# A Method for the Detection of Narrowband Interferer

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***Abstract*— A novel method for the detection of narrowband interferer is presented, specifically designed for time-hopping impulse radio receivers. Closed form expressions for the performance have been derived and are presented employing the central limit theorem. The detection probability and false alarm probability have been considered as measure for the performance. The correctness of the closed form expressions has been validated by means of Monte-Carlo simulations. Finally, the impact of several parameters on the performance is discussed.**

***Index Terms*— UWB, Impulse Radio, Coexistence, Detection.**

## I. INTRODUCTION

Almost all communications systems in use today deploy a sinusoid as an elementary waveform, on which information is mapped via certain kinds of modulation. Signal energy is hence concentrated in a well defined frequency band, which facilitates noise and interference suppression, e.g. by means of band-pass filtering. Unfortunately, such narrow band systems are inherently sensitive to fading. To obtain a robust wireless communication system, a fading margin has to be implemented with a negative effect on the system capacity . In the last ten years the interest in ultra wide band (UWB) technology has grown [1], [2] and [3], due to its inherent resilience against fading and implementation simplicity in case of impulse transmission. A system deploying this technique is often referred to as an impulse radio (IR). Due to its high bandwidth a UWB signal is able to resolve its surrounding with high resolution, which in principle allows a single UWB device to be used for communication and radar applications. Currently, regulation authorities in both Europe and the USA are in the process of developing legislation for UWB unlicensed deployment. Clearly, no gigahertz bandwidth at the lower frequencies can be allocated for UWB on a exclusive basis. The US telecommunications regulator (FCC) has adopted a First Report and Order to revise Part 15 regarding UWB devices. Particularly, the frequency range from 3.1 GHz to 10.6 GHz has been released for UWB communication devices. Other communication systems, like IEEE802.11, occupy slices of this vast frequency range and operate at relatively high spectral densities. Consequently, UWB systems will have to coexist with a number of radio systems, in order to be of interest from a technological and/or economic perspective.

In general, the interference from radio systems may be characterized as narrowband compared to UWB. Consequently, the UWB receiver could suppress the interference by means of signal processing utilizing the spectral differences between the interference and UWB signals. At initiation, the receiver is unaware on the momentary spectral characteristics of the interference from radio system. A UWB system could assume a worst case scenario and avoid the use of spectral bands, which are potentially used by others radio systems. Hence, the UWB system is very robust, but wasting capacity. An alternative approach is a receiver that measures the momentary interference situation and adapts to it. For instance, the receiver could inform the peer transmitter(s). Hence, the peer could notch that frequency at which the interferer is active by means of the TH-code [7],[6]. This paper presents a method for the detection of narrowband interferer for a time hopping impulse radio receiver.

## II. THE CONCEPT

Let's assume a pulse amplitude modulated (PAM) impulse radio (IR) signal, i.e.

$$y(t) = \sum_k a_k p(t - (kN_h + c_k)T_c) \tag{1}$$

where $p(t)$, $T_c$ and $N_h$, represent the pulse, the chip duration and the number of chips per bit, respectively. Both $a_k$ and $c_k$ are independent discrete random variables with the probabilities $P(a_k = a) = \frac{1}{2}$ and $P(c_k = c) = N_h^{-1}$, where $a \in \{-1, 1\}$ and $c \in \{0, 1, ..., N_h - 1\}$.

Now, let's assume a receiver matched to the signal $p(t)$. Consequently, the received signal will be passed through a filter with impulse response $h(t)$=$p(-t)$ whose output will be denoted by $f(t)$. For the sake of simplicity the issue of causality has been neglected and the pulse duration is assumed to be shorter than the chip duration. The demodulator output corresponding to the $k$-th pulse $\tilde{f}[k] = f(t)$, given that $t = (c_k + kN_h)T_c$. Consequently, the demodulator generates a stream of samples taken pseudo-randomly in time with an average sample period of $N_hT_c$.

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If the UWB receiver is disturbed by a narrowband interferer, the signal $f[m]$ will contain a residue of the narrowband interferer. The detection of the presence of a narrowband signal with an unknown amplitude, phase, and frequency in an uniformly sampled data stream is a well-known problem [4]. Therefore, additional zeros are inserted into the demodulator output sample stream at those chip positions, where no actual sample is taken. Consequently, a uniformly sampled date stream is obtained with a sample period equal to the chip duration. The sample taken at the $n$-th chip

$$f[n] = f(t)c(t)|_{t=nT_c} \tag{2}$$

where $c(t)$ represents the periodic, non-uniform sampling scheme, such that,

$$c(t) = \sum_k \delta(t - (kN_h + c_k)T_c) \tag{3}$$

where $\delta(k)$ denotes the Kronecker delta pulse. A well-known signal processing technique to bring out a narrowband signal is a spectrogram. Basically, a spectrogram is a power spectral density (PSD) estimation. Similar to [4], the spectrogram of the signal $f$ is defined as

$$I_{n_0}^{(f)}[k] \triangleq \left| \sum_{n=0}^{N-1} f[n+n_0] \exp\left(-j2\pi \frac{k}{N} n\right) \right|^2 \tag{4}$$

where $n_0$ denotes the start of the observation interval of length $N$, which is used to estimate the PSD. Furthermore, $k$ denotes a frequency identifier, where the actual frequency under evaluation is equal to $k/N$. Consequently, the frequency resolution is equal to $1/N$ and the spectrogram will be computed for $k \in \{0, 1, ..., N/2-1\}$. Whenever a narrowband interferer is active at frequency $k$ at time $n_0$, the spectrogram value at the specific $(k, n_0)$-combination will in average be larger. Here, a threshold detector is adopted, which reports a detection at frequency $k$, if the spectrogram $I[k]$ exceeds a certain threshold. Evidently, the performance of the algorithm will depend on the statistical properties of the spectrogram and on the threshold. In the following sections, both issues will be evaluated.

## III. The Statistical Properties of the Spectrogram

The spectrogram is a non-linear function of the received signal. Therefore, two alternative expressions for Eq.(4) are introduced, which help to derive the statistical properties of the spectrogram. If a signal is real valued in the time domain, the multiplication of this signal with its conjugated in the discrete Fourier domain is equivalent to the discrete Fourier transform of a circular convolution of the signal with itself, such that

$$I_{n_0}^{(f)}[k] = \sum_{n=0}^{N-1} r_{n_0}^{(f)}[n] \exp\left(-j2\pi \frac{k}{N} n\right) \tag{5}$$

where $r_{n_0}$ denotes a circular convolution over the interval defined by the time window, such that

$$r_{n_0}^{(f)}[n] = \begin{cases} \displaystyle\sum_{m=n_0}^{n_0+N-1} f[m+n]f[m] & m+n < N \\ \displaystyle\sum_{m=n_0}^{n_0+N-1} f[m+n-N]f[m] & m+n \geq N \end{cases} \tag{6}$$

Furthermore, the relation $\exp(jx) = \cos(x) + j\sin(x)$ allows us to rewrite the spectrogram

$$I_{n_0}^{(f)}[k] = \left(X_{n_0}^{(f)}[k]\right)^2 + \left(Y_{n_0}^{(f)}[k]\right)^2 \tag{7}$$

where

$$X_{n_0}^{(f)}[k] \triangleq \sum_{n=0}^{N-1} f[n+n_0] \cos\left(\frac{2\pi\ k}{N} n\right) \tag{8}$$

$$Y_{n_0}^{(f)}[k] \triangleq \sum_{n=0}^{N-1} f[n+n_0] \sin\left(\frac{2\pi\ k}{N} n\right) \tag{9}$$

To simplify the notation, the function identifier $(f)$ will be omitted whenever the signal $f[n]$ is the input. Furthermore, the $n_0$ identifier is omitted, since the statistical properties of the spectrogram are independent of its value.

The signal $f[m]$ is composed out of three terms, a IR signal term $y[m]$, a noise term $n[m]$ and the narrowband interferer $s[m]$, which are all random of nature. Since $X[k]$ is the sum of $N$ random variables (RVs), it is reasonable to assume that $X[k]$ is a random variable (RV) with a Gauss distribution given that $N$ is significantly large. Furthermore, it is evident that $Y[k] = 0$ if $k = 0$. Otherwise, $Y[k]$ is assumed to be a RV with a Gauss distribution, independent but identically distributed as $X[k]$. Consequently, $I[k]$ will be a RV which arises as a result of summing the squares of $v$ i.i.d. Gaussian RV, where $v = 1$ if $k = 0$ and otherwise equal to two. In both cases, $I[k]$ will have a chi-squared distribution, which is given by

$$p(x;\lambda;\beta;v) = \frac{1}{2\beta}\left(\frac{x}{\lambda}\right)^{\frac{v-2}{4}} \exp\left[-\frac{x+\lambda}{2\beta}\right] I_{\frac{v-2}{2}}\left(\frac{\sqrt{\lambda x}}{\beta}\right) \tag{10}$$

for all $x \geq 0$ and zero if $x < 0$, where $I_r(u)$ denotes the modified Bessel function of the first kind and order $r$. Furthermore, $\lambda$, $\beta$ and $v$ denote the non-centrality parameter, the scale parameter and the shape parameter, respectively. The values of $\lambda$, $\beta$ and $v$ depend on both the mean and the variance of $X[k]$ and $Y[k]$, according to

$$\begin{aligned} \lambda[k] &= \mathrm{E}\{X[k]\}^2 + \mathrm{E}\{Y[k]\}^2 \\ \beta[k] &= \mathrm{var}\{X[k]\} + \mathrm{var}\{Y[k]\} \\ v[k] &= 2 - \delta[k] \end{aligned} \tag{11}$$

Furthermore, the expectation for the spectrogram [4],

$$\mathrm{E}\{I[k]\} = \lambda[k] + \beta[k] \tag{12}$$

and is also equal to

$$\mathrm{E}\{I[k]\} = \sum_{n=0}^{N-1} \mathrm{E}\{r[n]\} \exp\left(-j2\pi \frac{k}{N} n\right) \tag{13}$$

Consequently, the expectation $\mathrm{E}\{r[n]\}$ and the non-centrality parameter $\lambda[k]$ completely describe the distribution of $I[k]$. The three terms that construct the signal $f[m]$ are assumed statistically independent, such that

$$\begin{aligned}\mathrm{E}\{X[k]\} &= \mathrm{E}\left\{X^{(y)}[k]\right\} + \mathrm{E}\left\{X^{(n)}[k]\right\} + \mathrm{E}\left\{X^{(s)}[k]\right\} \\ \mathrm{E}\{Y[k]\} &= \mathrm{E}\left\{Y^{(y)}[k]\right\} + \mathrm{E}\left\{Y^{(n)}[k]\right\} + \mathrm{E}\left\{Y^{(s)}[k]\right\}\end{aligned} \quad (14)$$

and

$$\begin{aligned}\mathrm{var}\{X[k]\} &= \mathrm{var}\left\{X^{(y)}[k]\right\} + \mathrm{var}\left\{X^{(n)}[k]\right\} + \mathrm{var}\left\{X^{(s)}[k]\right\} \\ \mathrm{var}\{Y[k]\} &= \mathrm{var}\left\{Y^{(y)}[k]\right\} + \mathrm{var}\left\{Y^{(n)}[k]\right\} + \mathrm{var}\left\{Y^{(s)}[k]\right\}\end{aligned} \quad (15)$$

$y$, $n$ and $s$ denote the IR signal term, the noise term and the narrowband interferer, respectively. Consequently,

$$\lambda[k] = \left(\sqrt{\lambda^{(y)}[k]} + \sqrt{\lambda^{(n)}[k]} + \sqrt{\lambda^{(s)}[k]}\right)^2 \quad (16)$$

$$\beta[k] = \beta^{(y)}[k] + \beta^{(n)}[k] + \beta^{(s)}[k] \quad (17)$$

where $\lambda^{(y)}$ and $\beta^{(y)}$ denote the non-centrality parameter and the scale parameter that describe the distribution of the spectrogram of the signal $y$, i.e. $I^{(y)}[k]$. The same notation rule is applied to $\lambda^{(n)}$, $\lambda^{(s)}$, $\beta^{(n)}$ and $\beta^{(s)}$. In the following three subsections, $\lambda^{(y)}$, $\beta^{(y)}$, $\lambda^{(n)}$, $\beta^{(n)}$, $\lambda^{(s)}$, and $\beta^{(s)}$ are constructed.

### A. The IR signal

Let's assume that the IR signal term in the signal $f[m]$ is described by

$$y[m] = \sqrt{E_p}\sum_l a_l \delta[m - c_l - lN_h - \theta] \quad (18)$$

where $E_p$ denotes the energy per pulse. Furthermore, $\theta$ is a RV with distribution $P(\theta = x) = N_h^{-1}$ where $x \in \{0, 1, ..., N_h - 1\}$. Now, let's concentrate on the expectation

$$\mathrm{E}\{X[k]\} = \sum_{n=0}^{N-1} \mathrm{E}\{y[n]\}\cos\left(\frac{2\pi k}{N}n\right) \quad (19)$$

Based on the fact that the expectation $\mathrm{E}\{a_l\} = 0$, it is easy to verify that both $\mathrm{E}\{y[n]\}$ and thus $\mathrm{E}\{X[k]\}$ are equal to zero. Similarly, it can be proven that $\mathrm{E}\{Y[k]\} = 0$, such that the $\lambda^{(y)}[k] = 0$.

Let's continue with the derivation of the expectation

$$\begin{aligned}\mathrm{E}\{y[m]y[m+n]\} &= E_p \sum_{k,l} \mathrm{E}\{a_k a_l\} \\ &\cdot \mathrm{E}\{\delta[m - c_k - kN_h - \theta]\delta[m + n - c_l - lN_h - \theta]\}\end{aligned} \quad (20)$$

Based on the assumption that the pulse duration is shorter than the chip duration, the expectation $\mathrm{E}\{a_l a_m\} = \delta(m - n)$. Consequently, the right side of Eq.(20) can be simplified to

$$E_p \sum_k \mathrm{E}\{\delta[m - c_k - kN_h - \theta]\delta[m + n - c_k - kN_h - \theta]\} \quad (21)$$

Based on the stationarity of the signal $y[m]$, the expectation depends only on the value of $n$, i.e.

$$\mathrm{E}\{y[m]y[m+n]\} = \frac{E_p}{N_h}\delta[n] \quad (22)$$

Substitution of Eq.(22) into Eq.(13) leads to the spectrogram of the IR signal

$$\mathrm{E}\left\{I^{(y)}[k]\right\} = \frac{E_p N}{N_h} \quad (23)$$

which is equal to $\beta^{(y)}[k]$ since $\lambda^{(y)}[k]$ was found to be zero.

### B. White Gaussian noise in the spectrogram

The noise term in the signal $f[m]$ is described by

$$n[m] = \sum_l n_l \delta[m - c_l - lN_h] \quad (24)$$

Assuming a zero mean, white Gaussian noise signal at the demodulator input and based on the fact that the pulse duration is shorter as the chip duration, the expectation $\mathrm{E}\{n_l\} = 0$ and $\mathrm{E}\{n_l n_m\} = \sigma_n^2 \delta(l - m)$, where $\sigma_n^2$ denotes the noise variance. The problem is equivalent to the one solved in the previous subsection, such that

$$\begin{aligned}\lambda^{(n)}[k] &= 0 \\ \beta^{(n)}[k] &= \frac{\sigma_n^2 N}{N_h}\end{aligned} \quad (25)$$

### C. Narrowband signal in the spectrogram

Let us assume a narrowband interferer whose phase remains approximately constant over the spectrogram duration $NT_c$, such that the narrowband signal term,

$$\tilde{s}[m] = \sqrt{2E_s}\cos\left(\frac{2\pi k_n}{N}m + \varphi\right) \quad (26)$$

where $E_s$, $k_n$ and $\varphi$ denote the average power, frequency and phase of the narrowband interferer, respectively. To simplify matters, $k_n$ is assumed to have an integer value. Let's commence with the computation of

$$\mathrm{E}\left\{X^{(s)}[k]\right\} = \sum_{n=0}^{N-1} \tilde{s}[n]\mathrm{E}\{c[n]\}\cos\left(\frac{2\pi k}{N}n\right) \quad (27)$$

It is easy to verify that $\mathrm{E}\{c[n]\} = N_h^{-1}$, which after substitution of Eq.(26) into Eq.(27) leads to

$$\mathrm{E}\left\{X^{(s)}[k]\right\} = \frac{\sqrt{E_s}N}{N_h}\left(\delta\left[k - k_n + N\right] - \delta\left[k_n - k\right]\right)\cos(\varphi) \quad (28)$$

The same approach leads to the following expression for

$$\mathrm{E}\left\{Y^{(s)}[k]\right\} = \frac{\sqrt{E_s}N}{N_h}\left(\delta\left[k - k_n + N\right] - \delta\left[k_n - k\right]\right)\sin(\varphi) \quad (29)$$

Let's continue with the derivation of the expectation $\mathrm{E}\left\{I^{(s)}[k]\right\}$. The blanking code and the narrowband interferer are independent, such that

$$\mathrm{E}\{s[n+m]s[n]\} = \tilde{s}[n+m]\tilde{s}[n]\mathrm{E}\{c[n+k]c[n]\} \quad (30)$$

Since $\mathrm{E}\left\{I^{(s)}[k]\right\}$ is the DFT of $\mathrm{E}\{s[n+m]s[n]\}$ and a multiplication of two signals in the discrete time domain is equivalent to the circular convolution of both signals in the discrete Fourier domain, then

$$\mathrm{E}\left\{I^{(s)}[k]\right\} = \frac{1}{N^2} I^{(\tilde{s})}[k] \otimes \mathrm{E}\left\{I^{(c)}[k]\right\} \tag{31}$$

where $I^{(\tilde{s})}[k]$ and $I^{(c)}[k]$ denote the spectrogram of $\tilde{s}[m]$ and $c[m]$, respectively. It is easy to derive that

$$I^{(\tilde{s})}[k] = E_s N^2 \delta\left[N - k_n\right] + E_s N^2 \delta\left[k_n\right] \tag{32}$$

where the frequency $k$ is circular such that $\delta[-k] = \delta[N-k]$. It is straight-forward derivation leads to the autocorrelation function of the blanking code

$$\mathrm{E}\{c[n+k]c[n]\} = \frac{1}{N_h}\delta(k) + \frac{1}{N_h^2}\left(1 - \mathrm{tri}\left(\frac{k}{N_h}\right)\right) \tag{33}$$

where $\mathrm{tri}\,(x) = 1 - |x|$ if $-1 < x < 1$ and otherwise equal to zero, such that the expectation of its spectrogram

$$\mathrm{E}\left\{I^{(c)}[k]\right\} = \frac{N}{N_h} + \frac{N^2}{N_h^2}\delta[k] - \frac{N}{N_h^3}\left(\frac{\sin\left(N_h \frac{\pi k}{N}\right)}{\sin\left(\frac{\pi k}{N}\right)}\right)^2 \tag{34}$$

Again, the frequency $k$ is circular such that $\mathrm{E}\left\{I^{(c)}[-k]\right\}$=$\mathrm{E}\left\{I^{(c)}[N-k]\right\}$. Consequently, the expectation, including the effect of the periodic non-uniform sampling, becomes

$$\mathrm{E}\left\{I^{(s)}[k]\right\} = \frac{E_s}{N^2}\left(\mathrm{E}\left\{I^{(c)}[k-k_n]\right\} + \mathrm{E}\left\{I^{(c)}[k_n-k]\right\}\right) \tag{35}$$

Eq.(34) shows that the blanking code spreads the energy of the narrowband signal over the spectrogram, resulting in a higher false alarm probability as will be shown later on. Fortunately, the delta peak ensures that the frequency of the narrowband signal can still be identified. The level of distortion depends on the parameters $N$ and $N_h$. To conclude this subsection, let's summarize the results,

$$\begin{aligned}\beta^{(s)}[k] &= \frac{E_s}{2}\left(\mathrm{E}\left\{I^{(c)}[k-k_n]\right\} + \mathrm{E}\left\{I^{(c)}[k_n-k]\right\}\right) - \lambda^{(s)}[k] \\ \lambda^{(s)}[k] &= \frac{E_s N^2}{2N_h^2}\left(\delta\left[k - k_n + N\right] - \delta\left[k_n - k\right]\right)\end{aligned} \tag{36}$$

## IV. The Detector

As stated before, the event of the detection of a narrowband interferer at frequency $k$ occurs whenever the spectrogram $I_{n_0}[k]$ exceeds the threshold $\eta$. The probability of this event is given by

$$P[k] = P\{I_{n_0}[k] > \eta\} = \int_{\eta}^{\infty} p\left(x; \lambda[k]; \beta[k]; v[k]\right) dx \tag{37}$$

The event of a correct detection can only occur if $k = k_n$, such that the detection probability $P_d = P[k|k=k_n]$, otherwise a false alarm event has occurred, i.e. $P_f[k] = P[k|k \neq k_n]$. Here, the same threshold $\eta$ is used for every frequency and is chosen such that a Chi-squared distribution with $\lambda = 0$, $\nu = 2$ and a $\beta = \frac{1}{N}\sum_{k=0}^{N-1}\beta[k]$ leads the desired false alarm probability. If $\lambda \neq 0$, Eq.(37) must be evaluated numerically, otherwise it can be solved closed form [5].

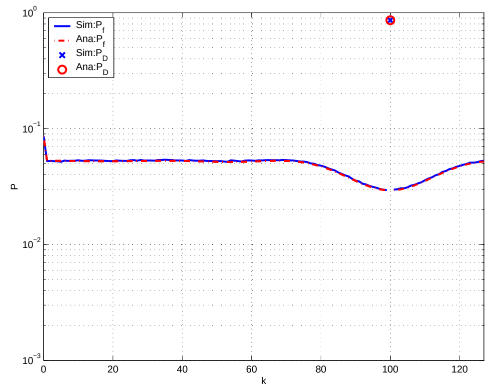


Fig. 1. The detection and false alarm probability of the detector for a desired false alarm probability of $5 \cdot 10^{-2}$

A narrowband interferer that is present in a spectrogram is most likely also present in the adjacent spectrogram. Consequently, the performance of the detector can be further improved by summing up $M$ spectrograms, i.e. the combined spectrogram

$$J_{n_0}[k] = \sum_{n=0}^{M-1} I_{n_0+nN}[k] \tag{38}$$

The individual spectrograms are computed over non-overlapping section of the signal $f[m]$ and are therefore statistically independent. Consequently, the combined spectrogram $J_{n_0}[k]$ will be a RV, which arises as a result of summing the squares of $M{\cdot}v$ i.i.d. Gaussian RV, where $v = 1$ if $k = 0$ and otherwise equal to two. Evidently, the combined spectrogram will be Chi-squared distributed as well, where its

$$\begin{aligned}\lambda^{(J)}[k] &= M\lambda[k] \\ \beta^{(J)}[k] &= M\beta[k] \\ v^{(J)}[k] &= Mv[k]\end{aligned} \tag{39}$$

Consequently, the expression for the detection and false alarm probability remains valid if multiple spectrograms are combined before the threshold detection is being applied.

## V. Performance Analysis

A performance evaluation of the narrowband interferer detector will be presented, where the detection and false alarm probability are considered to be a measure for the performance. To validate the correctness of the analysis, both probabilities have been compared with simulation results. The parameter values are $N = 512$, $N_h = 16$, $E_s = 1$, $\sigma_n^2 = 1$, $E_p = 1$ and $k_n = 100$. The performance is depicted in Fig.1 and Fig.2 for a desired false alarm probability of $5{\cdot}10^{-2}$ and $5{\cdot}10^{-3}$ , respectively. The X-axis denoted the frequency. The event of a correct detection can be found at the frequency with a value equal to 100 and the event of a false alarm at the other frequencies. The probability of both events can be found on the Y-axis.

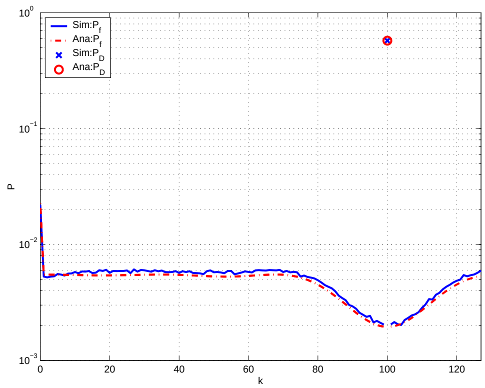


Fig. 2. The detection and false alarm probability of the detector for a desired false alarm probability of $5 \cdot 10^{-3}$

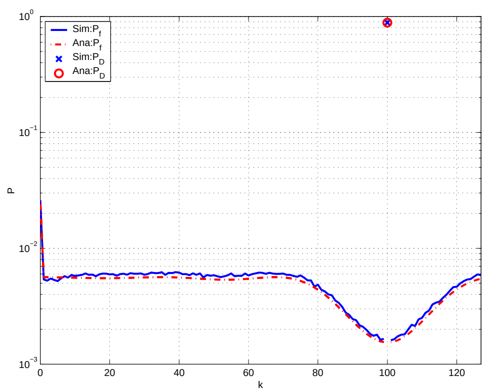


Fig. 3. The detection and false alarm probability of the detector for a desired false alarm probability of $5 \cdot 10^{-3}$ and $M = 2$

Both Fig.1 and Fig.2 indicate that the mathematical expressions for the detection and false alarm rate match the simulation results well, although the closed-form expressions seems to under-estimate the false alarm rate, which probably result from the assumption that $X$ and $Y$ are RVs with a normal distribution. Still the closed-form expressions give a good indication on the performance of the algorithm. A comparison of both figures reveals that a lower desired false alarm probability results in a decreased detection probability, meaning that the algorithm becomes less sensitive if a lower false alarm probability is desired. Both figures illustrate that the false alarm is not constant over the frequency. As stated before, the blanking code spreads a part of the energy of the narrowband signal over the spectrogram, but not uniformly. Consequently, the false alarm varies as function of $k$.

The sensitivity can be improved without increasing the false alarm probability by means of spectrogram combining. Here, the principle of spectrogram combining is validated and compared with the previous results. To ensure a fair comparison, the parameters have been kept constant. The desired false alarm rate has been set to $5 \cdot 10^{-3}$. The performance has been depicted in Fig.3 and Fig.4, where the former corresponds to $M = 2$ and the latter to $M = 4$.

A comparison of both figures confirm an improved sensitivity of the detector without increasing the false alarm probability. Furthermore, the simulation results validate the correctness of the closed-form expressions for both probabilities.

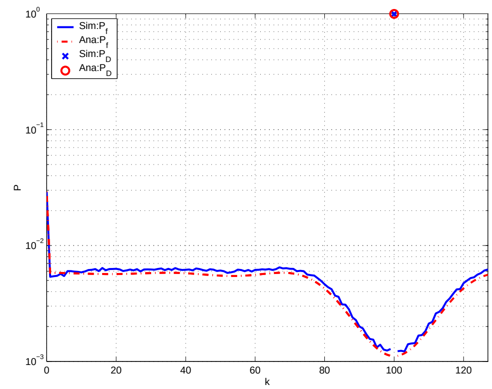


Fig. 4. The detection and false alarm probability of the detector for a desired false alarm probability of $5 \cdot 10^{-3}$ and $M = 4$

## VI. Conclusions and Open Issues

A method for the detection of narrowband interference including its center frequency has been presented for a time hopping impulse radio receiver. The impulse radio demodulator output is used as input for the narrowband detector, such that possibly no additional hardware is required. A spectrogram based technique combined with a threshold detector has been chosen as narrowband interferer detector. Exact closed form expressions for the detection and false alarm probability have been derived and validated by means of simulation. Furthermore, the performance is shown to be improved by combining several spectrograms before threshold detection.

Several topics critical to the applicability of the method remain open. For instance, the pulse duration was assumed to be shorter than the chip duration, which in several cases not the reality. A release of this assumption will modify the statistical properties of the spectrogram and thus the detector performance. Furthermore, the computation complexity has not been analyzed. Evidently, the fast Fourier transformation can reduce the computation complexity. Furthermore, the complexity may be further reduced, since the up-sampled demodulator output contains many zero valued samples.

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