A study of OFDM signal detection using cyclostationarity

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***Abstract*— In a Spectrum Pooling context, *Cognitive Radio* nodes have to perform spectrum sensing in order to find available spectrum holes. In this paper, OFDM cyclostationarity is studied as the best detector choice to find white spaces in the spectrum. OFDM signal detection can be performed using the induced cyclostationarity of the cyclic prefix. On the other hand, OFDM signal identification can be performed through the use of pilot patterns. These two cyclostationary based detectors are tested in different channel conditions using simulations. In addition, some unsolved issues of the OFDM detection context are discussed.**

***Keywords*— Spectrum Sensing, Cognitive Radio, Cyclostationarity, OFDM.**

1. INTRODUCTION

The growing demand of ubiquitous access to internet and multimedia services over wireless terminals represents an interesting challenge for researchers. One of the issues to deal with is the large amount of required bandwidth to deliver such services. Diverse studies made in several cities (Europe and US) showed that some bands currently occupied by license owners are mostly underused. In our context, this is a waste of a very scarce resource, the spectrum. In other words, these studies proved that conventional policies of spectrum licensing are inefficient.

In Mitola's work [1], a new policy of spectrum usage called *Spectrum Pooling* (SP) was proposed. With SP public access to a licensed band is enabled. Licensed systems, also called Primary System (PS), which do not use its bandwidth all time are able to rent part of it to a committing system, or Secondary System (SS), with the condition that performance of PS is not affected. PS systems do not take part in the process of spectrum pooling because the policy aims to share spectrum in already occupied bands with already installed equipments. All the work is done by sophisticated nodes called *Cognitive Radios*, which are capable of sensing the spectrum in order to find white spaces and accordingly reconfigure their physical parameters to occupy the available resources. As a result, SS obtains access to new bands and licensed owners get revenues for the spectrum they do not use [2].

Spectrum pooling is an interesting approach to enhance spectral efficiency. As a consequence, it seems to be a good solution to spectrum scarcity. The IEEE 802.22 Working Group is setting up a standard that allows Cognitive Radios to exploit unused TV bands [3].

The main tasks an SS detector must perform are *spectrum sensing* and *reconfigurability*. The former consists of sensing the spectrum in order to find the unused bands. This information (available bands) is commonly referred to as channel allocation information (CAI). Since the availability of resources is highly varying, the sensing must be performed periodically. In addition, reliable detection of licensed users is crucial because missing detection leads to increasing interference from SS users over PS users. Cognitive nodes use spectrum sensing information to dynamically modify their physical parameters to not interfere with licensed users. This property, known as reconfigurability, consists of the modification of one or more of the following parameters: operating frequency, modulation scheme and transmission power. The goal of reconfigurability is to achieve higher performance without causing damage to PS [4].

In this paper we focus in the sensing problem within the OFDM context.

The presentation of this work is the following. In Section 2, the sensing problem is introduced for SP systems. A brief review of cyclostationarity concepts is presented in Section 3. Then, the schemes to be studied are summarized in Section 4. Comparative simulations in an OFDM context in addition to a discussion of the performance results are presented in Section 5. Finally, conclusions are derived in Section 6.

2. THE SENSING PROBLEM

The cognitive radio must distinguish between used and unused bands. Thus, it should be able to determine if a signal from a primary user (PS) is present or not. The basic hypothesis model for PS detection is the

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following

$$x(t) = \begin{cases} n(t) & \mathcal{H}_0 \\ h(t) * s(t) + n(t) & \mathcal{H}_1 \end{cases} \quad (1)$$

where $n(t)$ is AWGN, $h(t)$ is the impulse response of the channel, $s(t)$ the transmitted signal and $*$ denotes convolution.

On the other hand, cognitive radio networks which contend for an unused space can coexist with PS in the same frequency bands [5, 6, 7]. In these cases, detection of the different SS is an important task. Indeed, a cognitive radio network needs coordination to perform changes of band. This may be achieved using fixed common channels, although almost invariably this results in a considerable complexity and overhead. An alternative approach that solves both problems is to use signatures intentionally embedded in the SS signals to discriminate between them. For this situation the hypothesis model is slightly different, and it is the defined as follows

$$x(t) = \begin{cases} h(t) * S^{ns}(t) + n(t) & \mathcal{H}_0 \\ h(t) * S^{s}(t) + n(t) & \mathcal{H}_1 \end{cases} \quad (2)$$

where $s^s(t)$ and $s^{ns}(t)$ are the signals transmitted with signature and without signature, respectively.

Many detection techniques may be employed to distinguish between $\mathcal{H}_0$ and $\mathcal{H}_1$. In the following, we discuss three of the most popular techniques [4].

The *matched filter* is the optimum detector in AWGN, because it maximizes the SNR. This method requires less time to achieve good detection performance due to coherent detection. Also, it is possible to distinguish between different SS, as in hypothesis model (2). Signal parameters and synchronization are required. If the information of the PS is not known accurately the performance of the filter decreases significantly. In our case, neither signal parameters nor the channel are available and, therefore, the matched filter may not be a good choice.

The *energy detector* is very simple, and contrary to the matched filter it does not require accurate knowledge of the signal parameters. Only the noise power is required to establish the test threshold. Since this method is incoherent it has poor performance in noisy environments. For the same reason, it is not possible to distinguish between different signals. Then, this detector is only applicable to hypothesis model (1). Silent periods where the SS suspends the transmission and senses the spectrum must be employed to avoid that an outgoing signal of the SS interferes with the PS and results in a false alarm. Both the detector performance and the SS transmission efficiency are affected by these silent periods. Thus, the energy detector is not a suitable detector for the problem at hand.

The *Cyclostationary feature detector* detects periodic features in the transmitted signal which may be the result of, e.g., the modulation carrier, pulse train, cyclic prefix, pilot pattern. A cyclostationary signal is characterized by its time periodic mean and autocorrelation functions. As in conventional signal analysis, the signal processing may be performed in time or frequency domain according to the application of interest. This detector does not require to know accurately the signal parameters, only those related to the cyclostationary feature [8, 7]. AWGN is not cyclostationary, so this test performs well in low SNRs. Network identification based on hypothesis (2) may be achieved. To be used with (2) it is only required that the cyclostationary features of each system are different. For the same reason, silent periods are not necessary. Therefore, this detector is the best choice to detect PS [9].

In this paper two recently proposed algorithms for the detection of cyclostationarity features in OFDM signals are studied. The first one, exploits the redundancy that is generated by the cyclic prefix (CP) [8]. The second one, makes use of the pilot subcarriers to detect the OFDM signals [7].

3. CYCLOSTATIONARITY CONCEPTS

A cyclostationary signal is characterized by its time periodic mean and autocorrelation functions [10]. Considering discrete time, the autocorrelation function is defined as $R_{xx}(n,\nu) = E\{x(n)x^*(n+\nu)\}$, where $x(n)$ is a zero-mean complex almost-cyclostationary process. Since $R_{xx}(n,\nu)$ is a periodic function of discrete time n, it can be represented by

$$R_{xx}(n,\nu) = \sum_{\alpha\in\mathcal{A}} R_x^\alpha(\nu) e^{j2\pi\alpha n} \quad (3)$$

where the coefficients $R_x^\alpha(\nu)$, that define the cyclic autocorrelation function (CAF) at cycle frequency α are given by

$$R_x^\alpha(\nu) = \lim_{N\to\infty} \frac{1}{2N+1} \sum_{n=-N}^{N} R_{xx}(n,\nu) e^{-j2\pi\alpha n} \quad (4)$$

and $\mathcal{A} \triangleq \{\alpha \in [-1/2; 1/2) : R_x^\alpha(\nu) \not\equiv 0\}$. If $x(n)$ is a *cycloergodic* process, which is usually the case, the expectation operator implicit in (4) can be replaced by its time average using a sample mean. So, a more useful expression for the CAF is the following [11]

$$R_x^\alpha(\nu) = \lim_{N\to\infty} \frac{1}{2N+1} \sum_{n=-N}^{N} x(n)x^*(n+\nu) e^{-j2\pi\alpha n} \quad (5)$$

The cyclic spectral density function (SDF) is obtained by taking the Fourier transform

$$S_x^\alpha(k) = \sum_{\nu\in\mathbb{Z}} R_x^\alpha(\nu) e^{-j2\pi\nu k} \quad (6)$$

We see that for $\alpha = 0$, SDF reduces to the conventional power spectral density. However, for $\alpha \neq 0$, it can be shown that $S_x^\alpha(k)$ is the density of correlation between components at frequencies k and $k+\alpha$. Then, a useful approximation for $S_x^\alpha(k)$ is given by

$$\hat{S}^{(N)}(\alpha,k)=\frac{1}{N}\sum_{s=0}^{N-1}X^{(N)}(s)X^{*(N)}(s-\alpha)W^{(N)}(k-s) \tag{7}$$

where $X^{(N)}(k)$ is the Fourier transform of $x(n)$ defined by

$$X^{(N)}(k)=\sum_{n=0}^{N-1}x(n)e^{-j\frac{2\pi}{N}nk} \tag{8}$$

and $W^{(N)}(k)$ is a spectral window of length N that satisfies $\sum_k W^{(N)}(k)=1$. The windowing makes the estimator consistent. A tradeoff between bias and variance can be obtaining by a proper window selection [12].

4. CYCLOSTATIONARITY BASED DETECTION

We consider here a discrete-time baseband OFDM signal of N_c subcarriers that can be described by

$$s(m)=\sqrt{\frac{E_s}{N_c}}\sum_{k\in\mathbb{Z}}\sum_{n=0}^{N_c-1}c_k(n)e^{j2\pi\frac{n}{N_c}(m-D-k(N+D))} \times g(m-k(N+D)) \tag{9}$$

where E_s is the signal power, $c_k(n)$ are the transmitter symbols at nth subcarrier of kth OFDM symbol[1], D is the cyclic prefix length and $g(m)$ is the pulse shaping filter. Considering a noisy multipath slow fading channel, the equivalent baseband discrete-time channel impulse response is given by $\{h(l)\}_{l=0}^{L-1}$. Thus, the sampled OFDM signal at the receiver is modeled as

$$y(m)=e^{j(2\pi\epsilon\frac{m-\tau}{N}+\theta)}\sum_{l=0}^{L-1}h(l)s(m-l-\tau)+\eta(m) \tag{10}$$

where ϵ is the carrier frequency offset (normalized to the subcarrier spacing), θ is the initial arbitrary carrier phase, τ is the timing offset and $\eta(m)$ is a zero mean circularly-symmetric complex-valued AWGN of variance σ^2 per complex dimension.

4.1. Cyclic prefix detection

This scheme exploits the cyclic correlation function which is generated by the cyclic prefix of the OFDM signal. This means that the cyclostationary features depend only on the CP length (which is chosen to be longer than the channel impulse response to avoid ISI). Since CP is usually set by standards it cannot be modified (it is not possible to intentionally embed a signature in order to distinguish between two systems), CP based detection is based on hypothesis model 1.

From (9), considering at this point only an AWGN channel and that all carriers are modulated, it is easy to show that $R_{yy}(u,N_c)$ can be expressed as: $E_s\{\sum_k g(u+N_c-k(N_c+D))g^*(u-k(N_c+D))\}$, which is a periodic function of u of period $1/\alpha_0=N_c+D$. Therefore, (5) can be approximated by

$$\hat{R}_y^{k\alpha_0}(N_c)=\frac{1}{U-N_c}\sum_{n=0}^{U-N_c-1}y(n)y^*(n+N_c)e^{-j2\pi k\alpha_0 n} \tag{11}$$

where $\hat{R}_y^{k\alpha_0}(N_c)$ is the estimation of the cyclic correlation at cycle frequency $k\alpha_0$ and time lag N_c, and U is the number of available symbols at reception.

To build a robust cost function, $2N_b+1$ cyclic frequencies are taken into account, where N_b is the amount of positive cyclic frequencies. The considered cost function is given by [8]:

$$J_{cp}(N_b)=\sum_{k=-N_b}^{N_b}\left|\hat{R}_y^{k\alpha_0}(N_c)\right|^2 \tag{12}$$

In our detection context it is difficult to develop a detection test based on $\mathcal{H}_1$ (of hypothesis model 1), because signal parameters of PS are assumed unknown. Thus, the detection threshold is calculated from hypothesis $\mathcal{H}_0$. Under $\mathcal{H}_0$ the cycle coefficients of the received signal ($\hat{R}_y^{k\alpha_0}(N_c)$) are asymptotically normal with mean 0 and variance σ^4/U. Also, due to the OFDM (orthogonal) structure, these cyclic coefficients are asymptotically uncorrelated and hence mutually independent. As $J_{cp}(N_b)$ is a sum of $(2N_b+1)$ absolute-squared cycle coefficients, the distribution of the cost function $J_{cp}(N_b)$ is Chi-Squared with $2(2N_b+1)$ degrees of freedom ($\chi^2_{2(2N_b+1)}$) [8].

To build a detection test, a constant λ is defined so that $\mathcal{P}\{J_{cp}(N_b)\geq\lambda|\mathcal{H}_0\}=P_{fa}$, where P_{fa} is the fixed false alarm probability and λ is the test threshold. Then, considering $\gamma[2(2N_b+1),x]$ as the cumulative distribution function (cdf) of $\chi^2_{2(2N_b+1)}$, it is possible to evaluate the threshold as:

$$\lambda=\frac{\sigma^4}{U}\,\gamma^{-1}\left[(2N_b+1),1-P_{fa}\right] \tag{13}$$

Finally it is possible to perform the following test in order to distinguish between $\mathcal{H}_0$ and $\mathcal{H}_1$.

- if $J_{cp}(N_b)\leq\lambda$, then $\mathcal{H}_0$ is decided,
- if $J_{cp}(N_b)>\lambda$, then $\mathcal{H}_1$ is decided.

Note that the scale factor $\frac{\sigma^4}{U}$ is introduced as a normalization because the variances of the cyclic coefficients are not equal to one.

To summarize the method, the steps needed to perform the detection test are described in the following: 1) CAF is calculated for each of the $2N_b+1$ considered cycle frequencies using Eq. (11), 2) to obtain the cost function, absolute-squared CAF estimates are summed as stated in Eq. (12), where an estimate of the noise variance is necessary to calculate the test threshold in Eq. (13), and; 3) the value of the cost function is compared against the threshold λ.

[1]OFDM symbols are sometimes called blocks

4.2. Pilot induced detection

In case it is necessary to determine which signal of a signal set is transmitted, as stated in hypothesis model 2, each signal must have a distinctive signature to facilitate cyclostationary detection. Cyclostationary features may be induced using special preambles or repeating the information symbols over multiple carriers [6, 5]. However, the inserted preambles are not always present in the signal, which difficult the detection. Furthermore, repeated carriers reduce system capacity. An interesting approach to overcome these problems is to take advantage of existing pilot carriers and embed signature on them. This method is known as *pilot induced cyclostationary* signature (PIC) [7]. Pilot carriers are always present in OFDM signals for channel estimation and synchronization.

The pilot pattern is described by an index set $\mathcal{I}(k)$ that specifies the carriers which contain pilots symbols in the k^{th} OFDM block. The OFDM symbols $c_k(n)$ in (9) are divided in two sets: data symbols $a_k(n)$, when $n \notin \mathcal{I}(k)$, and pilot symbols $b_k(n)$, when $n \in \mathcal{I}(k)$. It is considered that under hypothesis $\mathcal{H}_0$ of model 2, the received signal does not have pilots in the same places than a signal under $\mathcal{H}_1$.

The demodulated signal is the DFT of the received block, and is expressed as follows

$$Y_k(n) = \frac{1}{\sqrt{N_c}} \sum_{m=0}^{N_c-1} y[k(N_c+D)+D+m] e^{-j2\pi \frac{nm}{N}} \quad (14)$$

Detection is performed by measuring the energy of the cyclic correlation function induced by the pilot pattern. The cost function is defined as [7]

$$J_{pic} = \sum_{(p,q)\in\xi} \left(\sum_{\alpha\in\mathcal{A}(p,q)} \left| \hat{R}^{\alpha}_{\tilde{Y}(p,q)}(d^{(p,q)}) \right|^2 \right) \quad (15)$$

where

$$\hat{R}^{\alpha}_{\tilde{Y}(p,q)}(d^{(p,q)}) = \frac{1}{M-d^{(p,q)}} \sum_{k=0}^{M-d^{(d,q)}-1} \tilde{Y}_k(p) \times \tilde{Y}^*_{k+d^{(p,q)}}(q) e^{-j2\pi\alpha k}, \quad (16)$$

$\mathcal{A}_{(p,q)} \triangleq \{\alpha \in [-1/2; 1/2] : R^{\alpha}_{Y(p,q)}(d^{(p,q)}) \neq 0\}$, p and q are the correlated carriers, $d^{(p,q)}$ is the separation in blocks between p and q and $\xi = \{(p,q)|\mathcal{A}_{(p,q)} \neq \emptyset$ and $d^{(p,q)} + K \leq M\}$, where K and M are the pilot pattern period and the number of received OFDM symbols, respectively. Also is it useful to define N_p as the cardinality of set ξ. In addition, to make the criterion J_{pic} less sensitive to the received signal gain, each term $Y_k(n)$ in Eq. (16) is normalized as follows

$$\hat{Y}_k(n) = \frac{Y_k(n)}{\sqrt{\widehat{Var}[Y(n)]}} \quad (17)$$

where $\widehat{Var}[]$ denotes estimated signal variance defined as: $\widehat{Var}[Y(n)] = 1/M \sum_{k=0}^{M-1} |Y_k(n)|^2$.

In order to embed a cyclostationary signature in the signal, pilot symbols are designed such that $b_k(p) = b_{k+d^{(p,q)}}(q)e^{i\varphi}$, with $\varphi \in [-\pi; \pi)$. Then, the processes $\{c_k(p)\}_k$ and $\{c_k(q)\}_k$ are jointly cyclostationary with cyclic coefficients

$$R^{\alpha}_{c(p,q)}(d^{(p,q)}) = \frac{\sigma_b^2 e^{-j(2\pi\alpha+\varphi)}}{K} \sum_{m\in\mathbb{Z}} \delta\left[\alpha - \frac{m}{K}\right] \quad (18)$$

which are non zero for α belonging to the set $\mathcal{A}_{(p,q)} = \{\frac{m-\lfloor K/2 \rfloor}{K},\ m \in \{0,1,\cdots,K-1\}\}$, where $\lfloor\ \rfloor$ stands for integer flooring.

In the same way that in the former method, the cyclic coefficients $R^{\alpha}_{Y(p,q)}(d^{(p,q)})$ are jointly Gaussian with zero mean and variance $1/(M-d)$. To evaluate the cost function in Eq. (15), K of these coefficients corresponding to the K induced cycle frequencies are summed. This results in a Chi-Square random variable of $2K$ degrees of freedom. Finally, J_{pic} is the result of the sum of N_p of these χ^2_{2K}. To find the test threshold it is mandatory to know the cdf of J_{pic}. Since some parameters of the signal are unknown, as in CP detector, the test statistics is based on hypothesis $\mathcal{H}_0$. Reference [7] proposes to approximate the cdf of J_{pic} by a Laguerre series. Since usually N_p is in the order of 30, we propose an alternative approach by exploiting to use the *Central Limit Theorem* (CLT) to find an expression to J_{pic}'s cdf. With this in mind, first the mean of J_{pic} is calculated and then its variance. The mean of J_{pic} is given by

$$\begin{aligned} \mathbb{E}\{J_{pic}\} &= \mathbb{E}\left\{\sum_{m=1}^{N_p} \sigma_R^2 \chi^2_{2K}(m)\right\} \\ &= N_p \sigma_R^2 \mathbb{E}\left\{\chi^2_{2K}\right\} \\ &= N_p \sigma_R^2 2K \end{aligned} \quad (19)$$

where $\sigma_R^2 = 1/(M-d)$ is the CAF variance and χ^2_{2K} is the central Chi-Squared random variable corresponding to CAF coefficients $R^{\alpha}_{Y(p,q)}[d^{(p,q)}]$. The variance of J_{pic} may be calculated as

$$\begin{aligned} \mathbb{V}ar\{J_{pic}\} &= \mathbb{V}ar\left\{\sum_{m=1}^{N_p} \sigma_R^2 \chi^2_{2K}(m)\right\} \\ &= N_p \sigma_R^4 \mathbb{V}ar\left\{\chi^2_{2K}\right\} \\ &= N_p \sigma_R^4 4K \end{aligned} \quad (20)$$

In Eqs. (19) and (20) the χ^2 distributions are considered iid. Finally, $J_{pic} \sim \mathcal{N}(N_p\sigma_R^2 2K,\ N_p\sigma_R^4 4K)$, so now it is straightforward to find a test threshold that meets some particular probability of false alarm. If the cdf of the normal distribution is defined as $\mathcal{P}(J_y \leq x|\mathcal{H}_0) = \psi(x)$, then the threshold λ is given by

$$\lambda = \{x \mid \mathcal{P}(J_y \leq x|\mathcal{H}_0) = 1 - P_{fa}\} = \psi^{-1}(1 - P_{fa}) \quad (21)$$

and the decision test is therefore

- if $J_{pic} \leq \lambda$, then $\mathcal{H}_0$ is decided,
- if $J_{pic} > \lambda$, then $\mathcal{H}_1$ is decided.

In order to clarify PIC detector, the steps needed to perform the test are summarized in the following: 1) CP removed received blocks are transformed into frequency domain using Eq. (14), 2) each pair of carriers $(p, q) \in \xi$ is normalized according to Eq. (17) to make the test independent of signal gains, 3) over each pair (p, q) CAFs are calculated for the K values of $\alpha \in \mathcal{A}$ as established by Eq. (18) and Eq. (16). To calculate the cost function, the absoluted-squared CAFs are first summed for each $\alpha \in \mathcal{A}$ and then over all pairs $(p, q) \in \xi$ as indicated in Eq. (15), 4) since J_{pic}'s cdf under $\mathcal{H}_0$ is too complicated for closed-form solution, we propose to use the CLT. Following this approach mean and variance of J_{pic} are calculated by Eq. (19) and Eq. (20), respectively, 5) When the cdf of J_{pic}, $\psi(x)$, is known, it is possible to find the threshold λ using Eq. (21), and; 6) the cost function is compared against the threshold to perform the test.

5. SIMULATIONS AND DISCUSSION

Simulations were made using OFDM signals of $N_c = 512$ carriers, cyclic prefix $D = N_c/8 = 64$, and sample rate $T_c = 0.5\mu s$ which corresponds to inter-carrier spacing of ~ 3.9kHz that is the similar to $2K$ DVB-T signal. $2K$ DVB-T has an inter-carrier spacing of 4.464kHz [13].

Each SNR value is averaged over 2000 trials generated with a different channel realization. Hypothesis $\mathcal{H}_0$ and $\mathcal{H}_1$ of the corresponding model are equally likely. P_{fa} is set to 1%. For the CP detector, each trial consist of 30 OFDM symbols, whereas in PIC detector 60 symbols are used.

The pilot pattern used in PIC detector to embed a signature is scattered and has period 2 ($K = 2$), i.e., it consists of two blocks. Pilot spacing is 12 carriers and the offset between blocks is 6 carriers. Pilot symbols are BPSK modulated as in [7] and equally likely. In order to maintain low test complexity and allow a considerable number of networks to share the spectrum, N_p (pairs of correlated carriers) is set to 30. Along the test, the separation between p and q carriers is set to 30 and $d^{(p,q)} = 1$.

Static and dynamic multipath channels are considered. The simulated propagation channel $\{h_k(l)\}_{l=0}^{L}$ has length $L = D$, and exponential decay profile $\mathbb{E}\{|h_k(l)|^2\} = Ge^{-l/\beta}$, where G is chosen such that $\sum_{l=0}^{L} \mathbb{E}\{|h_k(l)|^2\} = 1$ and $\beta = N_p/4$. Doppler frequency is set to 75Hz which corresponds to $\sim 2\%$ of inter-carrier spacing and a speed of 80Km/h with a carrier of 1GHz.

5. 1. Cyclic prefix detection

As seen in Eq. (12), when N_b increases the detection test has more redundance which results in better performance at the expense of system complexity. Fig. 1 shows the percentage of correct detection versus SNR for N_b from 0 to 4 over an AWGN channel.

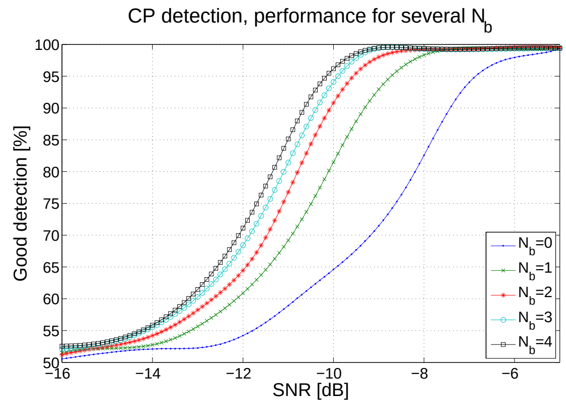

Figure 1: Comparison between several choices of N_b over an AWGN channel.

Channel effect over J_{cp} is to mis-match CP with the end of the OFDM symbol, so correlation decreases. The worse the channel the bigger the mis-match. Fig. 2 shows the results of the simulation for an AWGN channel, a static multipath channel and a dynamic multipath channel for $N_b = 4$.

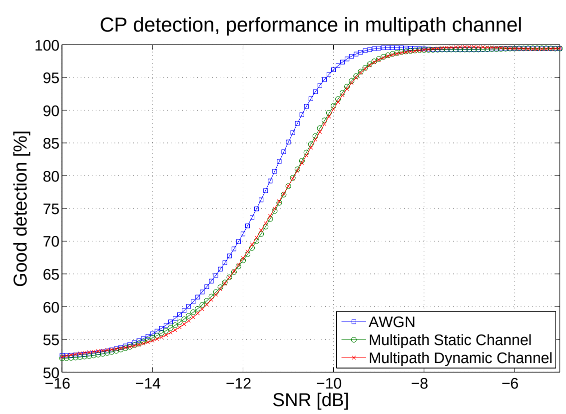

Figure 2: Performance of the J_{cp} test in different channel conditions.

5. 2. Pilot based detector

Fig. 3 shows the performance of J_{pic} test in AWGN, static multipath channel and dynamic multipath channel.

5. 3. Discussion

The studied systems do not require channel estimation, so channel condition has a big impact in performance. If information of the channel is not available

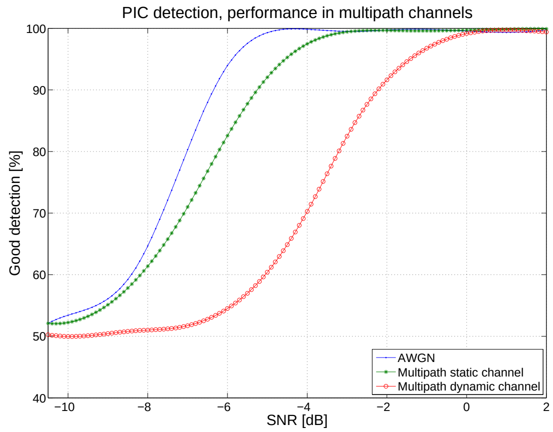

Figure 3: Performance of J_{pic} test for different channel conditions.

the only way to mitigate the lack of information is to increase redundancy.

It is known that OFDM systems are sensitive to Carrier Frequency Offset (CFO) and timing errors (TE) because they destroy orthogonality between carriers. On the other hand, the SP concept assumes that the PS parameters are unknown to the secondary system, so system synchronization is a big concern in SP. Since the test performance is affected by synchronization errors, it is possible to find ϵ and τ of Eq. (10) by looking for the maximum of J_{cp} or J_{pic} in a CFO versus TE plane [7]. However, an accurate estimation of impairments requires a fine search grid which increases the delay.

Note that the performance of J_{pic} is worse than the performance of J_{cp}. This follows from three different factors. Firstly, according to hypothesis model (2), the interference power is higher for the presence of the signal in $\mathcal{H}_0$. Secondly, J_{pic} employs less redundancy than J_{cp} in the cyclostationary signature which reduces test quality. Finally, correlated symbols pertain to different OFDM symbols and, therefore, the test is more sensitive to time-variant channels. On the other hand, the CP detector is quite robust to time-variant channels since correlation is calculated between symbols within an OFDM symbol.

A significant improvement in the performance will be achieved if channel estimation is performed although this may compromise the assumption that PS parameters are unknown.

6. CONCLUSIONS

Two cyclostationary methods that correspond to two hypothesis models for signal detection and identification were studied within different channel environments. The first one, J_{cp}, exploits the cyclic correlation function which is generated by the cyclic prefix of the OFDM signal. J_{cp} is simpler and more robust to channel distortion, because it employs more redundance. The second method, J_{pic}, takes advantage of existing pilot carriers and embed signature on them. J_{pic} is more general, because it allows detection of different services.

From this study it is clear that the operating environment of the detectors must be specified in detail in order to move towards a real application. This involves: performance of coarse and fine Carrier Frequency Offset (CFO) and timing errors (TE) estimators, and analysis of blind vs. non-blind algorithms. Also it is necessary to make an analytical study of CFO and TE effects over cyclostationarity detectors for OFDM signals.

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