2 \leq p, \end{aligned} \tag{13}$$

where $\tilde{\mathbf{r}} = \mathbf{r}^{2g+1,2g+a}$.

The optimization problem defined in (13) is called a "linear least squares optimization problem with multiple quadratic inequality constraints" which is a well-studied optimization problem. To solve this problem, we first calculate the pseudo inverse of the argument on the right hand side of equation (13) as

$$\mathbf{r} = -(D^{\dagger} D)^{-1} D^{\dagger} \, \widetilde{\mathbf{s}}. \tag{14}$$

If $\mathbf{r}$, which is computed from (14), satisfies the power constraints, then $\mathbf{r} = \mathbf{r}_{opt}$, the optimum solution. If it violates any one of the power constraints, then at least one constraint is tight. In this case, to the best of our knowledge, no analytical solution for solving (13) is known that gives a closed form expression. However, there are efficient solvers that solve the problem iteratively employing numerical algorithms [21]. In this work, to solve (13), we used `cvx`, a package for specifying and solving convex programs [22], [23].

B. *Simulation results and discussion*

Simulations are run to investigate the performance of the proposed joint method. An OFDM-based cognitive radio using $N = 256$ subcarriers is considered where a cyclic prefix of length 64 is added to each symbol. Data subcarriers are modulated with BPSK symbols and the upsampling factor is $L = 16$. The channel between the secondary transmitter and the primary receiver is assumed to be a frequency selective fading channel. The model that we use for the channel is the SUI-4 channel model [24] which is a tapped-delay-line model with 4 taps. In the following simulations, interference power is calculated as the normalized norm of the interference vector in the primary band. We examine the performance of the joint method in two different scenarios.

1) Single wideband interference: In this case, the detected primary user has a rather wide bandwidth which is spread over 32 subcarriers from subcarrier #112 to subcarrier #143. Fig. 4 shows the power spectral density of the output OFDM signal at the location of primary receiver in four different cases. The first case is the conventional OFDM signal spectrum where only the subcarriers in the primary bandwidth are deactivated. The second one is the OFDM signal spectrum using the AST method where the length of symbol extension is 4. In the third case, OFDM signal spectrum using the AIC method with 4 cancellation subcarriers on each side of the primary bandwidth is depicted. Finally, the fourth one is the signal spectrum using the proposed joint technique with 4 cancellation carriers at each side of the primary bandwidth and an extension of length 4. Note that Fig. 4 is not a fair comparison of the performance of the different techniques. Therefore, in order to study the time/frequency trade-off, the amount of interference power

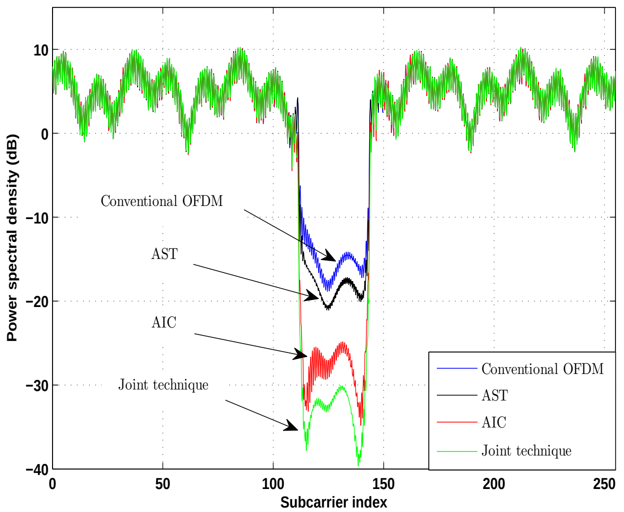

Fig. 4. Single wideband interference, Power spectrum of the output OFDM signal; $N = 256$, number of primary bands=1, $B = 32$.

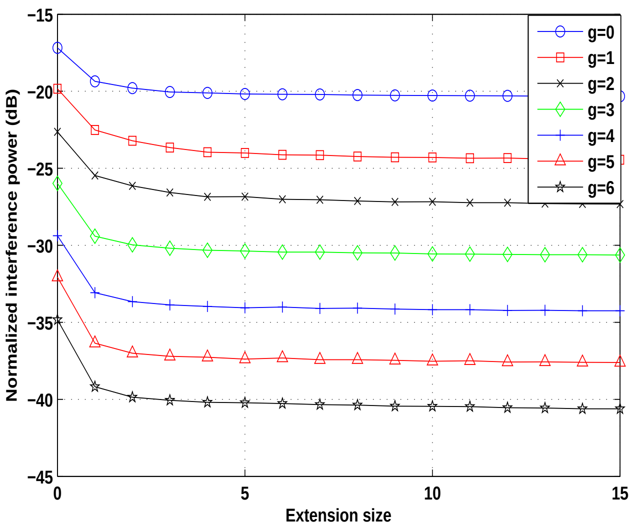

Fig. 5. Effect of adding extension samples on the amount of interference reduction in single wideband interference case.

for different numbers of cancellation carriers and extension lengths is computed. The results are as follows.

Trade-off study: We study the tradeoff between the number of cancellation carriers and the extension size in terms of interference reduction, and design the system to maximize the rate for a fixed interference level. Indeed, we find the best combination of time extensions and cancellation subcarriers to better improve the performance. Fig. 5 depicts the change of interference level for different numbers of cancellation carriers g on each side of the primary band and the number of time extensions. It can be seen from Fig. 5 that there is a dominant break point on each curve in the first extension sample, which implies that while keeping the number of cancellation carriers fixed, the most significant gain is achieved by adding the first extension sample. Furthermore, beyond a single extension, the marginal interference reduction of adding 2 more extension samples is less than that achieved by adding a pair of cancellation carriers. Also, unlike adding extensions, the interference reduction obtained by adding cancellation carriers does not appear to have diminishing returns as each pair of cancellation subcarriers generally reduces the interference by about 4 dB. Because adding 2 extension samples or a pair of cancellation carriers degrades complexity and throughput features approximately equally, the best trade-off is achieved by using only one sample extension and several cancellation carriers for a desired interference reduction level.

Extension approximation: Since the extension is calculated to suppress the sidelobes by smoothing the transition between successive OFDM symbols, instead of solving (13) to find the optimal extension sample, one can compute the average of the two consecutive OFDM symbols endpoints. If the transition between the symbols is approximated by a linear curve fitting, i.e.,

$$\eta_1^{(k)} = \frac{x_{N-1}^{(k-1)} + x_{N-G}^{(k)}}{2}, \tag{15}$$

where $\eta_1^{(k)}$ is the extension sample between the $(k-1)^{th}$ and the k^{th} OFDM symbols, then this approximation significantly reduces the complexity as there is now no need to compute (9) and find the extension via solving (13), i.e., the size of matrices will be reduced and the cancellation carriers can be found separately. The cost of this approximation has been observed in simulations to be small.

2) Multiple narrowband interference: In the second scenario, there are multiple primary bands which are relatively narrow compared to the total available bandwidth of the cognitive system and are used by the same primary receiver. An example of narrow primary bands is IEEE 802.22 (WRAN) standard which is a standard for license-exempt devices to work on a non-interfering basis in the TV Broadcast Service spectrum. A cognitive radio in this band can use up to three consecutive TV channels (18 MHz). Police dispatch devices and wireless microphones which require approximately 200 KHz of bandwidth are considered as narrowband primary users in this band.

In simulations, we assume that the primary bands are spread over six narrow bands whose width are equivalent to 4 subcarriers. Two cancellation subcarriers are inserted on each side of each primary band and the length of the extension is 10. The rest of the parameters are the same as the single wideband case.

Fig. 6 shows the performance of the joint method compared to the conventional OFDM system, the AST method that uses an extension of length 10, and the AIC method where two cancellation subcarriers are inserted on each side of each primary band.

Trade-off study: Similar to the wideband interference case, Fig. 7 depicts the tradeoff between the number of subcarriers g on each side of each primary band and the size of the extension in interference reduction. It can be observed from Fig. 7 that in this case, unlike the single wideband interference case, increasing the size of the symbol extension provides more

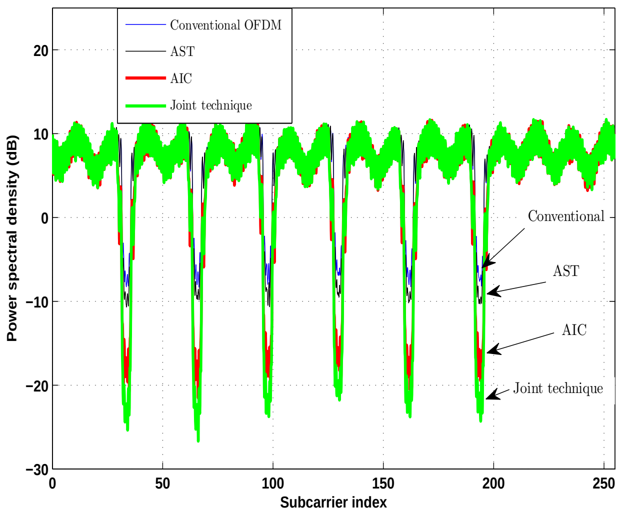

Fig. 6. Multiple narrowband interference: Power spectrum of the output OFDM signal; $N = 256$, # of primary bands $= 6$, $B = 4$.

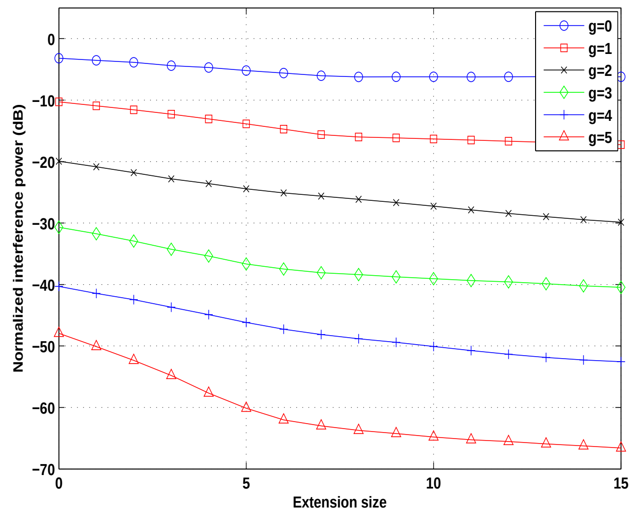

Fig. 7. The effect of adding extension samples on the amount of interference reduction in multiple narrowband interference.

consistent interference suppression, i.e., the marginal return is not negligible after one extension sample. We attribute the better performance of symbol extension to the fact that by sharpening the subcarrier sidelobes, each additional extension sample reduces the interference in all of the narrow primary bands.

Also, note that in this case, each of the narrow primary bands needs cancellation subcarriers at its edge, resulting in $2g{\times}m$ subcarriers where m is the number of primary bands. Hence, adding a pair of cancellation carriers decreases the throughput $2m$ times more than adding an extension sample. Therefore, in terms of data throughput, it can be better to increase the length of symbol extension than to add cancellation subcarriers. For example, in Fig. 7, subcarriers reduce the interference by at most 0.83 dB/subcarrier (10 dB for each addition of 12 subcarriers), whereas the extensions can reduce interference by 1 to 2 dB/extension sample for $g \geq 3$.

As a result, we conclude that the best time/frequency trade-off depends on the configuration of the detected spectral opportunities, whether there is a single large primary band or there are multiple narrow primary bands. In either cases, at the best trade-off point, the joint method achieves a higher interference reduction compared to the pure AIC or AST techniques using the same amount of resources.

IV. Multiple-Antenna Cognitive Transmitter: Joint Antenna Optimization

In this section, we present a new method, called the joint antenna method, for suppressing the interference to the primary user in multiple-antenna OFDM-based cognitive systems based on the idea introduced in Section III. Our proposed method minimizes the interference to the primary user *at the location of the primary receiver*, requiring the knowledge of the channel state information. Moreover, interference minimization is jointly performed over all transmitter antennas. In this case, we will show that a good improvement in interference reduction can be achieved. We use the simulation results to study the time/frequency trade-off as well.

A. The joint antenna method

We assume that the secondary transmitter uses n antennas with sufficient spatial separation, that send streams of OFDM symbols and try to avoid causing interference to a single primary receiver, as shown in Fig. 8. The set of the secondary transmitter antennas and primary receiver antenna forms a multiple-input single-output (MISO) system. Let $\mathbf{h}_i = [h_{i,0}, h_{i,1}, \dots, h_{i,NL-1}]^T$ denote the upsampled frequency response of the channels between the i^{th} secondary transmitter antenna and the primary receiver antenna. Therefore, the upsampled spectrum of the received signal at the primary receiver is

$$\mathbf{Y} = \sum_{i=1}^{n} H_i \mathbf{s}_i, \tag{16}$$

where

$$H_i = \begin{bmatrix} h_{i,0} & 0 & \dots & 0 \\ 0 & h_{i,1} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & h_{i,NL-1} \end{bmatrix}$$

and $\mathbf{s}_i$ is the upsampled signal transmitted by the i^{th} secondary transmitter antenna, $i = 1, 2, \dots, n$.

In the multiple-antenna case, to avoid secondary signals interfering with the primary user, the secondary transmitter forms the transmission OFDM symbols on each antenna in such a way that, after passing through the channels, their effect at the primary band cancel each other as much as possible and the power in the primary band is minimized. To this end, we extend the joint time/frequency optimization technique introduced in Section III for the single-antenna case to multiple

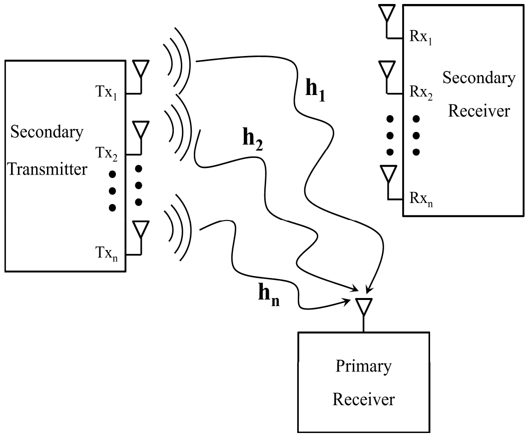

Fig. 8. Multiple-antenna cognitive system.

antennas where the optimization is done jointly, using the channel state information, over multiple antennas. We refer to the proposed method as the *joint antenna* method.

In the joint antenna method, cancellation carriers are inserted in every transmission OFDM symbol of each transmitter antenna in the frequency domain. Each symbol is also extended in the time domain by a symbol extension. The optimal values of the extensions and the cancellation carriers of the n OFDM symbols of the n transmitter antennas are jointly computed considering the effect of the channel, in order to minimize the interference to the primary receiver. Therefore, similar to Section III, an OFDM symbol pair is considered for each transmitter antenna where the first symbol is already optimized.

Let $\widetilde{\mathbf{s}}_i$ denote the upsampled interference vector of the non-optimized symbol pairs of the i^{th} secondary antenna, which is calculated in the same way as in Section III. Thus, the total interference vector at the primary receiver is

$$\widetilde{\mathbf{s}} = \sum_{i=1}^{n} \widetilde{H}_i \widetilde{\mathbf{s}}_i, \tag{17}$$

where

$$\widetilde{H}_i = \begin{bmatrix} h_{i,(t+1)L} & 0 & \dots & 0 \\ 0 & h_{i,(t+1)L+1} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & h_{i,(t+B)L} \end{bmatrix}.$$

We denote the complex values of the second symbol's extension samples and cancellation carriers in the OFDM symbol pair of the i^{th} transmitter antenna by the complex vectors $\boldsymbol{\eta}_i$ and $\boldsymbol{\mu}_i$, respectively. Therefore, the interference contribution of the extension samples of the i^{th} antenna in the primary band at the location of the primary receiver is determined as $\widetilde{H}_i Z \boldsymbol{\eta}_i, i = 1, 2, \dots, n$. Similarly, the interference contribution of the cancellation carriers of the i^{th} antenna in the primary band at the location of the primary receiver is $\widetilde{H}_i C \boldsymbol{\mu}_i, i = 1, 2, \dots, n$, where matrices C and Z are defined in Section III. Thus, the interference minimization problem is expressed as

$$\begin{aligned} &(\boldsymbol{\mu}_{1_{opt}}, \dots, \boldsymbol{\mu}_{n_{opt}}, \boldsymbol{\eta}_{1_{opt}}, \dots, \boldsymbol{\eta}_{n_{opt}}) \\ &= \underset{(\boldsymbol{\mu}_1,\dots,\boldsymbol{\mu}_n,\boldsymbol{\eta}_1,\dots,\boldsymbol{\eta}_n)}{\arg\min} \|\widetilde{\mathbf{s}} + \sum_{i=1}^{n} \widetilde{H}_i (C\boldsymbol{\mu}_i + Z\boldsymbol{\eta}_i)\|^2, \\ &\text{s.t.} \quad |\mu_{ij}|^2 \le \alpha, \quad i = 1, \dots, n, \ j = 1, \dots, 2g, \\ &\text{and} \quad \|\boldsymbol{\eta}_i\|^2 \le p, \quad i = 1, \dots, n, \end{aligned} \tag{18}$$

where $\alpha = \mathbb{E}\{|X_i|^2\}, i = 1, \dots, N$, is the power constraint on each cancellation carrier and p is the power constraint on the symbol extension of each antenna and is chosen according to Section III. By introducing

$$J \triangleq [\widetilde{H}_1 C \ \dots \ \widetilde{H}_n C \ \ \widetilde{H}_1 Z \ \dots \ \widetilde{H}_n Z] \tag{19}$$

and

$$\mathbf{t} \triangleq [\boldsymbol{\mu}_1^T \ \dots \ \boldsymbol{\mu}_n^T \ \ \boldsymbol{\eta}_1^T \ \dots \ \boldsymbol{\eta}_n^T]^T, \tag{20}$$

equation (18) can be expressed as

$$\begin{aligned} \mathbf{t}_{opt} &= \arg\min_{\mathbf{t}} \|\widetilde{\mathbf{s}} + J\mathbf{t}\|^2, \\ \text{s.t.} \quad & |t_j|^2 \le \alpha, \quad j = 1, \dots, 2gn, \\ \text{and} \quad & \|\tilde{\mathbf{t}}_i\|^2 \le p, \quad i = 1, \dots, n, \end{aligned} \tag{21}$$

where $\tilde{\mathbf{t}}_i = \mathbf{t}^{2gn+(i-1)a+1,2gn+ia}$. Equation (21) is a "linear least squares problem with multiple quadratic inequality constraints". The same procedure as in Section III is used to solve this problem.

B. Simulation results and discussion

The MISO case is also examined using numerical simulations. We consider a secondary transmitter with two spatially separate antennas. The model we use for the channel between each secondary transmitter and the primary receiver is the SUI-4 channel model, the same as in Section III. The two channels are assumed to be independent of each other. The OFDM communication system uses $N = 256$ subcarriers per antenna. The detected primary user bandwidth is assumed to occupy 32 subcarriers between subcarrier #112 and subcarrier #143. BPSK modulation is employed to modulate the data subcarriers. A cyclic prefix of length 64 is used and the upsampling factor is $L = 16$.

Fig. 9 shows the spectrum of the OFDM signal at the receiver in four cases. First, the conventional MISO-OFDM spectrum where only the subcarriers at the primary band are deactivated in each transmitted OFDM signal. In the second case, separate antenna optimization, the MISO system is considered as two separate single-input single-output (SISO) systems, where each of the transmitted sequences are separately designed using the optimization method described in Section III, considering the channel state information. The sequences are then passed through the channels and the spectrum of the received signal is computed. The third case is similar to the second case except that it is assumed that the transmitter antennas do not have channel state information, or equivalently, assuming $H_i = I, \ i = 1, \cdots, n$, where I is the identity matrix. Finally, in the fourth scenario, joint antenna optimization is performed where the two transmitted

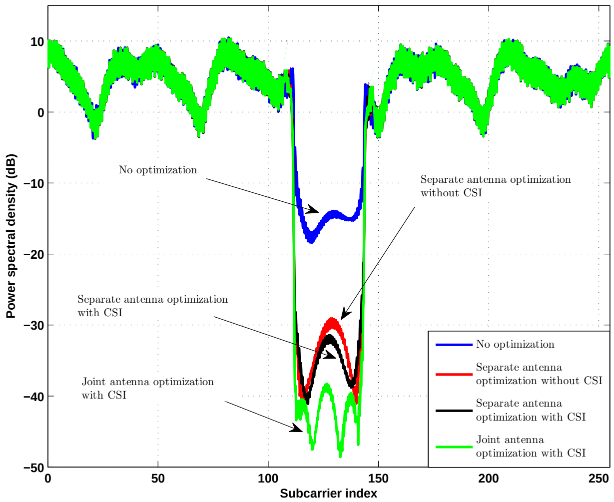

Fig. 9. Comparison of the spectra of MISO-OFDM signal at the primary receiver in the frequency selective fading channel; 4 cancellation carriers on each side of the primary band and a time extension of length 4 are used.

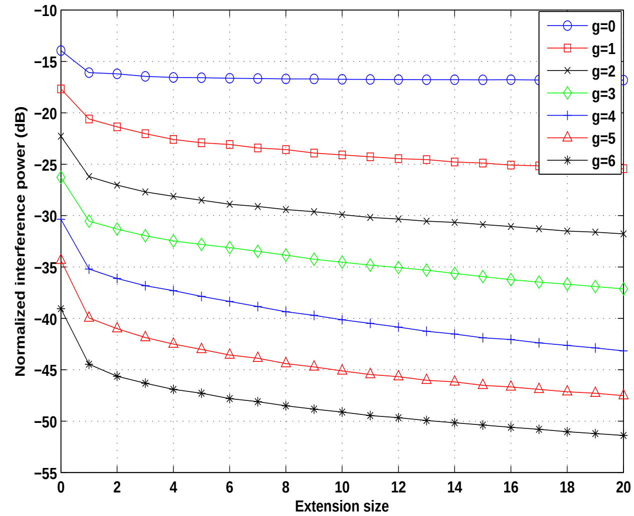

Fig. 10. Effect of adding extension samples on the amount of interference reduction.

sequences are jointly optimized over time, frequency and space (antennas) using the channel information. As Fig. 9 shows, the joint antenna optimization method suppresses the interference by almost 10 dB more than the separate antenna optimization method, and thus, channel state information at the transmitter can provide significant improvement for interference reduction in multiple-antenna systems.

Trade-off study: In Fig. 10, the trade-off between the number of cancellation carriers and the extension size with respect to the amount of interference reduction is depicted. Similar to the single-antenna case, it can be seen from Fig. 10 that adding the first extension sample gives the most significant gain in interference suppression.

Extension approximation: We can conclude from Fig. 10 the same result of the single-antenna case that at the best trade-off, instead of calculating the extension sample using the optimization problem stated in (18), we can solve the optimization problem only for the cancellation carriers and use a single sample for the extension. The extension sample can be easily approximated as the average of the two endpoints of the two consecutive OFDM symbols, i.e.,

$$\eta_{i1}^{(k)} = \frac{x_{i,N-1}^{(k-1)} + x_{i,N-G}^{(k)}}{2}, \quad i = 1, 2, \ldots, n, \qquad (22)$$

where $\eta_{i1}^{(k)}$ is the extension sample between the $(k-1)^{th}$ and the k^{th} OFDM symbols of the i^{th} antenna.

Finally, simulation results for the multiple narrowband primary user case also demonstrate that, similar to the single-antenna case, interference reduction per extension sample in dB/sample for the first few extensions is greater than the reduction per subcarrier in dB/subcarrier. Therefore, the trade-off study in the multiple-antenna secondary transmitter case has the same result as the single-antenna case.

V. Conclusion

In this paper, the problem of interference reduction in OFDM-based cognitive radios in single-antenna and multiple-antenna secondary transmitter for single primary user band and multiple primary user bands is considered. In the single-antenna case, we propose a new joint time/frequency scheme to investigate the trade-off between active interference cancellation and adaptive symbol transition techniques. The new method optimizes jointly over the symbol extension and cancellation subcarriers to minimize the interference to the primary user. In view of symbol extension, it is shown that for a single wideband primary, most of the gain in interference cancellation is achieved by adding the first extension sample. Hence, the complexity can be significantly reduced by using one extension sample whose value is easily calculated as the average of the two endpoint of two successive OFDM symbols. Furthermore, we show that the effect of the channel on the transmitted secondary signals can be used to improve interference cancellation. Using this fact, in the multiple-antenna case, we propose a new method, called the joint antenna method, in which the transmitted sequences from the secondary transmitter antennas are designed such that the interference at the primary receiver antenna is minimized. Simulation results also demonstrate significant improvement in jointly optimizing over two antennas compared to two separate antenna interference minimization.

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