WEAK INTERFERENCE DETECTION WITH SIGNAL CANCELLATION IN SATELLITE COMMUNICATIONS

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ABSTRACT

Interference is identified as a critical issue for satellite communication (SATCOM) systems and services. There is a growing concern in the satellite industry to manage and mitigate interference efficiently. While there are efficient techniques to monitor strong interference in SATCOM, weak interference is not so easily detected because of its low interference to signal and noise ratio (ISNR). To address this issue, this paper proposes and develops a technique which takes place on-board the satellite by decoding the desired signal, removing it from the total received signal and applying an Energy Detector (ED) in the remaining signal for the detection of interference. Different from the existing literature, this paper considers imperfect signal cancellation, examining how the decoding errors affect the sensing performance, derives the expressions for the probability of false alarm and provides a set of simulations results, verifying the efficiency of the technique.

Index Terms— Energy Detector, signal cancellation, truncated chi-squared distribution, interference detection, satellite communications.

1. INTRODUCTION

Interference is identified as a critical issue for the satellite communications industry, especially for the satellite owners/operators [1]-[2]. The proper management of interference can be carried out at different steps (interference monitoring, detection, classification, localisation and mitigation) [3]. However, in this paper, we focus on the detection of interference, particularly On-Board Interference Detection (OBID) by introducing a dedicated spectrum monitoring unit (SMU) within the satellite payload. Nevertheless, we should note that the developed techniques can be used for on-ground interference detection, as well. The advantages of OBID with respect to ground-based solutions are summarized as follows: i) simplification of the ground based stations in multibeam systems ii) faster reaction time and iii) avoidance of the additional downlink noise and possible distortions related to the satellite transponder [3]-[4].

One method for OBID is proposed in [4], where the adopted detection approach is the conventional Energy Detector (CED) [5]-[9]. This is an efficient technique, especially for strong interference scenarios. However, it faces challenges for the detection of weak interference because in this case, the ED is much more sensitive to the noise and desired signal power uncertainties, as well to the number of available samples. To overcome this issue, the idea of "ED with signal cancellation" is proposed in [10] by exploiting the frame structure and pilot symbols of the standards. This technique provides reliable detection of weak interfering signals under the assumption of adequate number of samples, namely pilots. However, sometimes the detection at low values of ISNR may require more samples than the number of pilots supported by the standard.

To address these concerns, in this paper, we propose a detection scheme based on the concept of "ED with signal cancellation", which does not require pilots symbols. It focuses on the data domain, decoding the desired data signal, removing it from the total received signal and applying an ED in the remaining signal for the detection of interference. There are similar works [11]-[14], nevertheless, [11] assumes perfect signal decoding, [12] considers a system which knows when the decoding is successful and only then it cancels the signal, while [13] and [14] assume that the remaining signal, after the signal or interference cancellation, follows a Gaussian distribution.

The contributions of this paper are summarized as follows: i) we propose a reliable scheme for the detection of weak interference applied on-board the satellite, based on the decoding and cancellation of the desired signal and ii) we examine how the imperfect signal cancellation, due to the decoding errors, affects the interference detection performance and derive analytical expressions for the probability of false alarm (P_{FA}). The latter expressions are based on the actual deterministic values of the remaining signal and not on their approximations by Gaussian distribution, as in [13]-[14].

The rest of this paper is organized as follows. Section 2 presents the system model. In Section 3, the derivation of the P_{FA} is described. Section 4 depicts the numerical results, and Section 5 concludes the paper.

Notation: Bold-face lower case letters are used to denote vectors. The superscript $(\cdot)^T$ represents the transpose of $(\cdot)$. $\|\cdot\|$ declares the standard vector norm, while $|\cdot|$ is the absolute value. $E\{\cdot\}$ and $V\{\cdot\}$ denote the expectation and the variance of $\{\cdot\}$, respectively and $\mathcal{R}\{\cdot\}$ and $\mathcal{I}\{\cdot\}$ represent the real and the imaginary part of $\{\cdot\}$, respectively.

2. SYSTEM MODEL

2.1. Signal Model

We consider a common satellite communication system , where the satellite, the desired Earth Station (ES) and the interferer are equipped with one antenna. Our goal is to detect the uplink radio frequency interference. Hence, the detection problem can be formulated as the following binary hypothesis test, which is a baseband symbol sampled model:

$$\mathcal{H}_0: \quad \mathbf{y}{=}h\mathbf{s}{+}\mathbf{w}, \tag{1}$$

$$\mathcal{H}_1: \quad \mathbf{y}{=}h\mathbf{s}{+}\mathbf{w}{+}\mathbf{i}, \tag{2}$$

where h denotes the scalar flat fading channel from the desired ES to the satellite, which is assumed to be known at the satellite receiver

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(i.e. estimated in advance) and it is assumed to be real after the phase compensation with channel power γ, $\mathbf{s} = [s(1) \ \cdots s(N)]^T$ denotes an $N \times 1$ vector, referred to as the signal transmitted by the desired ES with power P_s and it is a modulated signal, $\mathbf{i} = [i(1) \ \cdots i(N)]^T$ denotes an $N \times 1$ vector, referred to as the received signal from the interferer, $\mathbf{w} = [w(1) \ \cdots w(N)]^T$ denotes an $N \times 1$ vector referred to as the additive noise at the receiving antenna of the satellite, modelled as an independent and identically distributed (i.i.d.) complex Gaussian vector with zero mean and covariance matrix given by $E\left\{\mathbf{w}\mathbf{w}^H\right\} = \sigma_w^2 \mathbf{I}_N$, where $\mathbf{I}_N$ denotes an identity matrix of size N, and $\mathbf{y} = [y(1) \ \cdots y(N)]^T$ denotes an $N \times 1$ vector, referred to as the total received signal at the satellite.

2.2. Proposed algorithm

In this subsection, the various steps of the proposed algorithm are described. We should note that this technique needs a partially regenerative satellite (at least for the SMU), where the received signal can be demodulated.

Algorithm: *ED with imperfect signal cancellation (EDISC)*

Step 1: Decode the transmitted signal by the desired ES: $\hat{\mathbf{s}}$ denotes the decoded or estimated signal.
Step 2: Remove this estimated signal from the total received signal at the satellite: $\mathbf{y}' = \mathbf{y} - h\hat{\mathbf{s}}$.
Step 3: Apply the ED in the remaining signal as shown in (3)

$$T\left(\mathbf{y}'\right) = \left\|\mathbf{y}'\right\|^2 = \sum_{n=0}^{N-1} \left|y'(n)\right|^2 \begin{array}{l} < \varepsilon \rightarrow \mathcal{H}_0 \\ > \varepsilon \rightarrow \mathcal{H}_1 \end{array}, \tag{3}$$

where ε is a proper selected threshold responsible for the decision of presence/absence of interference.

This algorithm can be applied for any modulation scheme supported by DVB-S2X [15] standard (QPSK, 8PSK, 16APSK,...), but in this paper, we focus on QPSK modulated signals. However, for simplicity, we start our analysis considering a BPSK signal, which as shall be shown later, can be easily extended to QPSK scenario.

Applying the first and second step of our algorithm under the BPSK case, the hypothesis test of (1)-(2) can be reformulated as follows:

$$\mathcal{H}_{0_B} = \left\{ \begin{array}{l} \mathcal{H}_{00_B} : y'(n) = w(n), \\ \mathcal{H}_{01_B} : y'(n) = 2hs(n) + w(n), \end{array} \right. \tag{4}$$

$$\mathcal{H}_{1_B} = \left\{ \begin{array}{l} \mathcal{H}_{10_B} : y'(n) = i(n) + w(n), \\ \mathcal{H}_{11_B} : y'(n) = i(n) + 2hs(n) + w(n), \end{array} \right. \tag{5}$$

where $n = 0, 1, ..., N-1$, the index B denotes the BPSK scenario, $\mathcal{H}_{00_B}$ and $\mathcal{H}_{10_B}$ represent the hypothesis when the received signal is decoded correctly and the interference is absent and present, respectively, while $\mathcal{H}_{01_B}$ and $\mathcal{H}_{11_B}$ correspond to the wrong decoding case when the interference is absent and present, respectively.

Then, the final step is to use the ED of (3) in order to decide if the interference exists or not. We should mention that our detector is designed based on Neyman-Pearson criterion [16], hence, the detection threshold ε is calculated through the P_{FA}, which represents the scenario when we decide wrongly that the interference is present, while it is absent in the reality. Therefore, in the rest of this paper, we focus on the hypothesis $\mathcal{H}_0$ for the derivation of the P_{FA}.

3. PROBABILITY OF FALSE ALARM

As mentioned earlier, the performance of our proposed detector is examined through the P_{FA}. Hence, in this section we derive the P_{FA}, first for the BPSK and then for the QPSK scenario.

3.1. Probability of false alarm for BPSK signals

In this subsection, we derive the P_{FA} under the BPSK scenario as described by Theorem 1.

Theorem 1: *Consider a satellite with one receive antenna, which collects a large number of samples N. The satellite receiver decodes the received samples, removes them from the total received signal and applies an ED in the remaining signal. Then, the P_{FA_B} is given by*

$$P_{FA_B} = \sum_{k=0}^{N} \left(\begin{array}{c} N \\ k \end{array} \right) P_{k_B} P_{e_B}^k \left(1 - P_{e_B}\right)^{N-k}, \tag{6}$$

where k denotes the number of wrong decoded bits, P_{e_B} is the probability of bit error for BPSK [17] and P_{k_B} is the probability of false alarm for the case that k bits are decoded wrongly which can be approximated as follows

$$P_{k_B} = Q\left(\frac{\varepsilon - \mu_{H_{0_B}}}{\sqrt{V_{H_{0_B}}}}\right), \tag{7}$$

where $\mu_{\mathcal{H}_{0_B}}$ and $V_{\mathcal{H}_{0_B}}$ are the mean and variance of the test statistic $T\left(\mathbf{y}' \left|\mathcal{H}_{0_B}\right.\right)$, respectively, which are also related to k as we show in the following proof.

Proof: Considering the fact that the number of wrong bits ranges from 0 to N, the first part of the theorem, can be proved by applying the concept of Binomial distribution [18]. For the second part, the probability of false alarm P_{k_B} of the ED of (3) is determined by $P_{k_B}(\varepsilon) = \Pr\left(T\left(\mathbf{y}'\right) > \varepsilon | \mathcal{H}_{0_B}\right)$, where it becomes clear that the definition of the distribution of the test statistic $T\left(\mathbf{y}' \left|\mathcal{H}_{0_B}\right.\right)$ is necessary. Unlike the CED, in our case we do not deal with the sum of N squared Gaussian random variables, but with truncated [19]-[20] ones. The issue to be solved in this case, is that the distribution of the sum of N truncated central or non-central chi-squared variables is not tractable. Therefore, the question to be answered is if we can approximate the distribution of the test statistic $T\left(\mathbf{y}' \left|\mathcal{H}_{0_B}\right.\right)$ using the central limit theorem (CLT) [21].

The test statistic $T\left(\mathbf{y}' \left|\mathcal{H}_{0_B}\right.\right)$ consists of the following three categories of variables: i) all the symbols have decoded correctly, ii) all the symbols have decoded wrongly and iii) some symbols have decoded correctly and some other have decoded wrongly. It is clear that the CLT can be applied (under the assumption of large number of samples) for the first and second case, because both involve a sequence of i.i.d random variables. However, the third case includes a sequence that is independent but not identically distributed. Nevertheless, even in this case, we can use the CLT because Lyapunov's and Lindeberg's conditions for non-identical variables [22] are satisfied. Therefore, the mean and variance of $T\left(\mathbf{y}' \left|\mathcal{H}_{0_B}\right.\right)$ can be respectively expressed by $\mu_{\mathcal{H}_{0B}} = (N-k)\mu_{\mathcal{H}_{00B}} + k\mu_{\mathcal{H}_{01B}}$ and $V_{\mathcal{H}_{0B}} = (N-k)V_{\mathcal{H}_{00B}} + kV_{\mathcal{H}_{01B}}$, where $\mu_{\mathcal{H}_{00_B}}$, $\mu_{\mathcal{H}_{01_B}}$, $V_{\mathcal{H}_{00_B}}$ and $V_{\mathcal{H}_{01_B}}$ are the mean and variance of the test statistic $T\left(y' \left|\mathcal{H}_{00_B}\right.\right)$ and $T\left(y' \left|\mathcal{H}_{01_B}\right.\right)$, respectively, where y' means only one sample. ∎

Hence, we need to find the mean and variance for $T\left(y' \left|\mathcal{H}_{00_B}\right.\right)$ and $T\left(y' \left|\mathcal{H}_{01_B}\right.\right)$. However, first, we should introduce the following three lemmas, which will be used in the calculation of these parameters.

Lemma 1: *The mean $\mu_c^{[a,b]}$ and variance $V_c^{[a,b]}$ of a central chi-squared variable with one degree of freedom, truncated to the interval $[a,b]$ where $0 \le a \le b \le \infty$ is given as follows:*

$$\mu_c^{[a,b]} = 1 + 2\left[\frac{af_{\chi_1^2}(a) - bf_{\chi_1^2}(b)}{F_{\chi_1^2}(b) - F_{\chi_1^2}(a)}\right] \tag{8}$$

and

$$V_c^{[a,b]} = 2 - 4\left[\frac{af_{\chi_1^2}(a) - bf_{\chi_1^2}(b)}{F_{\chi_1^2}(b) - F_{\chi_1^2}(a)}\right]^2 + 2\left[\frac{a^2 f_{\chi_1^2}(a) + af_{\chi_1^2}(a) - b^2 f_{\chi_1^2}(b) - bf_{\chi_1^2}(b)}{F_{\chi_1^2}(b) - F_{\chi_1^2}(a)}\right]. \tag{9}$$

where $F_{\chi_1^2}$ and $f_{\chi_1^2}$ denote respectively the cumulative distribution function (cdf) and the probability density function (pdf) of a central chi-squared variable with one degree of freedom.

Lemma 2: *The mean $\mu_c^{[a,b]}$ and variance $V_c^{[a,b]}$ of a non-central chi-squared variable, each with one degree of freedom and non-centrality parameter λ, truncated to the interval $[a,b]$ where $0 \le a \le b < \infty$ is given as follows:*

$$\mu_{nc}^{[a,b]} = \frac{F_{\chi_{3,\lambda}^2}(b) - F_{\chi_{3,\lambda}^2}(a) + \lambda\left[F_{\chi_{5,\lambda}^2}(b) - F_{\chi_{5,\lambda}^2}(a)\right]}{F_{\chi_{1,\lambda}^2}(b) - F_{\chi_{1,\lambda}^2}(a)} \tag{10}$$

and

$$V_{nc}^{[a,b]} = \frac{3\left[F_{\chi_{5,\lambda}^2}(b) - F_{\chi_{5,\lambda}^2}(a)\right] + 6\lambda\left[F_{\chi_{7,\lambda}^2}(b) - F_{\chi_{7,\lambda}^2}(a)\right]}{F_{\chi_{1,\lambda}^2}(b) - F_{\chi_{1,\lambda}^2}(a)} + \frac{\lambda^2\left[F_{\chi_{9,\lambda}^2}(b) - F_{\chi_{9,\lambda}^2}(a)\right]}{F_{\chi_{1,\lambda}^2}(b) - F_{\chi_{1,\lambda}^2}(a)} - \left(\mu_{nc}^{[a,b]}\right)^2, \tag{11}$$

where $F_{\chi_{1,\lambda}^2}$ denotes the cdf of a non-central chi-squared variable with one degree of freedom and non-centrality parameter λ.

Lemma 3: *The mean $\mu_{nc\,\inf}^{[-\infty,a]}$ and $V_{nc\,\inf}^{[-\infty,a]}$ of a squared Gaussian variable, where the Gaussian variable is truncated to the interval $[a,\infty]$ or $[-\infty,\alpha]$, with mean μ and variance σ^2, is given as follows:*

$$\mu_{nc\,\inf}^{[-\infty,a]} = \mu^2 - 2\mu\sigma\frac{f_x(d)}{F_x(d)} + \sigma^2\left(1 - d\frac{f_x(d)}{F_x(d)}\right), \tag{12}$$

$$V_{nc\,\inf}^{[-\infty,a]} = \mu^4 - 4\mu^3\sigma\frac{f_x(d)}{F_x(d)} + 6\mu^2\sigma^2\left(1 - d\frac{f_x(d)}{F_x(d)}\right) + 4\mu\sigma^3\left(-d^2\frac{f_x(d)}{F_x(d)} - 2\frac{f_x(d)}{F_x(d)}\right) + \sigma^4\left(-d^3\frac{f_x(d)}{F_x(d)} - 3d\frac{f_x(d)}{F_x(d)} + 3\right) - \left(\mu_{nc\,\inf}^{[-\infty,a]}\right)^2, \tag{13}$$

where $d = a - \mu$ and F_x, f_x denote respectively the cdf and pdf of the normal distribution.

Proof: The proof of Lemma 1, 2 and 3 is not presented in this paper due to lack of space, but they can be proved following the definition of truncated distributions ([23]-[25]). ■

We can now develop Theorem 2 for the derivation of $\mu_{\mathcal{H}_{00_B}}$, $\mu_{\mathcal{H}_{01_B}}$,$V_{\mathcal{H}_{00_B}}$ and $V_{\mathcal{H}_{01_B}}$.

Theorem 2: *Consider that our proposed algorithm of ED with signal cancellation is applied for a large number of samples N. Then, the mean and the variance of $T\left(y' \left|\mathcal{H}_{00_B}\right.\right)$ can be respectively given by*

$$\mu_{\mathcal{H}_{00_B}} = \mu_c^{[0,P_t]} P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]} + \mu_c^{[P_t,\infty]} P_{w_n}^{\left[\sqrt{P_t},\infty\right]} \tag{14}$$

and

$$\begin{aligned} V\left\{\mathcal{H}_{00_B}\right\} &= V_c^{[0,P_t]} P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]} + V_c^{[P_t,\infty]} P_{w_n}^{\left[\sqrt{P_t},\infty\right]} + \\ &+ \left(\mu_c^{[P_t,\infty]}\right)^2 \left(1 - P_{w_n}^{\left[\sqrt{P_t},\infty\right]}\right) P_{w_n}^{\left[\sqrt{P_t},\infty\right]} + \left(\mu_c^{[0,P_t]}\right)^2 \times \\ &\times \left(1 - P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]}\right) \times P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]} - 2\mu_c^{[0,P_t]} \mu_c^{[P_t,\infty]} \times \\ &\times P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]} P_{w_n}^{\left[\sqrt{P_t},\infty\right]}, \end{aligned} \tag{15}$$

while the mean and the variance of $T\left(y' \left|\mathcal{H}_{01_B}\right.\right)$ can be respectively expressed as follows

$$\mu_{\mathcal{H}_{01_B}} = \mu_{nc}^{[0,P_t]} P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]} + \mu_{ncinf}^{\left[3\sqrt{P_t},\infty\right]} P_{w_n}^{\left[3\sqrt{P_t},\infty\right]} \tag{16}$$

and

$$\begin{aligned} V\left\{\mathcal{H}_{01_B}\right\} &= V_{nc}^{[0,P_t]} P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]} + V_{ncinf}^{\left[3\sqrt{P_t},\infty\right]} P_{w_n}^{\left[3\sqrt{P_t},\infty\right]} + \\ &+ \left(\mu_{ncinf}^{\left[3\sqrt{P_t},\infty\right]}\right)^2 \left(1 - P_{w_n}^{\left[3\sqrt{P_t},\infty\right]}\right) P_{w_n}^{\left[3\sqrt{P_t},\infty\right]} + \left(\mu_{nc}^{[0,P_t]}\right)^2 \times \\ &\times \left(1 - P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]}\right) P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]} - 2\mu_{nc}^{[0,P_t]} \mu_{ncinf}^{\left[3\sqrt{P_t},\infty\right]} \times \\ &\times P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t},\right]} P_{w_n}^{\left[3\sqrt{P_t},\infty\right]}, \end{aligned} \tag{17}$$

where $P_t = \gamma P_s$ and $\mu_c^{[0,P_t]}$, $V_c^{[0,P_t]}$, $\mu_c^{[P_t,\infty]}$ and $V_c^{[P_t,\infty]}$, $\mu_{nc}^{[0,P_t]}$, $V_{nc}^{[0,P_t]}$, $\mu_{ncinf}^{[P_t,\infty]}$, $V_{ncinf}^{[P_t,\infty]}$ are defined by Lemma 1, 2 and 3, while $P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]}$ and $P_{w_n}^{\left[\sqrt{P_t},\infty\right]}$, $P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]}$, $P_{w_n}^{\left[3\sqrt{P_t},\infty\right]}$ are given by $P_{w_n}^{\left[-\sqrt{P_t},\sqrt{P_t}\right]} = \frac{P\left(-\sqrt{P_t} \le w[n] \le \sqrt{P_t}\right)}{P_{\mathrm{cB}}}$, $P_{w_n}^{\left[\sqrt{P_t},\infty\right]} = \frac{P\left(w[n] \ge \sqrt{P_t}\right)}{P_{\mathrm{cB}}}$, $P_{w_n}^{\left[\sqrt{P_t},3\sqrt{P_t}\right]} = \frac{P\left(\sqrt{P_t} \le w[n] \le 3\sqrt{P_t}\right)}{P_{\mathrm{cB}}}$, $P_{w_n}^{\left[3\sqrt{P_t},\infty\right]} = \frac{P\left(w[n] \ge 3\sqrt{P_t}\right)}{P_{\mathrm{cB}}}$ and P_{cB} is the probability of correct decision for BPSK.

Proof: The proof is given in Appendix A. ■

All the parameters of (6) have been defined. However, the calculation of the detection threshold ε, through (6), may be complicated, particularly as the number of samples increases. Nevertheless, the probability of false alarm can be approximated by

$$P_{FA_{B_a}} = Q\left(\frac{\varepsilon - N\left(1 - P_{e_B}\right)\mu_{\mathcal{H}_{00_B}} - NP_{e_B}\mu_{\mathcal{H}_{01_B}}}{\sqrt{N\left(1 - P_{e_B}\right)V_{\mathcal{H}_{00_B}} + NP_{e_B}V_{\mathcal{H}_{01_B}}}}\right), \tag{18}$$

where the index B_a denotes approximation under the BPSK scenario, and hence, this equation approximates and simplifies (6), based on the fact that for a large number of samples, the expected number of correct and wrong decoded bits is $N(1 - P_{e_B})$ and NP_{e_B}, respectively. Now, the calculation of the threshold ε is straightforward, based on the inverse function of the $P_{FA_{B_a}}(\cdot)$.

3.2. Probability of false alarm for QPSK signals

In the previous subsection, we derived the probability of false alarm under the BPSK scenario. Now, the extension of (6) to QPSK case is straightforward and it is given as follows:

$$P_{FA_Q} = \sum_{k=0}^{2N} \left(\begin{array}{c} 2N \\ k \end{array} \right) P_{k_Q} P_{e_Q}^{k} \left(1 - P_{e_Q}\right)^{2N-k}, \quad (19)$$

where $P_{k_Q} = P_{k_B}$ and $P_{e_Q} = P_{e_B}$. Hence, the only difference with (6) is the factor 2, due to the fact that a QPSK signal constitutes of two orthogonal BPSK ones. From the other side, the approximated P_{FA} of (18) can be expressed as follows:

$$P_{FA_{Q_a}} = Q\left(\frac{\varepsilon - a\mu_{\mathcal{H}_{00_Q}} - b\left(\mu_{\mathcal{H}_{01_Q}} + \mu_{\mathcal{H}_{02_Q}}\right) - c\mu_{\mathcal{H}_{03_Q}}}{\sqrt{aV_{\mathcal{H}_{00_Q}} - b\left(V_{\mathcal{H}_{01Q}} + V_{\mathcal{H}_{02_Q}}\right) - cV_{\mathcal{H}_{03_Q}}}}\right) \quad (20)$$

where $a = \left(1 - P_{e_Q}\right)^2$, $b = \left(1 - P_{e_Q}\right) P_{e_Q}$, $c = P_{e_Q}^2$, the index Q denotes the QPSK scenario, P_{e_Q} is the probability of bit error for QPSK and is the same as for BPSK, $\mathcal{H}_{00_Q}$ denotes that both the real and imaginary part are decoded correctly, $\mathcal{H}_{01_Q}$ means that the real part is decoded wrongly and the imaginary part is decoded correctly, $\mathcal{H}_{02_Q}$ means that the real part is decoded correctly and the imaginary part is decoded wrongly, while $\mathcal{H}_{03_Q}$ denotes that both the real and imaginary part are decoded wrongly. Furthermore, we can easily see that $\mu_{\mathcal{H}_{00_Q}} = 2\mu_{\mathcal{H}_{00_B}}$ and $V_{\mathcal{H}_{00_Q}} = 2V_{\mathcal{H}_{00_B}}$, $\mu_{\mathcal{H}_{01_Q}} = \mu_{\mathcal{H}_{00_B}} + \mu_{\mathcal{H}_{01_B}}$ and $V_{\mathcal{H}_{01_Q}} = V_{\mathcal{H}_{00_B}} + V_{\mathcal{H}_{01_B}}$, $\mu_{\mathcal{H}_{02_Q}} = \mu_{\mathcal{H}_{00_B}} + \mu_{\mathcal{H}_{01_B}}$ and $V_{\mathcal{H}_{02_Q}} = V_{\mathcal{H}_{00_B}} + V_{\mathcal{H}_{01_B}}$, and finally, $\mu_{\mathcal{H}_{03_Q}} = 2\mu_{\mathcal{H}_{01_B}}$ and $V_{\mathcal{H}_{03_Q}} = 2V_{\mathcal{H}_{01_B}}$.

4. NUMERICAL RESULTS

In this section, we present a numerical result to show the detection performance of the proposed detection scheme, namely EDISC. We will compare this scheme with CED and with "ED with perfect signal cancellation (EDPSC)" as both given in [4]. For simplicity, throughout this section, we assume that the interference follows a Gaussian distribution with zero mean and variance σ_i^2, while the transmitted signal from the desired ES is QPSK modulated. 10000 Monte Carlo simulations are carried out, the detection threshold is set such that $P_{FA} = 0.1$ and the channel power is taken as unit.

Figures 1 depicts the probability of detection versus the ISNR comparing the aforementioned three techniques: i) CED, ii) EDPSC and iii) EDISC. We can notice that the proposed ED presents a sightly degraded detection performance than the one of the EDPSC approach, but significantly better that the one of the CED approach, because this desired signal cancellation concept improves the ISNR and thus, enhances the detection reliability of the system. Furthermore, we can see that the detection performance, based on a threshold determined through the approximated P_{FA}, performs closely to the more accurate one, verifying the reliability of (20).

5. CONCLUSION

In this paper, we proposed an interference detection scheme based on the ED with imperfect signal cancellation in the data domain, assuming a perfectly known channel and we examined how the decoding errors affect the detection performance measures. Furthermore, we derived the theoretical expressions for the probability of false alarm, considering digitally modulated signals (i.e. BPSK, QPSK). Finally, the numerical results showed that our proposed technique outperforms the conventional ED, verifying the efficiency of the proposed technique. Future works include extension of the technique to higher

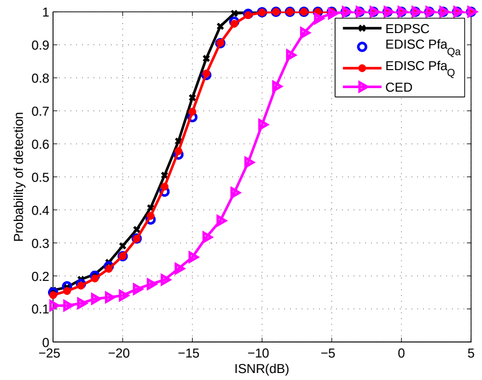

Fig. 1: Pobability of detection versus the ISNR under the QPSK scenario for $N = 100$, $E_s = 7$ dB and $\sigma_w^2 = 0$ dB.

modulation levels, as well as, considering the effect of channel estimation errors on the detection reliability.

A. APPENDIX
DERIVATION OF THE MEAN $\mu_{\mathcal{H}_{00_B}}$

The derivation of the mean $\mu_{\mathcal{H}_{00_B}}$ is based on the law of the total expectation [26] and it can be written as follows

$$E\left\{|y'|^2|\mathcal{H}_{00_B}\right\} = E\left\{|y'|^2 \left| s = +\sqrt{P_s}\,,\ \hat{s} = +\sqrt{P_s}\right.\right\} \times \\ \times P\left(s = +\sqrt{P_s}\right) P\left(\hat{s} = +\sqrt{P_s}\,|c_B\right) + E\left\{\,|y|^2\right| \\ \left| s = -\sqrt{P_s}, \hat{s} = -\sqrt{P_s}\,\right\} P\left(s = -\sqrt{P_s}\right) P\left(\hat{s} = -\sqrt{P_s}\,|c_B\right) \quad (21)$$

Furthermore, the expression of (21) can be further expressed by

$$E\left\{|y'|^2|\mathcal{H}_{00_B}\right\} = \\ = \frac{1}{2}E\left\{|y'|^2\,|w \geq -\sqrt{P_t}\right\} P\left(w \geq -\sqrt{P_t}\,|c_B\right) \\ + \frac{1}{2}E\left\{|y'|^2\,|w \leq +\sqrt{P_t}\right\} P\left(w \leq +\sqrt{P_t}\,|c_B\right), \quad (22)$$

where $P\left(s = \sqrt{P_s}\right) = P\left(s = -\sqrt{P_s}\right) = \frac{1}{2}$. But, the remaining noise under $\mathcal{H}_{00_B}$ does not follow any more a normal distribution, but a truncated normal distribution in the following intervals: i) $w \leq -\sqrt{P_t}$, ii) $-\sqrt{P_t} \leq w \leq \sqrt{P_t}$ and iii) $w \geq \sqrt{P_t}$. Thus, exploiting the symmetry of the BPSK constellation, (22) is given by

$$E\left\{|y'|^2|\mathcal{H}_{00_B}\right\} = \\ = E\left\{|y'|^2\left|-\sqrt{P_t} \leq w \leq \sqrt{P_t}\right.\right\} P\left(-\sqrt{P_t} \leq w \leq \sqrt{P_t}\,|c_B\right) \\ + E\left\{|y'|^2\,|w \geq \sqrt{P_t}\right\} P\left(w \geq +\sqrt{P_t}\,|c_B\right), \quad (23)$$

where $P\left(-\sqrt{P_t} \leq w \leq \sqrt{P_t} \leq \sqrt{P_t}\,|c_B\right) = \frac{P\left(-\sqrt{P_t} \leq w \leq \sqrt{P_t}\right)}{P_{c_B}}$ and $P\left(w \geq +\sqrt{P_t}\,|c_B\right) = \frac{P\left(w \geq \sqrt{P_t}\right)}{P_{c_B}}$ according to Bayes' Theorem [27]. Then, after some manipulations, (23) takes its final shape, which is given by (14).

The same methodology can be used for the proof of $\mu_{\mathcal{H}_{01_B}}$, $V_{\mathcal{H}_{00_B}}$ and $V_{\mathcal{H}_{01_B}}$. Note that the proof of the variances is based on the law of total variance [26].

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