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Mitigation of spectral leakage for single carrier, block-processing cognitive radio receivers

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ABSTRACT

Cognitive Radio (CR) is a multiuser, wireless communication concept that allows for a dynamic and adaptive assignment of spectral resources. The coexistence of multiple users, often transmitting at significantly different power levels, makes CR receivers vulnerable to spectral leakage caused by components' nonlinearities and time-truncation of the processed signal records. In this work we propose a method for mitigating the latter with an adaptive choice of the length of the processing block size. With simulations we show that a significant leakage reduction that leads to receiver dynamic range improvement of around 10 dB can be achieved with the proposed method.

1. Introduction

Cognitive Radio (CR) is a wireless communication concept that aims for an efficient use of spectral resources through dynamic spectrum management [1]. A conventional wireless receiver uses a variable-frequency oscillator and a filter to tune the desired signal to a common intermediate frequency or baseband, where it is then sampled and quantized by an Analog-to-Digital Converter (ADC). Unlike conventional wireless transceivers, operating in preallocated sub-bands, CRs need to support any momentarily unoccupied sub-bands in a wide frequency range of interest. As an example of a practical application of CR, consider the recently developed IEEE 802.22 standard [2] that was aimed at using CR for opportunistic transmissions in a 50–700 MHz frequency band that became sparsely allocated [3] after the switchover to digital television in the United States in June 2009. An ideal architecture for an opportunistic cognitive radio receiver would be a wideband Low-Noise-Amplifier (LNA) and an analog-to-digital converter directly following the receiving antenna. A digital processor would then process the output of the ADC in order to extract information from a dynamically assigned channel of interest. Such an architecture, while appealing with its flexibility, poses significant implementation challenges. The CR wideband RF antennas receive signals from various transmissions, often at significantly different power levels. Since the interfering transmissions cannot be removed by analog filtering, the CR RF front-end needs to have the capability of receiving a weak signal of interest in the presence of a very strong interferer. Two main reasons for the distortion of the weak message signal in strong interference, and therefore the degradation of the receiver's dynamic range, are nonlinearities in the receiver's components and time-truncation of the blocks of processed signal records. In this work we consider the latter cause of degradation and propose a method for adaptation of the processing window size for a dynamic range improvement.

Multiple techniques have been considered in the past for dynamic range enhancement for cognitive radio. In [4], the authors proposed spatial filtering techniques for multiple antenna CR systems. In [5] Yang et al. considered a multi-stage receiver architecture, in which the interference was estimated and subtracted in the initial stage of the receiver. In [6], exclusion of parts of the spectrum classified as quasi-stationary was proposed. The time-domain limitation of signal processing blocks and effects of windowing have been studied in [7].

Block transmissions, for which groups of data symbols are processed as a unit, allow for the implementation of Frequency Domain channel Equalization (FDE), which for broadband transmissions in rich multipath environments can bring significant complexity relaxations when compared to time-domain equalization [8]. To allow for FDE, block transmissions employ a Cyclic Prefix (CP). The CP is a copy of the end of the signal block attached in front of the block. To avoid Inter-Block Interference (IBI), the length of the CP is chosen larger than the maximum delay spread of the channel. Due to the dynamic character of wireless channels, the value of the delay spread changes over time and space [9]. For the high reliability demanded of modern communication systems, the tail of the distribution of the delay spread dictates a

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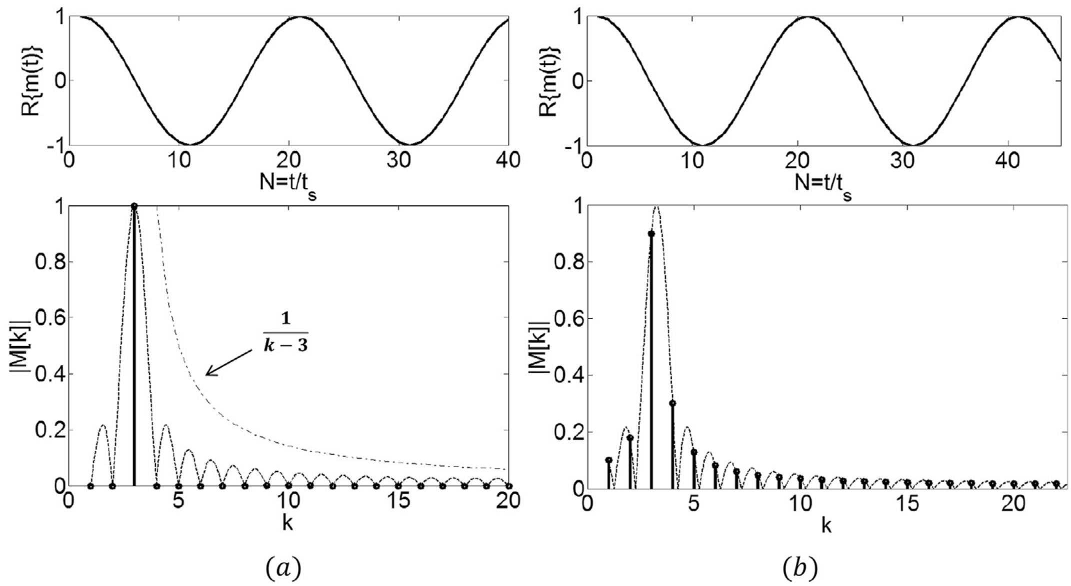

Fig. 1. Spectral representation of a sampled complex sinusoid with the DFT. Observation time is such that: (a) the frequency of the oscillation overlaps with one of the DFT grid points; (b) the frequency lies off the grid which leads to the spectral leakage.

conservative choice of the length of the cyclic prefix, which is fixed during the design stage.

Since truncation in the time domain causes spreading in the frequency domain, weak signals in a block transmission can be corrupted by strong in-band interferers (see Section 2). In this work we propose enhancement of the dynamic range of a block processing CR receiver by an adaptive choice of the processing block size to minimize the spectral leakage into the frequency sub-bands occupied by the signal of interest. In particular we will show that with an adaptive non-complete removal of the CP at the receiver, which can be employed in wireless environments with channel delay spreads even slightly shorter than the length of the CP, significant receiver dynamic range improvements can be achieved for block transmissions. The proposed method can be used together with the previously researched approaches, listed above, in order to further improve the receiver's dynamic range. It is important to stress that while the receiver adaptively changes the size of the processing block, no changes are applied to the transmission scheme. In particular, the transmitter block size is not subject to the adaptive change.

2. Problem statement

The Discrete Fourier Transform (DFT) is an invertible signal processing operation that projects a time-limited signal onto a set of complex frequencies, and gives a discrete representation of the signal's spectrum. The values of the frequencies that the signal is projected onto build a discrete grid, equidistantly dividing the entire sampling bandwidth. A complex sinusoid oscillating with a frequency off the discrete frequency grid cannot be represented with a single element of the grid, and its energy leaks between multiple elements, which leads to misinterpretation of its spectral content. This is visualized in Fig. 1, where the observation time of a complex oscillation is shown for two different cases: (a) the frequency of the sinusoid overlaps with that of one of the grid points (left subplots); (b) the frequency of the sinusoid lies off the grid which leads to spectral leakage (right subplots). For wide-band, block-processing receivers, the finite block length can cause misinterpretation of the Fourier coefficients of the in-band interference, possibly orders of magnitude stronger than the message of interest, which can lead to significant contamination of the message.

A cyclic prefix is employed in the Single-Carrier Frequency Division Multiplexing (SC-FDM) scheme [10], which is often considered for broadband transmissions over wireless channels. For example SC-FDM has been selected as an uplink communication scheme for the Long Term Evolution (LTE) standard for wireless, high-speed data communication for mobile phones and data terminals [11]. Similar to Orthogonal Frequency Division Multiplexing (OFDM), SC-FDM processes data in blocks; however, unlike OFDM, it utilizes a single carrier modulation at

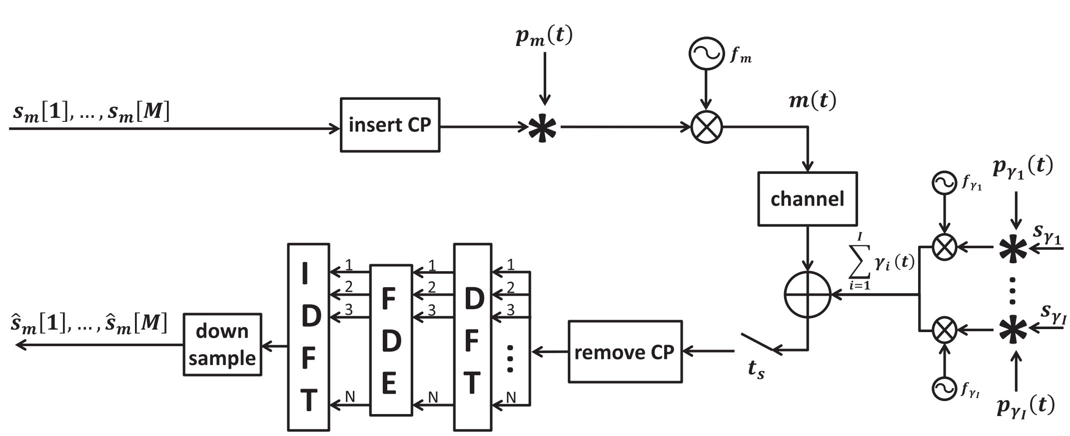

Fig. 2. Block diagram for a single carrier frequency division multiplexing cognitive radio transmission subject to interference.

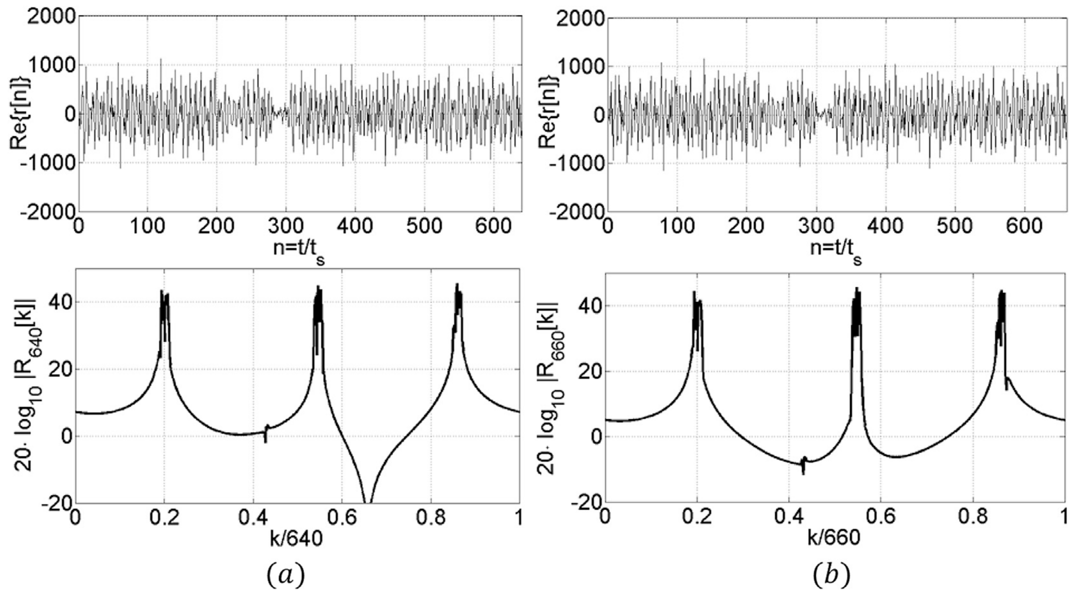

Fig. 3. Sampled time domain (top) and DFT (bottom) capture ($r[n]$ and $R_N[k]$) of a QPSK block transmission of 5 symbols of interest equipped with 1 cyclic prefix symbol, transmitted at a symbol rate W_m=3.84 MHz and power P_m for two different choices of the processing block size: (a) N=640 after a complete CP removal, (b) N=660 after a partial CP removal. The signal of interest, located at $f_m/f_s \approx 0.4297$ was contaminated with three interferers occupying 7.68 MHz sub-bands, with total power P_γ. The signal-to-interference ratio and sampling frequency were set to, respectively: $SIR = 10\log_{10}\frac{P_m}{P_\gamma} = -60\ dB$ and f_s=491.52 MHz.

the transmitter, which allows for reduction of the Peak-to-Average Power Ratio (PAPR) [10] and thus improves the efficiency of RF power amplifiers. The inherent DFT operation at the receiver (Fig. 2) makes SC-FDM suitable for spectral sensing applications of cognitive radio. If the length of the CP is longer than the longest delay path of the channel, then the received blocks arriving at the receiver over different channel paths appear as circularly shifted versions of the transmitted block. A circular shift of a discrete-time record corresponds to a multiplication of its DFT with a linear phase. Therefore FDE can be implemented as a simple, frequency domain multiplication with an inverse of the estimate of the channel transfer function, in contrast to time domain equalization, which can involve time-domain adaptive filters with tens of taps and hundreds of multiplication operations required per data symbol.

Consider a single-carrier, block transmission and a wide-band, highly over-sampling, cognitive radio receiver are visualized in Fig. 2. Message data symbols s_m after cyclic prefix insertion are pulse-shaped with pulse $p_m(t)$, up-converted to frequency f_m, and sent over the channel. The received signal, including an additive interference $\sum_{i=1}^{I}\gamma_i(t)$, possibly with power orders of magnitude higher than that of the message, is over-sampled with a sampling rate $f_s = 1/t_s$. A block of high-rate samples is captured, the cyclic prefix is discarded, and the DFT of the sample sequence of length N is calculated for frequency domain channel equalization. The transmitted data is then recovered from the time representation of the equalized and digitally filtered parts of the spectrum. In the reminder of this paper we refer to N as the size of the processing block after (potentially partial) cyclic prefix removal.

Consider an off-grid complex interferer sinusoid with unit amplitude and a frequency f_γ such that: $f_\gamma \cdot t_s \cdot N = l + a; l \in \mathbb{N}; -0.5 < a \leq 0.5$. The DFT of the sinusoid is ((5.103) [12], pg. 262)

$$\mathbb{O}(f_\gamma, k) = \frac{\sin\left(\pi\left(f_\gamma \cdot t_s \cdot N - k\right)\right)}{\sin\left(\frac{\pi}{N}\left(f_\gamma \cdot t_s \cdot N - k\right)\right)} \cdot e^{j\pi\left(f_\gamma \cdot t_s \cdot N - k\right)\frac{N-1}{N}}, \quad k = 1, \ldots, N. \tag{1}$$

For an interferer $\gamma(t)$ with a continuous bandwidth $(f_\gamma - W_\gamma/2, f_\gamma + W_\gamma/2)$, spectral power leakage into a set K_m of the discrete frequencies occupied by the message of interest (either a message to be sensed or to be received) is therefore

$$L_N[K_m] = \sum_{k \in K_m} \left| \int_{f_\gamma - W_\gamma/2}^{f_\gamma + W_\gamma/2} \Gamma_\gamma(f) \mathbb{O}(f, k) df \right|^2 \tag{2}$$

where $\Gamma_\gamma(f)$ are the values of the continuous Fourier transform of the interferer. If the power of the interferer is orders of magnitude higher than the power of the message of interest, the leakage can cause significant degradation of the message's quality. The level of the leakage into a particular frequency band of interest differs for different choices of the size of the processing block N. This work addresses a method for mitigation of the interference power leaking into the band of interest via an adaptive choice of N obtained by partial removal of the cyclic prefix at the receiver.

3. Proposed method for receiver dynamic range enhancement for block transmissions

For conventional block receivers, the cyclic prefix is removed completely before the block is processed to retrieve the transmitted message of interest. The complete CP removal allows for the cancelation of IBI caused by the channel delay spread τ_{CH} not exceeding the length L_{CP} of the cyclic prefix and leads to a fixed processing block size, further referred to as N_B. As we saw in the Section 2 (Fig. 1 and eq. (1)) the character of the DFT leakage depends on the size N of the processing block. In other words, the spectral representation of the frequency components comprising the signal changes with the change of N. If the delay spread of a considered channel is smaller than the length of the CP employed, then without compromising the ability to cancel the IBI completely, a potential leakage reduction and therefore dynamic range enhancement can be achieved with an adaptive, partial removal of the cyclic prefix. In particular, an adaptive choice of the length N of the processing block can be made from a search set $\boldsymbol{N}_{search} = \left\{N_B,\ N_B + 1,\ \ldots,\ N_B + \frac{L_{CP} - \tau_{CH}}{t_s}\right\}$, where t_s is the sampling period. The current value of τ_{CH} that determines the size of $\boldsymbol{N}_{search}$ can be estimated at the receiver using conventional channel order estimation methods in OFDM systems as studied for example in [13–15].

To gain an intuition on how the spectral leakage can be controlled with the size N of the observation window, consider an interferer γ

consisting of a set of I complex oscillations with amplitudes Γ_{γ_i} and frequencies f_{γ_i}, i=1,…,I, and a message m consisting of a set of M complex oscillations with frequencies f_{m_μ}, $\mu = 1, \ldots, M$. Depending on the size N of the observation window, f_{m_μ}, $\mu = 1, \ldots, M$, can either lie on or off the discrete frequency grid. Denote the rounded value of x to the nearest integer as $[\![x]\!]$. The discrete frequency set K_m of the message is thus $K_m = \left\{ [\![\frac{f_{m_1}}{f_s}N]\!], \ldots, [\![\frac{f_{m_M}}{f_s}N]\!] \right\}$. Ignoring the rounding operation when building the set K_m, defining $\Delta_{i,\mu} = \frac{(f_{\gamma_i} - f_{m_\mu})}{f_s}$, with (1) and (2), the total interference experienced by the message due to the leakage caused by the limited observation window size N is

$$L_N[K_m] = \sum_{\mu=1}^{M} \left| \sum_{i=1}^{I} \Gamma_{\gamma_i} \frac{\sin(\pi\Delta_{i,\mu}\cdot N)}{\sin(\pi\Delta_{i,\mu})} e^{j\pi\Delta_{i,\mu}\cdot N} \right|^2 = \sum_{\mu=1}^{M} \left| \sum_{i=1}^{I} \Gamma_{\gamma_i} \frac{1}{2j\cdot\sin(\pi\cdot\Delta_{i,\mu})} \cdot \left(e^{2j\pi\Delta_{i,\mu}\cdot N} - 1\right) \right|^2. \quad (3)$$

Denote $\tilde{\Gamma}_{i,\mu} = \Gamma_{\gamma_i} \cdot \frac{1}{2j\cdot\sin(\pi\cdot\Delta_{i,\mu})}$. The minimization of the interference leakage experienced by the message over the choice of the window size N simplifies to

$$L_{min}[K_m] = \min_{N \in \mathbf{N}_{search}} \sum_{\mu=1}^{M} \left| \sum_{i=1}^{I} \tilde{\Gamma}_{i,\mu} \cdot \left(e^{2j\pi\Delta_{i,\mu}\cdot N} - 1\right) \right|^2 \quad (4)$$

which is a function of $\tilde{\Gamma}_{i,\mu}$'s that depend on the signal transmitted by the interferers, weighted with factors that depend only on the frequency spacing between the message and the interferers. Eq. (4) shows how the interference leakage can be controlled with the choice of the value of the length N of the processing block.

Fig. 3 visualizes possible reduction of the leakage of interferers' power into the frequency band occupied by the message of interest that can be achieved with the partial removal of the CP and hence an adjustment of the processing window size from 640 to 660. While increasing the spectral leakage to some parts of the spectrum, the adjustment leads to a roughly 10 dB reduction of the leakage into frequency bins around the message carrier frequency located at $f_m/f_s \approx 0.4297$.

In practice the receiver does not have access to $L_N[K_m]$ (2), (3); thus, some other measurable quantity must be used to determine the optimal N. We will employ the total received power in the frequency band occupied by the message of interest. To establish its utility we show that the power of the message of interest and the interference decorrelate quickly with the size N of the processing block. Consider a received signal $r(t)$ consisting of the message of interest $m(t)$ with power P_m and an interferer $\gamma(t)$ with power P_γ. The baseband sampled received signal can be written as $r[n] = m[n] + \gamma[n]$. Its power is

$$P_r = \frac{1}{N} \sum_{n=1}^{N} |r[n]|^2 = P_m + P_\gamma + 2C \quad (5)$$

$C = \frac{1}{N}\sum_{n=1}^{N} \mathscr{R}\{m[n]\}\cdot\mathscr{R}\{\gamma[n]\} + \mathscr{I}\{m[n]\}\cdot\mathscr{I}\{\gamma[n]\}$. Let $m[n]$ and $\gamma[n]$ be respective digital streams of symbols s_m and s_γ, with periods T_m and T_γ, pulse-shaped with analog pulses $p_m(t)$ and $p_\gamma(t)$, sampled with frequency $f_s = 1/t_s$:

$$m[n] = \sum_{l=-\infty}^{\infty} s_m[l]\cdot p_m(nt_s - lT_m), \quad \gamma[n] = \sum_{l=-\infty}^{\infty} s_\gamma[l]\cdot p_\gamma(nt_s - lT_\gamma). \quad (6)$$

Assume $s_m[l]$ and $s_\gamma[l]$ to be uncorrelated, zero-mean random variables, which is the case for commonly used digital modulation schemes. Then $\mathbf{E}\{C\} = 0$ and

$$Var\{C\} = \frac{2\cdot Var\{\mathscr{R}\{s_m\}\}\cdot Var\{\mathscr{R}\{s_\gamma\}\}}{N^2} \times \sum_{n=1}^{N} \sum_{l_1=-\infty}^{\infty} \sum_{l_2=-\infty}^{\infty} p_m^2(nt_s - l_1T_m)\cdot p_\gamma^2(nt_s - l_2T_\gamma). \quad (7)$$

Eq. (7) demonstrates that C is small relative to P_γ. For example, even when $\mathscr{R}\{s_m\}$ and $\mathscr{R}\{s_\gamma\}$ have unit variance (note the case of most interest is when $\mathscr{R}\{s_m\}$ is much larger), the ratio of $Var\{C\}$ and $Var\{\mathscr{R}\{s_\gamma\}$ is -17 dB for a very short processing block of 32 symbols, pulse-shaped with a raised cosine filter with a roll-off factor of 0.5, and oversampled at a rate of 32. Based on this empirical evidence, the power of the block of the received signal $r[n]$ concentrated in the frequency range occupied by the message of interest, which is easily measured, will be used to make the decisions on the size of the processing block. Numerical results will show this as an effective choice. Denote as $\mathbf{R}_N$ the DFT of length N of the received, sampled signal $r[n]$, $n = 1, \ldots, N$, carrying a message of interest, occupying a bandwidth $(f_m - W_m/2, f_m + W_m/2)$. The optimal size of the processing block can then be chosen as

$$N_{opt} = \underset{N \in \mathbf{N}_{search}}{\operatorname{argmin}} \sum_{k_m = [\![\frac{f_m - W_m/2}{f_s}N]\!]}^{[\![\frac{f_m + W_m/2}{f_s}N]\!]} |\mathbf{R}_N[k_m]|^2. \quad (8)$$

4. Numerical results

In this section we numerically study possible dynamic range improvements for the block CR receivers, employing adaptive, partial removal of the CP, and hence with the adaptive, block-to-block choice of the processing block size. Consider a CR receiver working with a sampling rate f_s=491.52 MHz receiving a transmission of interest together with strong interference. The transmission of interest $m(t)$, at a carrier frequency f_m, is a QPSK transmission with a symbol rate of W_m=3.84 MHz, pulse-shaped with a raised-cosine filter $p_m(t)$ with a roll-off factor $\beta = 0.5$. The interferer signal consists of three QPSK transmissions at carrier frequencies f_{γ_i}, with symbol rates W_{γ_i}, also pulse-shaped with $\beta = 0.5$ raised-cosine filters $p_{\gamma_i}(t)$, $i = 1, 2, 3$. It is important to stress here that the method's performance did not change for various settings of the interferers' excess bandwidth β. The processing block was built out of 32 QPSK message symbols and was equipped with a cyclic prefix of length $6.25\%\cdot 32 = 2$ symbols, which allows the avoidance of IBI caused by a channel with a maximal delay path difference $\Delta d_{CH_1} = c\cdot\tau_{CH_1} = c\cdot\frac{2}{3.84\ MHz} = 156.14$ m, where c is the speed of light. The number of samples building the processing block at the receiver after the complete CP removal was then $N_B = [\![\frac{32}{3.84\ MHz}\cdot f_s]\!] = 4096$.

We assume that the transmission occurs in a wireless environment with the maximal channel delay path difference Δd_{CH_2} slightly shorter than Δd_{CH_1}. In particular, a 6-tap channel with taps at {0.0039, 0.1602, 0.3164, 0.4727, 0.6289, 0.7852} $[L_{CP}]$ was assumed. The complex values of taps were chosen uniformly at random and multiplied with a real-valued decay factor $e^{-a\cdot t}$, where $a = 10^7$. In such an environment, a complete IBI cancelation is achieved even if a fraction of the CP (up to $\tau_{CH_1} - \tau_{CH_2}$ seconds) is not discarded and kept for further processing. This allows for a processing block size search set: $\mathbf{N}_{search} = \{N_B, \ldots, N_{max}\}$, with $N_{max} = 4096 + (\tau_{CH_1} - \tau_{CH_2})\cdot f_s$.

For the considered transmission, for $N_{max} = 4134$ corresponding to a search space of size 15% of the CP length, the upper subplot of Fig. 4 shows the Root Mean Square Error (RMSE) of the recovered QPSK message symbols as a function of the Signal-to-Interference Ratio (SIR), which is the ratio of the power of the signal of interest to the power of the interference, before and after the adjustment of N with (10). Fig. 4,

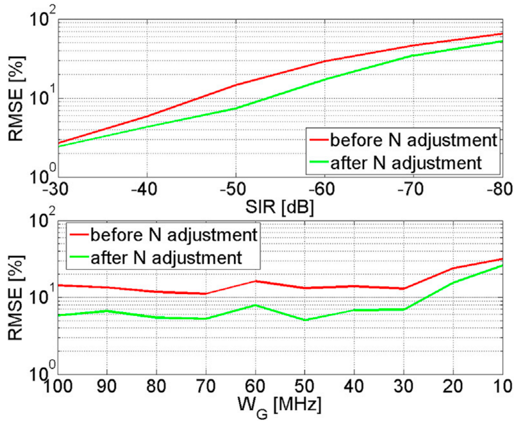

Fig. 4. RMSE of the recovered QPSK message symbols before and after the adjustment of the processing block size N with (10) as a function of the signal-to-interference ratio for a fixed guard bandwidth from (11) W_G=50 MHz (top subplot) and as a function of the guard bandwidth W_G for a fixed SIR=−50 dB (bottom subplot).

averaged over 500 random choices of message and interferers' symbols s_m and s_{γ_i}, carrier frequencies f_m and f_{γ_i} and interferers' symbol rates $W_{\gamma_i} \in \{1.92, 3.84, 7.68, 15.36, 30.72, 61.44\}$ MHz, shows that for a desired value of RMSE a significant reduction of required SIR value, and hence dynamic range improvement of roughly 10 dB, can be achieved with this simple adjustment of the block processing length N, as similar RMSE values are achieved at − 60 dB when the proposed method is employed and at − 50 dB when it is not. The random choice of f_m, f_{γ_i} and W_{γ_i} was subject to a constraint

$$|f_m - f_{\gamma_i}| < W_G + W_{\gamma_i}/2, \quad i = 1,2,3 \tag{9}$$

where W_G is a guard bandwidth set to 50 MHz, that dictates the minimal spacing between the message of interest and interference carriers. The lower subplot of Fig. 4 for $N_{max} = 4134$ and $SIR = -50\ dB$ shows the RMSE of the recovered QPSK message symbols as a function of the guard bandwidth W_G. The effect of the spectral leakage decreases with increasing SIR, and therefore the room for impact for the proposed leakage reduction technique (or any other) is limited at high SIRs. For fixed SIR the performance of the proposed method decreases with decreasing W_G. Small W_G can lead to small values of $\Delta_{i,\mu}$ from (4), which limits the rotation speed of $(e^{2j\pi\Delta_{i,\mu}\cdot N} - 1)$.

5. Conclusions

In this paper an approach for the mitigation of spectral leakage via an adaptive choice of processing window size was proposed for block processing, single-carrier CR receivers. The method is based on an adaptive partial removal of the CP and can be applied in environments with maximal channel delay paths shorter than the length of the CP. The proposed method does not require any structural changes to the receiver and allows for significant sensitivity improvements (around 10 dB) when compared to a studied fixed window size approach from [7]. The receivers' sensitivity improvements translate directly into a meaningful reduction of the cost of deployment of cellular wireless networks, as they lead to a lower spatial density of base stations required to provide a certain level of coverage. Among areas for future work are verification field tests in a wide range of wireless environments with different wireless channel properties.

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