On the Effect of Random Sampling Jitter on Cyclostationarity Based Spectrum Sensing Algorithms For Cognitive Radio

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***Abstract*—Cognitive radio is an enabling technology, which is expected to lead to a more efficient utilization of the available spectral resources due to its flexibility and its ability to sense its spectral environment. Recently, spectrum sensing methods based on exploiting the cyclostationary characteristics of communication signals have been drawing considerable interest. Imperfections in the cognitive radio receiver that affect the cyclic statistics of a signal of interest may lead to a degradation in the performance of spectrum sensing algorithms based on cyclostationarity. One such typical source of imperfection is random timing jitter in the sampling process. In this work, we explore the effect of random sampling jitter on the second order cyclostationary statistics of wide sense cyclostationary signals. General analytical expressions are derived for the cyclic statistics of sampled signals in the presence of sampling jitter and specific results are provided for two cases of interest. Subsequently, the effect of the jitter on a spectrum sensing algorithm is investigated via simulations.**

***Index Terms*—cognitive radio,spectrum sensing, cyclostationarity, sampling jitter.**

I. Introduction

Today's wireless systems are largely characterized by static spectrum allocations, which lead to a wasteful use of scarce and expensive spectral resources. The emergence of Cognitive Radio (CR) technology, which is considered to be the next evolutionary step in wireless communications, is expected to offer an unprecedented flexibility in terms of spectrum usage, paving way to more efficient dynamic and demand oriented spectrum allocation strategies. Opportunistic spectrum access schemes such as Spectrum Pooling can be regarded as the first steps into this direction[1].

Spectrum sensing is a key functionality of the CR, providing it with the required environmental awareness for an agile and flexible use of spectral resources. Recently,there has been an increasing interest on spectrum sensing methods based on exploiting the cyclostationary characteristics of communication signals, mainly due to the inherent robustness of such algorithms to noise and interference, and the signal selectivity they offer (see, for example [2], [3], [4],[5]). Virtually all communication signals exhibit cyclostationarity with cycle frequencies related to hidden periodicities underlying the signal, which can be used for this task, whereas some of the recently proposed methods also involve artificially embedding distinct cyclic signatures in signals [6]. Spectrum sensing algorithms based on cyclostationarity proposed for CR systems usually require an estimate of the second order cyclic statistics of the sampled version of the received signal, i.e. the cyclic autocorrelation function (CAF) or the spectral correlation density (SCD), in order to detect the presence of the cyclic signature of a signal of interest in the frequency environment.

In practice, imperfections in the signal generation or reception may affect the cyclic statistics of a signal of interest, leading to a degradation in the performance of cyclostationarity-exploiting spectrum sensing algorithms based on an ideal signal model. Sampling jitter is one such example, where the sampling instant undergoes a random shift in each sampling interval. Thus, evaluation of the effect of the sampling jitter on the cyclic statistics of a signal of interest in the presence of random sampling jitter is of considerable practical and theoretical interest.

The effect of the presence of sampling jitter on the power spectral density (PSD) of random signals is covered thoroughly in the existing literature. In [7], the effect of the pulse timing jitter on the cyclic statistics of a digitally modulated signal has been investigated. The purpose of this work is to provide a theoretical groundwork for investigating the effect of random sampling jitter on spectrum sensing methods based on cyclostationarity. First, we derive general analytical expressions for the second order discrete-time cyclic statistics of signals generated by sampling continuous time cyclostationary processes in the presence of random sampling jitter and provide specific results for two cases of interest. We show that presence of independent identically distributed (i.i.d.) sampling jitter leads to a complex attenuation of the spectral correlation density function which is both selective in the frequency and cyclic frequency parameters, and to the emergence of a constant additive term previously not present. Subsequently, the effect of the jitter on a spectrum sensing algorithm based on Dandawate's work in [8] is investigated via simulations.

II. Preliminaries

This section presents the mathematical preliminaries and definitions for discrete and continuous time cyclostationary processes, laying down the theoretical background for the rest of the work. A complex continuous time WSCS process $x(t)$ is characterized by a time varying autocorrelation function

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(TVAF) $R_x(t,\tau) = E\{x(t)x^*(t-\tau)\}$, which is periodic in time t with a fundamental period T_f and can be represented as a Fourier series:

$$R_x(t,\tau) = \sum_{\alpha} R_x^{\alpha}(\tau) e^{j2\pi\alpha t} \quad (1)$$

where the sum is taken over integer multiples of fundamental cycle frequency $\alpha_f = 1/T_f$. $R_x^{\alpha}(\tau)$ is referred to as the cyclic autocorrelation function, and is nonzero only for $\alpha = l/T_f$, i.e. for integer multiples of the fundamental cycle frequency. The SCD $S_x^{\alpha}(f)$ is defined as the Fourier transform of $R_x^{\alpha}(\tau)$ w.r.t. τ and can be seen as a measure of correlation between the spectral components of $x(t)$ separated in frequency by an amount of α. Clearly, for $\alpha = 0$, the SCD is equal to the PSD of $x(t)$. Similarly, a complex discrete time WSCS process $\tilde{x}_n$ has a time varying autocorrelation function $R_{\tilde{x}}[n,k] = E\{\tilde{x}_n \tilde{x}^*_{n-k}\}$, periodic in n with a period of N, which allows a discrete time Fourier series representation:

$$R_{\tilde{x}}[n,k] = \sum_{\beta} R_{\tilde{x}}^{\beta}[k] e^{j2\pi\beta n} \quad (2)$$

Where the sum is taken over N integer multiples of the fundamental cycle frequency $\beta_f = 1/N$. Note that the discrete-time CAF $R_{\tilde{x}}^{\beta}[k]$ is periodic in β with a period of 1 and is nonzero only for integer multiples of β_f. The discrete time Fourier transform of the discrete-time CAF results in an SCD $S_{\tilde{x}}^{\beta}(\omega)$ which is also periodic in ω with a period of 2π.

It is easily shown that a discrete-time random signal $\tilde{x}_n$ generated by sampling a continuous-time WSCS signal $x(t)$ with a fundamental cyclostationarity period T_f is itself WSCS with a fundamental period N, provided that T_f is an integer multiple of the sampling time T_s, i.e. $T_f/T_s = N$. In such a case, the CAF and SCD functions of $\tilde{x}_n = x(nT_s)$ can be expressed in terms of the CAF and SCD functions of the continuous time process $x(t)$ in the following manner:

$$R_{\tilde{x}}^{\beta=\frac{m}{N}}[k] = \sum_{p=-\infty}^{\infty} R_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}(kT_s) \quad (3)$$

and

$$S_{\tilde{x}}^{\beta=\frac{m}{N}}(\omega) = \frac{1}{T_s} \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}\left(\frac{\omega}{2\pi T_s} + \frac{q}{T_s}\right) \quad (4)$$

(See also [9]). From (4) it is evident that aliasing takes place in both frequency and cycle-frequency domains.

III. Effect of the Sampling jitter on the Cyclic Statistics

Assuming that the timing of the sampling instants are subject to random jitter, the resulting discrete-time process y_n can be modeled as

$$y_n = x(nT_s - \epsilon_n) \quad (5)$$

where ϵ_n is a discrete-time random process representing the jitter.The TVAF of y_n can be written in terms of $R_x(t,\tau)$:

$$R_y[n,k] = E\{R_x(nT_s - \epsilon_n, kT_s - \epsilon_n + \epsilon_{n-k})\} \quad (6)$$

Where the expectation operation is performed over the jitter process. Assuming that the jitter is independent of $x(t)$, after substituting (1) in (6) and expressing the CAF in terms of the SCD, (6) can be reexpressed as

$$R_y[n,k] = \sum_{l=-\infty}^{\infty} \int_{-\infty}^{\infty} S_x^{\alpha=\frac{l}{T_f}}(f) e^{j2\pi f k T_s} e^{j2\pi l n/N} \times E\{e^{-j2\pi f(\epsilon_n - \epsilon_{n-k})} e^{-j2\pi l \epsilon_n / T_f}\} df \quad (7)$$

Thus, the CAF of y_n is periodic in n with a period N, i.e y_n is WSCS with the same fundamental cycle frequency as $\tilde{x}_n$, provided that the jitter process is stationary. Defining the joint characteristic function of the jitter as

$$C(\theta,\phi,k) = E\{e^{-j2\pi\theta\epsilon_n} e^{j2\pi\phi\epsilon_{n-k}}\}, \quad (8)$$

the CAF of y_n can be expressed in terms of the SCD function of $x(t)$ and $C(\theta,\phi,k)$:

$$R_y^{\beta=\frac{m}{N}}[k] = \sum_{p=-\infty}^{\infty} \int_{-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}(f) \times C\left(f + \frac{m}{T_f} + \frac{p}{T_s}, f, k\right) e^{j2\pi f k T_s} df \quad (9)$$

and the SCD can be calculated by fourier transforming (9) w.r.t. to index k.

Equation (9) is general and applies to arbitrary stationary jitter statistics. In the following, the special case of an independent identically distributed (i.i.d.) jitter sequence is investigated. In this case, the joint characteristic function of the jitter can be simplified as:

$$C(\theta,\phi,k) = \begin{cases} \varepsilon(\theta)\varepsilon^*(\phi) & k \neq 0 \\ \varepsilon(\theta-\phi) & k = 0 \end{cases} \quad (10)$$

with $\varepsilon(\theta) = E\{e^{-j2\pi\theta\epsilon_n}\}$, or equivalently,

$$C(\theta,\phi,k) = \varepsilon(\theta)\varepsilon^*(\phi) - \delta_k(\varepsilon(\theta)\varepsilon^*(\phi) - \varepsilon(\theta-\phi)), \quad (11)$$

where δ_k is the discrete time impulse function. Substituting (11) in (9), the CAF of y_n becomes

$$R_y^{\beta=\frac{m}{N}}[k] = \begin{cases} \sum_{p=-\infty}^{\infty} \int_{-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}(f) \varepsilon(f + \frac{m}{T_f} + \frac{p}{T_s}) \times \varepsilon^*(f) e^{j2\pi f k T_s} df, & k \neq 0 \\ \sum_{p=-\infty}^{\infty} \int_{-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}(f) \times \varepsilon(\frac{m}{T_f} + \frac{p}{T_s}) df, & k = 0 \end{cases} \quad (12)$$

Finally, fourier transforming (12) leads to the SCD function:

$$S_y^{\beta=\frac{m}{N}}(\omega) = \frac{1}{T_s} \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}\left(\frac{\omega}{2\pi T_s} + \frac{q}{T_s}\right) \times \varepsilon\left(\frac{\omega}{2\pi T_s} + \frac{m}{T_f} + \frac{p}{T_s} + \frac{q}{T_s}\right) \varepsilon^*\left(\frac{\omega}{2\pi T_s} + \frac{q}{T_s}\right) + \sum_{p=-\infty}^{\infty} \int_{-\infty}^{\infty} S_x^{\alpha=\frac{m}{T_f}+\frac{p}{T_s}}(f) \Big(\varepsilon\left(\frac{m}{T_f} + \frac{p}{T_s}\right) - \varepsilon\left(f + \frac{m}{T_f} + \frac{p}{T_s}\right) \varepsilon^*(f) \Big) df \quad (13)$$

It is easily shown that $|\varepsilon(f)| \leq \varepsilon(0) = 1$ for any jitter distribution. Thus, comparing (13) with (4), it is evident that the presence of random i.i.d. sampling jitter leads to a complex attenuation of the SCD, selective both in the frequency parameter ω and cycle frequency parameter β, which can be seen in the first summand of (13) and it adds an additional, previously nonexistent constant term to the SCD function given in the second summand of (13).The magnitude of the attenuation increases with increasing β, for $0 \leq \beta \leq N/2$. Obviously, the amount of distortion in the SCD and the CAF functions increases with increasing jitter variance. Assuming that, in practice, the sampling jitter will not exceed a fraction of the sampling interval T_s, the "bandwidth" of the characteristic function $\varepsilon(f)$ will be in the same order of magnitude with the sampling rate $f_s = 1/T_s$ or larger. Thus,for a given sampling rate and jitter distribution, WSCS signals with a larger bandwidth (i.e. those sampled nearly at the Nyquist sampling rate) are more affected by the distortion of the SCD than signals with lower bandwidth (i.e. which are oversampled by a significant amount). This distinction will be more apparent in the next section, where we provide one example for each case.

IV. Application Examples

In this section, the effect of the sampling jitter on the cyclic statistics is investigated for two examples of practical interest. First, the SCD of a simple pulse train modulated by a wide sense stationary (WSS) sequence is evaluated. Subsequently, the case of an OFDM signal is explored. For both cases, magnitude estimates of the SCD functions are provided, where the jitter distribution is assumed to be gaussian i.i.d., for which the jitter characteristic function is given as:

$$\varepsilon(f) = e^{-2\pi^2\sigma^2 f^2}, \tag{14}$$

with the jitter variance σ^2.

A. SCD of a Digital Pulse Stream With i.i.d. Sampling Jitter

A pulse stream modulated by a discrete-time sequence can be modeled as:

$$x(t) = \sum_{n=-\infty}^{\infty} a_n p(t - nT_p), \tag{15}$$

with the WSS modulating sequence a_n, pulse waveform $p(t)$ and the pulse interval T_p. It is straightforward to show that $x(t)$ is WSCS with a fundamental period $T_f = T_p$. The SCD of $x(t)$ can be given as:

$$S_x^{\alpha=\frac{m}{T_p}}(f) = \frac{1}{T_p} S_a(2\pi f T_p) P(f + \frac{m}{T_p}) P^*(f), \tag{16}$$

where $P(f)$ is the Fourier transform of $p(t)$ and $S_a(\omega)$ the PSD of a_n. Assuming that the sampling is performed such that $T_s = T_p/N$, the resulting signal y_n is WSCS. The SCD

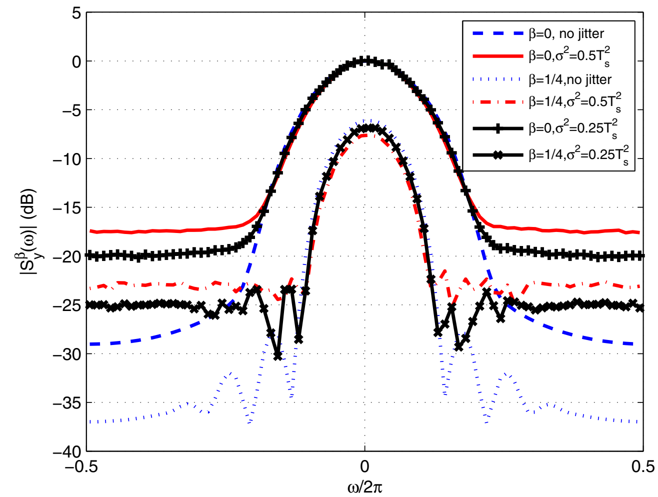

Fig. 1. Estimates of $|S_y^\beta(\omega)|$ for a QPSK modulated pulse train and for $\beta = 0$ and $\beta = 1/4$.

can be calculated by substituting (16) in (13):

$$\begin{aligned} S_y^{\beta=\frac{m}{N}}(\omega) &= \frac{S_a(\omega N)}{T_p T_s} \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} P(\frac{\omega}{2\pi T_s} + \frac{m}{T_p} + \frac{p}{T_s} + \frac{q}{T_s}) \\ &\times P^*(\frac{\omega}{2\pi T_s} + \frac{q}{T_s}) \varepsilon(\frac{\omega}{2\pi T_s} + \frac{m}{T_p} + \frac{p}{T_s} + \frac{q}{T_s}) \varepsilon^*(\frac{\omega}{2\pi T_s} + \frac{q}{T_s}) \\ &+ \frac{1}{T_p} \sum_{p=-\infty}^{\infty} \int_{-\infty}^{\infty} S_a(2\pi f T_p) P(f + \frac{m}{T_p} + \frac{p}{T_s}) P^*(f) \\ &\times \Big(\varepsilon(\frac{m}{T_p} + \frac{p}{T_s}) - \varepsilon(f + \frac{m}{T_p} + \frac{p}{T_s}) \varepsilon^*(f) \Big) df \end{aligned}$$

Fig.1 shows the normalized estimate of the magnitude of the SCD function for a QPSK modulated pulse stream where a raised cosine pulse waveform with a roll-of factor $\rho = 0.9$ is employed and the sampling time is chosen such that $N = T_p/T_s = 4$. Three different cases are displayed: jitter free sampling, Gaussian distributed jitter with $\sigma^2 = 0.25T_s^2$ and $\sigma^2 = 0.5T_s^2$. Only the spectral correlation surfaces for $\beta = 0$ (i.e. the PSD) and $\beta = 1/4$ are shown, and the surface for $\beta = 1/2$ is omitted, since for a raised cosine pulse shape, the amount of spectral correlation at this cycle frequency is too insignificant to be useful for detection. Note that even in the presence of sampling jitter with a rather high variance, the attenuation in the main lobes of the SCD surfaces due to jitter is very small, and the magnitude of the constant SCD term is so low that it can only be perceived if logarithmic scale is used. Clearly, this is due to the fact that the signal is oversampled, and for higher harmonics of the cycle frequency, for which the attenuation could have become effective despite the oversampling, the signal does not exhibit any significant amount of spectral correlation to begin with.

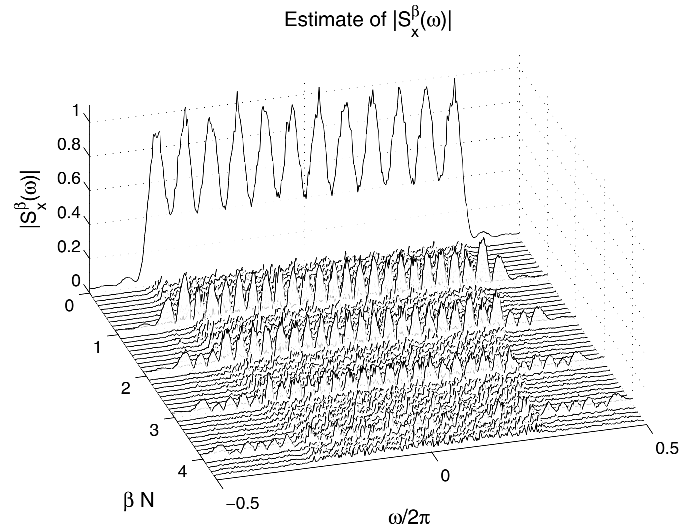

Fig. 2. Estimate of $|S_{\tilde{x}}^{\beta}(\omega)|$ for an OFDM signal in the absence of timing jitter.

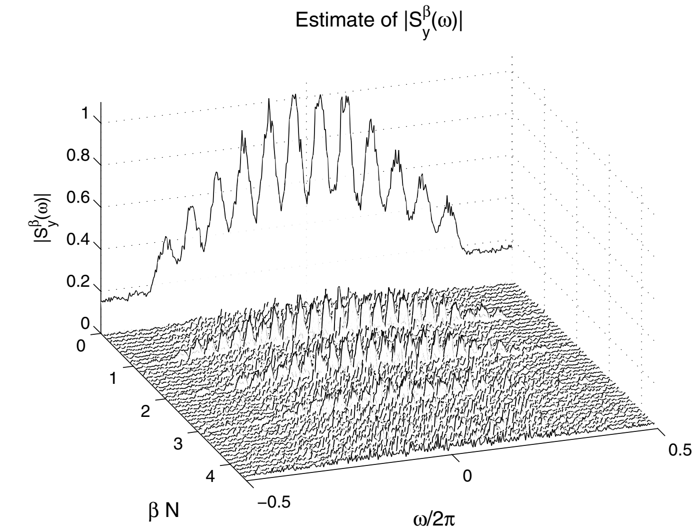

Fig. 3. Estimate of $|S_{y}^{\beta}(\omega)|$ for an OFDM signal in the presence of Gaussian i.i.d. timing jitter with $\sigma^2 = 0.25T_s^2$

B. SCD of an OFDM signal With i.i.d. Sampling Jitter

The baseband signal model for an OFDM signal can be given as:

$$x(t) = \sqrt{\frac{1}{N_c}} \sum_{n=-\infty}^{\infty} \sum_{i=0}^{N_c-1} d_{n,i} e^{j2\pi i \Delta_f (t-nT_d)} g_R(t-nT_d) \times e^{-j2\pi \frac{N_c-1}{2}\Delta_f t} \quad (17)$$

where $d_{n,i}$ is the n'th information symbol modulated on the i'th carrier, N_c is the number of carriers, Δ_f is the carrier separation, and $g_R(t)$ is the rectangular pulse function of length T_d. $T_d = T_u + T_g$ is the symbol length, where $T_u = 1/\Delta_f$ is the useful symbol duration and T_g is the length of the guard interval where the symbol is extended cyclically. It is straightforward to show that the OFDM signal is WSCS with a fundamental period of T_d. Assuming that the data sequence is zero mean and uncorrelated, the SCD can be calculated as:

$$S_x^{\alpha=\frac{m}{T_d}}(f) = \frac{A}{T_d} \sum_{i=0}^{N_c-1} G_R(f - i\Delta_f + \frac{N_c-1}{2}\Delta_f + \frac{m}{T_d}) G_R^*(f - i\Delta_f + \frac{N_c-1}{2}), \quad (18)$$

with $A = \sigma_d^2/N_c$, $\sigma_d^2 = E\{d_{n,i}d_{n,i}^*\}$ and $G_R(f) = \frac{\sin(\pi f T_d)}{\pi f}$, the Fourier transform of $g_R(t)$ [2]. The SCD of the sampled OFDM signal in presence of sampling jitter can easily be calculated by substituting (18) in (13). Fig. 2 displays the normalized magnitude estimate of the SCD function of a sampled OFDM signal where $N_c = 12$, $N = T_d/T_s = 20$, $T_g = T_d/5$. Clearly, in contrast to the previous case, the SCD of the signal exhibits a large spectral occupation, even for higher harmonics of the fundamental cycle frequency, and the sampling rate is only slightly higher than the Nyquist rate, thus we expect the cyclic statistics of this signal to be more sensitive to sampling jitter effects than that of the previous example, which is confirmed by Fig. 3 where the SCD magnitude of the signal in presence of Gaussian jitter with a variance $\sigma^2 = 0.25T_s^2$ is displayed. Unlike the previous case, the distortion of the SCD due to the sampling jitter is clearly discernable.

V. Detection Results

In the following, the effect of the sampling jitter on the detection performance of a cyclostationarity based spectrum sensing algorithm is investigated via simulations. In this work, we employ a modified version of the time domain constant false alarm rate (CFAR) test for the presence of cyclostationarity introduced by Dandawate et al. in [8] for discrete time cyclostationary processes, for this task. Since Dandawate's algorithm does not require any explicit assumptions on the data or noise distributions, it is expected to be relatively robust to nonidealities in the signal.

A. The Detection Algorithm

The consistent estimate of the discrete time conjugate CAF of the received signal x_n for a given cycle frequency β_0,is given as:

$$\hat{R}_x^{\beta_0}(k) = \frac{1}{N_o} \sum_{i=0}^{N_o-1} x_n x_{n-k}^* e^{-j2\pi\beta_0 n} = R_x^{\beta_0}(k) + \Delta_x^{\beta_0}(k) \quad (19)$$

where N_o is the length of the observation window and $\Delta_x^{\beta}(k)$ is the estimation error,which vanishes as $N_o \to \infty$. Using (19),the $1 \times 2L$ row vector consisting of cyclic autocorrelation estimates at the cycle frequency $\beta = \beta_0$ is defined as:

$$\hat{\mathbf{r}}_\mathbf{x} = \Big[Re\Big\{\hat{R}_x^{\beta_0}(k_1)\Big\}, ..., Re\Big\{\hat{R}_x^{\beta_0}(k_L)\Big\}, Im\Big\{\hat{R}_x^{\beta_0}(k_1)\Big\}, ..., Im\Big\{\hat{R}_x^{\beta_0}(k_L)\Big\}\Big] \quad (20)$$

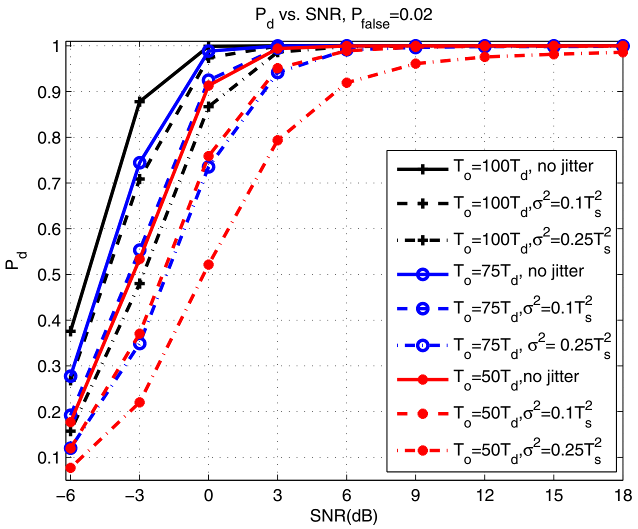

Fig. 4. P_d vs. SNR for an OFDM signal in presence of sampling jitter, $P_{false} = 0.02$

The test statistic for the employed spectrum sensing algorithm is given as [8]:

$$Z_x = N_o \hat{\mathbf{r}}_{\mathbf{x}} \hat{\boldsymbol{\Sigma}}_{\mathbf{x}}^{-1} \hat{\mathbf{r}}_{\mathbf{x}}^{\mathbf{T}}, \tag{21}$$

where $\hat{\boldsymbol{\Sigma}}_{\mathbf{x}}$ is the estimated asymptotic covariance matrix of the cyclic autocorrelation estimation error, which is calculated using fourth order cyclic statistics of the data (see [2] and [8] for details). It can be shown that, if the signal of interest is absent, the distribution of Z_x converges asymptotically to a central χ^2 distribution with $2L$ degrees of freedom, irrespective of the distribution of the input data. Hence, for a given threshold, the false alarm probabilities can be analytically calculated for large enough observation intervals N_o, regardless of the particular signal, leading to an asymptotically constant false alarm rate (CFAR) test. If the signal of interest is present, the distribution of the test statistics converges to a normal distribution [8]:

$$\lim_{N_o \to \infty} Z_x \stackrel{D}{=} \mathcal{N}(N_o \mathbf{r}_{\mathbf{x}} \boldsymbol{\Sigma}_{\mathbf{x}}^{-1} \mathbf{r}_{\mathbf{x}}^{\mathbf{T}}, 4N_o \mathbf{r}_{\mathbf{x}} \boldsymbol{\Sigma}_{\mathbf{x}}^{-1} \mathbf{r}_{\mathbf{x}}^{\mathbf{T}}),$$

where $\mathbf{r}_{\mathbf{x}}$ is the row vector consisting of the actual CAF values of the signal. Thus, any attenuation in the spectral correlation exhibited by the signal of interest is expected to lead to a decrease in the detection performance of the algorithm.

B. Simulation Results

In the following simulations, the effect of the sampling jitter on the performance of the detection algorithm in (21) is investigated. The signal of interest is an OFDM signal with QPSK modulation on all subcarriers, $N_c = 12$, $N = T_d/T_s = 20$, $T_g = T_d/5$. The channel is purely AWGN and the jitter is Gaussian i.i.d. Only the fundamental cycle frequency $\beta_0 = 1/20$ is considered for detection. Simulations are performed for 3 different values of observation times $T_o = N_o T_s$ and for each value of T_o, three cases are considered: absence of sampling jitter, jitter with variance $\sigma^2 = 0.1T_s^2$ and jitter with $\sigma^2 = 0.25T_s^2$.The probability of detection P_d and the false alarm rate P_{false} are defined as:

$P_d = Prob$(OFDM detected|OFDM signal present)

$P_{false} = Prob$(OFDM detected|No OFDM signal present)

The results of the simulations are summarized in Fig.4, where P_d vs. SNR curves for a fixed false alarm rate $P_{false} = 0.02$ are plotted. The results show that, as expected, the distortion of the cyclic statistics of the signal of interest due to the presence of random sampling jitter causes a severe degradation in the performance of the detection scheme, especially for lower values of T_o.

VI. Conclusion

In this work, we present a theoretical groundwork for analyzing the effect of random sampling jitter on cyclostationarity based spectrum sensing methods for cognitive radio systems. It is shown that the sampled versions of WSCS signals remain WSCS in presence of sampling jitter, provided that the jitter process is stationary. General expressions for the CAF and SCD functions under arbitrary stationary jitter have been derived and it is shown that the presence of independent identically distributed (i.i.d.) sampling jitter leads to a complex attenuation of the spectral correlation density function which is both selective in the frequency and cyclic frequency parameters, and to the emergence of a constant additive term. Examples of SCD functions are provided for two cases of practical interest and verified by numerical results. The detrimental effect of sampling jitter on the performance of a spectrum sensing algorithm based on cyclostationarity is investigated by simulations. Future work will include a thorough theoretical analysis of the jitter effects on existing cyclostationarity based spectrum sensing algorithms and the design and analysis of algorithms robust to the effects of random jitter.

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