

Citation for the published version:

Simpson, O., Sun, Y., Abdulkadir, Y., & El Cheikh, M. (2018). Spectrum Sensing of DVB-T2 Signals in Multipath Channels for Cognitive Radio Networks. European Wireless 2018 - 24th European Wireless Conference.

Document Version: Accepted Version

Link to the final published version available at the publisher:

<https://ieeexplore.ieee.org/document/8385554>

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Spectrum Sensing of DVB-T2 Signals in Multipath Channels for Cognitive Radio Networks

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Abstract— In this paper, spectrum sensing of digital video broadcasting-second generation terrestrial (DVB-T2) signals in different fading environments with energy detection (ED) is considered. ED is known to achieve an increased performance among low computational complexity detectors, but it is susceptible to noise uncertainty. By taking into consideration the edge pilot and scattered pilot periodicity in DVB-T2 signals, a low computational complex noise power estimator is proposed. It is shown analytically that the choice of detector depends on the environment, the detector requirements, the available prior knowledge and with the noise power estimator. Simulation confirm that with the noise power estimator, ED significantly outperforms the pilot correlation-based detectors. Simulation also show that the proposed scheme enables ED to obtain increased detection performance in fading channels.

Keywords — spectrum sensing; cognitive radio; DVB-T2; noise estimation

I. INTRODUCTION

The NEXT generation wireless standards, for instance fifth generation (5G) cellular networks will further transform the way information is accessed and communicated. At present the vast number of different wireless devices and technologies available, the rapid increase in the number of wireless subscribers, the introduction of new applications and the constant demand for higher data rates are all reasons for the radio frequency spectrum becoming more saturated. Each new access technology puts additional burden on the already constrained spectrum resources. This advances calls for systems and devices that are aware of their neighbouring radio environment, hence facilitating flexible, efficient, and reliable operation and utilization of the available spectral resources. Consequently, cognitive radio (CR) has been recognized as one of the capable technologies to alleviate the issue of wireless spectrum shortage [1]. CR archives this through the use of opportunistic spectrum sharing (OSS) [2] whereby secondary users (SUs) (unlicensed users) are allowed to opportunistically share the spectrum of primary users (PU) (licensed users) without causing any adverse impact on the PU transmission.

Spectrum sensing is a key enabling technology of CR. The techniques of sensing are commonly classified into matched filter detection (MFD), energy detection (ED), feature detection (FD) and waveform based detection (WFD) [3]. Of these algorithms, ED, also known as radiometry, is commonly used as a spectrum sensing technique due to its low computational complexity and hardware simplicity [3-7]. Additionally, it does not require prior information about the primary signal. ED is

capable of sensing various types of signals with varying characteristic, e.g. orthogonal frequency division multiplexing (OFDM) signals.

IEEE 802.22 is a standard for wireless regional area network (WRAN) using white spaces in the television (TV) frequency spectrum, where spectrum sensing is included as a mandatory feature [8]. The IEEE 802.22 standard is intended for cognitive operation in the digital TV bands, among others. According to the standard, a SU must be able to detect a PU digital video broadcasting (DVB-T) signal with probability of detection (P_d) at least 90% and probability of false alarm (P_f) no more than 10% at -22.2 dB signal-noise-ratio (SNR) [8]. CR operations in the DVB-T spectrum have also been considered in other research literature [9-11]. In Cognitive radio network's (CRN), DVB-T signals are one of the most important signal type used by PUs in TV bands [12]. This paper focuses on spectrum sensing of a digital video broadcasting-second generation terrestrial (DVB-T2) signals, a vital functionality in CR. DVB-T2 is the world's most advanced digital terrestrial television (DTT) system, offering more robustness, flexibility and 50% more efficiency than any other DTT system [13].

Aimed at sensing signals such as DVB-T2 signals, correlation-based detectors were proposed by exploiting correlation cyclic prefix or pilot [14,15]. The exact noise power in these detectors are replaced with the power of received signal which is impracticable because the received signal is likely highly to contain the primary signals. The detectors are robust to noise uncertainty, but in comparison with the perfect ED [16], they exhibit considerable performance degradation of about 8 ~ 14dB [15]. In this paper, using ED with a precise estimate of noise power in Multipath channel instead of using correlation-based detectors for DVB-T2 spectrum sensing is proposed. By taking into consideration the edge pilot and scattered pilot periodicity in the DVB-T2 signals, a novel low complexity noise power estimator which is able to obtain a precise noise power estimate from the received signals without regard consideration of the presence of the primary DVB-T2 signals is proposed, the signification of carrier frequency offsets (CFO's) is eliminated in the noise power estimate and frequency synchronization is not required. Due to the edge pilot and scattered pilot periodicity in DVB-T2, time synchronization is not required either. Simulation results show that the proposed noise power estimator enables ED to obtain a reliable detection performance. The rest of this paper is organized as follows: Section II describes the system model. In Section III, the proposed ED noise power estimation for DVB-T2 signals is presented. In Section IV, the computational complexity of the proposed system is analysed. Section V, presents Monte-Carlo

simulation results concerning the performance of the proposed algorithms. Conclusions are discussed in section VI.

II. System Model

A. DVB-T2 Signal

DVB-T2 is the world's most advanced DTT system, offering more robustness, flexibility and 50% more efficiency than any other DTT system [13]. It supports SD, HD, UHD, mobile TV, radio, or any combination thereof [14]. DVB-T2 signals are more resilient against certain types of interference than DVB-T. Since its publication in 1997, over 70 countries have deployed DVB-T services and 69 countries have now adopted or deployed DVB-T2 [12]. This well-established standard benefits from massive economies of scale and very low receiver prices. Due to the European analogue switch-off and increasing scarcity of spectrum, DVB drew up commercial requirements for a more spectrum-efficient and updated standard. DVB-T2 easily fulfils these requirements, including increased capacity, robustness and the ability to reuse existing reception antennas [15]. From a spectrum sensing point of view, important DVB-T2 parameters (see Table 1) are represented by: channel bandwidth (that ranges from 1.7 to 10 MHz), the OFDM Cyclic Prefix (CP) length (that ranges from 1/128 to 1/4 of the OFDM symbol length), and the presence of OFDM pilots (continual and scattered). DVB-T2 uses the same error correction coding as used in digital video broadcasting satellite second generation (DVB-S2) and digital video broadcasting cable second generation DVB-C2: LDPC (Low Density Parity Check) coding combined with BCH (Bose-Chaudhuri-Hocquengham) coding, offering a very robust signal. The number of carriers, guard interval sizes and pilot signals can be adjusted, so that the overheads can be optimised for any target transmission channel. The presence of pre-determined patterns in the transmitted DVB-T2 signal determines the cyclostationary property shown by OFDM signals. DVB-T2 uses OFDM modulation with a large number of sub-carriers delivering a robust signal, and offers a range of different modes, making it a very flexible standard.

Table 1. Main parameter of DVB-T2

	DVB-T2
FEC	LDPC + BCH 1/2, 3/5, 2/3, 3/4, 4/5, 5/6
Modes	QPSK, 16QAM, 64QAM, 256QAM
Guard interval	1/4, 19/256, 1/8, 19/128, 1/16, 1/32, 1/128
FFT size	1K, 2K, 4K, 8K, 16K, 32K
Scattered Pilots	1%, 2%, 4%, 8% of total
Continual Pilots	0.4%-2.4% (0.4%-0.8% in 8K-32K)
Bandwidth	1.7, 5, 6, 7, 8, 10 MHz
Typical data rate (UK)	40 Mbit/s
Max. data rate (@20 dB C/N)	45.5 Mbit/s (using 8M Hz)
Required C/N ratio (@24 Mbit/s)	10.8 dB

B. Problem Formulation and Signal model

As a common assumption in the literature on spectrum sensing, the primary signal is modeled as a Gaussian process. In [11], it is shown that in the case of DVB-T2 signals, this assumption is well motivated.

The problem of spectrum sensing is to decide whether there is a signal transmitted or not. That is, we wish to discriminate between the following two hypotheses:

$$\mathcal{H}_0 : r(k) = n(k) \quad k = 0, 1, \dots, K-1 \quad (1)$$

$$\begin{aligned} \mathcal{H}_1 : r(k) &= \eta e^{j2\pi\Delta_f k} x(k) + n(k) \\ &= \eta e^{j2\pi\Delta_f k} \sum_{m=0}^{M-1} h(m)s(k-m) + n(k) \end{aligned} \quad (2)$$

$k = 0, \dots, K-1$

where the hypotheses \mathcal{H}_0 and \mathcal{H}_1 correspond to the absence and presence of a DVB-T2 signal, respectively, k is the time index, Δ_f denotes the unknown CFO, $n(k)$ denotes the circularly symmetric complex Gaussian white noise with mean zero and variance σ_n^2 , i.e., $n(k) \sim \mathcal{CN}(0, \sigma_n^2)$; and DVB-T2 signal $s(k)$ arrives at the SU through a length- M multipath channel, resulting in the received primary signal $x(k)$. The DVB-T2 signal $s(k)$ consists of edge pilot signal, scattered pilot and data signals as shown in Figure 1. The time domain signal $s(k)$ can be represented as

$$s(k) = s_{ep}(k) + s_p(k) + s_d(k) \quad (3)$$

where $s_{ep}(k)$, $s_p(k)$ and $s_d(k)$ denote the edge pilot signal, scattered pilot and data signal, respectively. For a larger number of sub-carriers, both $s_{ep}(k)$, $s_p(k)$ and $s_d(k)$ are approximately Gaussian distributed. In Figure 1, PP2 has a repetition period of twelve OFDM symbols, while $s_d(k)$ has no periodicity [13]. Assuming that the sampling frequency is an integer multiple of OFDM symbol rate and twelve OFDM symbols include N even number samples, we have $s_p(k) = s_p(k+N)$. Using (3), the received primary signal $x(k)$ can also be divided into three parts

$$\begin{aligned} x(k) &= x_{ep}(k) + x_p(k) + x_d(k) \\ &= \sum_{m=0}^{M-1} h(m)s_{ep}(k-m) + \sum_{m=0}^{M-1} h(m)s_p(k-m) + \\ &\quad \sum_{m=0}^{M-1} h(m)s_d(k-m) \end{aligned} \quad (4)$$

It can be shown that $x_{ep}(k)$ and $x_p(k)$ has a period of N , i.e., $x_{ep}(k) = x_{ep}(k+N)$ and $x_p(k) = x_p(k+N)$ respectively, while $x_d(k)$ has no periodicity. In addition, $x_{ep}(k)$, $x_p(k)$ and $x_d(k)$ follow a normal Gaussian distribution. Let σ_{ep}^2 , σ_p^2 and σ_d^2 denote the power of $x_{ep}(k)$, $x_p(k)$ and $x_d(k)$,

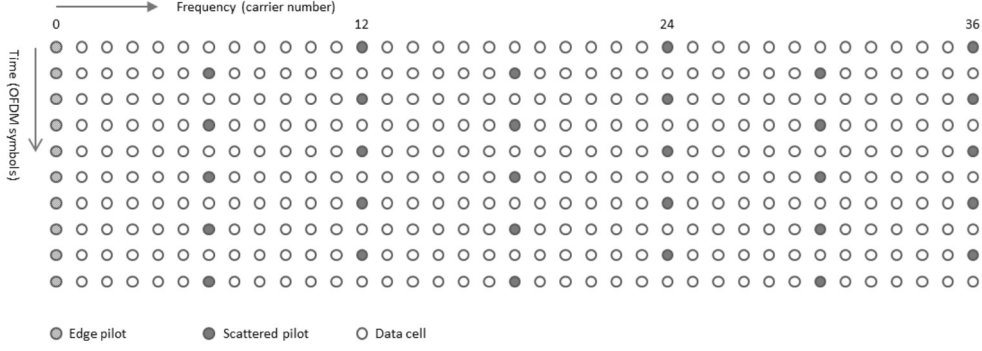


Figure 1 PP2 for DVB-T2 Signal symbol grid

respectively. i.e., $x_{ep}(k) \sim \mathcal{CN}(0, \sigma_{ep}^2)$, $x_p(k) \sim \mathcal{CN}(0, \sigma_p^2)$ and $x_d(k) \sim \mathcal{CN}(0, \sigma_d^2)$. According to (4), $\sigma_d^2 = \alpha \sigma_p^2$, where α is the power ratio of all data subcarriers to all pilot subcarriers in $s(k)$. With (2) and (4) the received signal can be represented as

$$r(k) = \eta e^{j2\pi\Delta_f k} x_{ep}(k) + \eta e^{j2\pi\Delta_f k} x_p(k) + \eta e^{j2\pi\Delta_f k} x_d(k) + n(k) \quad (5)$$

where $x_{ep}(k)$, $x_p(k)$, $x_d(k)$ and $n(k)$ are independently Gaussian distributed. The aim of this paper is to precisely estimate the power of the noise $n(k)$ using $r(k)$ without knowledge of η and Δ_f , therefore achieving reliable detection using ED.

C. Neyman-Pearson Time-domain symbol cross-correlation Detection

The test statistic of ED is given by [4]

$$T = \frac{1}{K \sigma_w^2} \sum_{k=0}^{K-1} |r(k)|^2 \quad (6)$$

As the precise noise power σ_w^2 is unknown, it is often assumed to be in a bounded range of $\sigma_{wl}^2, \sigma_{wu}^2$. The ED in a worst case scenario employs σ_{wu}^2 to replace the exact noise power. This bounded worst behavior (BWB) model causes the noise uncertainty [7], which is defined as $\beta = 10 \log_{10}(\sigma_{wu}^2 / \sigma_w^2)$ in dB.

In [13], NP A Neyman-Pearson Time-domain symbol cross-correlation (NP-TDSC) detector was proposed for sensing DVB-T signal, and the test-statistic is given by

$$T_{NP-TDSC} = \frac{1}{(K-N)\sigma_w^2} \left| \sum_{k=0}^{k=N-1} r(k)r^*(k+N) \right| \quad (7)$$

where the precise noise power σ_w^2 is also required. As in many

other correlation-based detectors, the unknown σ_w^2 can be replaced with the estimated power of the received signal [4]

$$\hat{\sigma}_r^2 = \frac{1}{K} \sum_{k=0}^{K-1} |r(k)|^2 \quad (8)$$

i.e., the noise power and the noise-plus-primary-signal power are not distinguished. NP-TDSC is robust against noise uncertainty, but even with exact noise power, it suffers from considerable performance loss compared to the perfect ED as shown in [14].

III. ED NOISE POWER ESTIMATION FOR DVB-T2

A. Noise Power Estimator

Let the correlation coefficient of $r(k)$ at lag N be $\rho_{\eta\Delta}$, i.e.,

$$\rho_{\eta\Delta} = \frac{1}{\sigma_r^2} E[r(k)r^*(k+N)] \quad (9)$$

where $E[\cdot]$ denotes the expectation operator and

$$\sigma_r^2 = \eta\sigma_{ep}^2 + \eta\sigma_p^2 + \eta\sigma_d^2 + \eta\sigma_w^2 \quad (10)$$

It can be verified that

$$E[r(k)r^*(k+N)] = \eta e^{-j2\pi\Delta_f N} \quad (11)$$

Hence, we have

$$\rho_{\eta\Delta} = e^{-j2\pi\Delta_f N} \frac{\eta\sigma_{ep}^2 + \eta\sigma_p^2}{\eta\sigma_{ep}^2 + \eta\sigma_p^2 + \eta\sigma_d^2 + \eta\sigma_w^2} \quad (12)$$

Let

$$\rho_\eta = e^{j2\pi\Delta_f N} \rho_{\eta\Delta} \quad (13)$$

According to (12), can also be represented as

$$\rho_\eta = \frac{\eta\sigma_{ep}^2 + \eta\sigma_p^2}{\eta\sigma_{ep}^2 + \eta\sigma_p^2 + \eta\sigma_d^2 + \eta\sigma_w^2} \quad (14)$$

which together with $\sigma_d^2 = \alpha\sigma_p^2 + \alpha\sigma_{ep}^2$ gives the noise power

$$\sigma_w^2 = [1 - (1 + \alpha)\rho_\eta] \sigma_r^2 \quad (15)$$

The power of the received signal σ_r^2 can be estimated with (8), with the received signal $r(k)$, $\rho_{\eta\Delta}$ can be estimated as

$$\rho_{\eta\Delta} = \frac{1}{(K-N)\hat{\sigma}_r^2} \left| \sum_{k=0}^{k-N-1} r(k)r^*(k+N) \right| \quad (16)$$

In order to deal with the phase induced by the unknown CFO, $\rho_{\eta\Delta}$ in (16) is separated into two parts as shown in (17).

$$\hat{\rho}_{\eta\Delta} = \frac{1}{(K-N)\hat{\sigma}_r^2} \underbrace{\sum_{k=0}^{k-N-1} r(2k+1)r^*(2k+1+N)}_{\alpha_1} + \frac{1}{(K-N)\hat{\sigma}_r^2} \underbrace{\sum_{k=0}^{k-N-1} r(2k)r^*(2k+N)}_{\alpha_2} \quad (17)$$

According to (16) and (13), α_1 and α_2 in (17) can be represented as

$$\alpha_1 = |\alpha_1| e^{-j(2\pi\Delta_f N + \delta\theta_1)} \quad (18)$$

$$\alpha_2 = |\alpha_2| e^{-j(2\pi\Delta_f N + \delta\theta_2)} \quad (19)$$

where $-2\pi\Delta_f N$ is the common phase induced by the unknown CFO, and represent the phase errors. Noting that N is an even number, α_1 and α_2 have no common elements in their summations. Hence, $\delta\theta_1$ and $\delta\theta_2$ are independent. The phase $-2\pi\Delta_f N$ in α_1 can be cancelled by using the phase of α_2 and vice versa. Hence, $\hat{\rho}_\eta$ can be estimated as

$$\hat{\rho}_\eta = \Re(\alpha_1 e^{-j\angle\alpha_2} + \alpha_2 e^{-j\angle\alpha_1}) \quad (20)$$

where $\angle\alpha_1(\angle\alpha_2)$ denotes the phase of $\alpha_1(\alpha_2)$, and $\Re(\cdot)$ denotes the operation of taking real part which is used because ρ_η is a real number. It is worth mentioning that the estimation of $\rho_{\eta\Delta}$ is independent of the starting time samples primary signal, so time synchronization is not required. Finally, the noise power can be estimated as

$$\hat{\sigma}_w^2 = [1 - (1 + \alpha)\hat{\rho}_\eta] \hat{\sigma}_r^2 \quad (21)$$

It is attractive that the proposed noise power estimator works no matter whether the primary signals are present or absent. This can be taken advantage of with the use of a sliding window approach to achieve a precise estimate of the noise power. the estimate for the sensing duration can be represented as

$$\hat{\sigma}_w^2(q) = \frac{1}{Q} \sum_{q'=q-Q+1}^q \hat{\sigma}_{w,q'}^2 \quad (22)$$

where $Q-1$ previous sensing durations are used, and denotes the estimate noise power using (21) for the q 'th

IV. COMPUTATIONAL COMPLEXITY

The calculation of $\hat{\sigma}_w^2$ with (8) requires approximately K multiplication, which can be shared by the calculation of α_1 and α_2 . Thus, the calculation of ρ_η with (20) requires about

$K-N$ multiplication. The complexity of ED with noise power estimator is only about $2K-N$ multiplication.

V. SIMULATION RESULTS

The effects of different parameters on the proposed algorithm were examined, such as the SUs with independent channels, channel availability and different values of the SNR. The PU signal is a DVB-T2 signal [12] with a 2K mode and $\frac{1}{4}$ guard interval. The DVB-T2 signal has 2048 subcarriers, including 2% scattered pilot subcarriers of total, 2.4% continual pilot subcarriers of total and 343 null subcarriers. The sample number of each OFDM symbol, i.e., $K = 2N$. $Q = 100$ sensing duration are employed for the noise power estimation. The CFO is -0.8π . The SNR is defined as $(\sigma_{ep}^2 + \sigma_p^2 + \sigma_d^2) / \sigma_w^2$. The bandwidth of the PU signal is 8 MHz and modulation type is QPSK. The average occupancy rate for the PU is set to 50%, i.e. the probability of presence and absence of the PU signal is fixed to an equal probability (0.5), respectively. The simulation is based on the Monte Carlo method in MATLAB with 100,000 iterations.

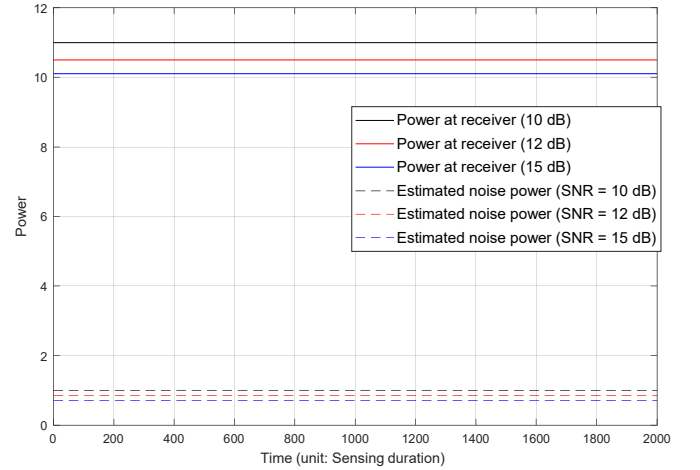


Figure 2. Power of PU plus noise at the SU receiver and estimated noise power when $Q = 100$ for SNR = 10 dB, 12 dB, 15 dB.

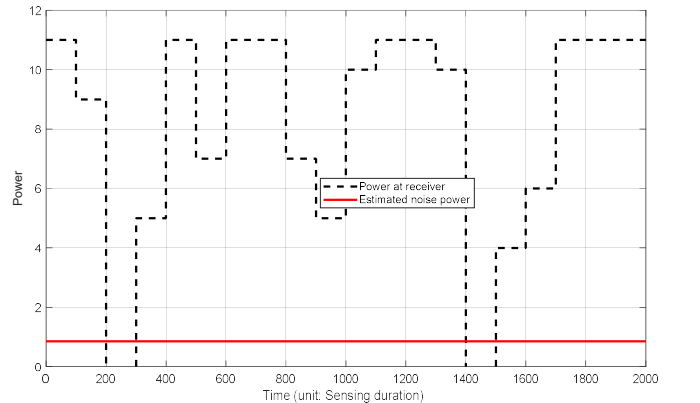


Figure 3. Power of PU plus noise at the SU receiver and estimated noise power when $Q = 100$ for varying SNR.

In figure 2 and figure 3, the estimated noise power when and $Q = 100$ for SNR = 10 dB, 12 dB, 15 dB and for a varying SNR is presented, respectively. The PU signal is absent from the 300th to the 400th sensing duration. The results illustrate that the noise power estimator gives an increased performance when the PU is present or absent.

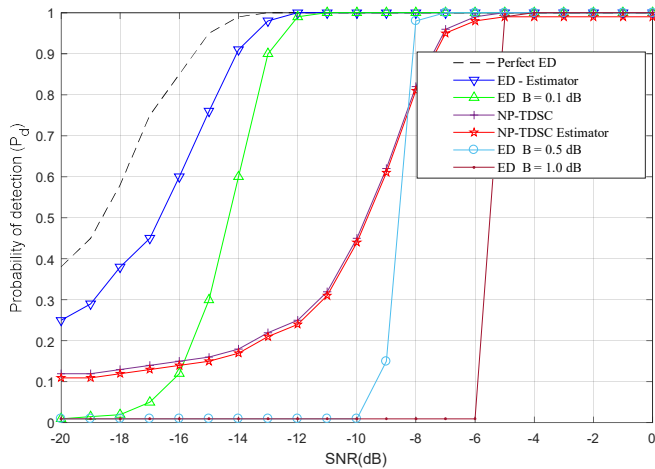


Figure 4. Probability of detection (P_d) of ED and NP-TDSC vs SNR (dB) when $Q = 100$. Note β is denoted B .

In figure 4, the P_d of ED and NP-TDSC for a target of probability of false alarm P_f of 0.1 is presented. It can be seen that the perfect ED exhibits the best performance, but from figure 5 it can be seen that the replacement of noise power using a BWB mode brings low false alarm probability but causes sever degradation of detection probability. The performance of ED with noise uncertainty $\beta = 1.0$ dB is reduced to that of NP-TDSC. To obtain a P_d of 0.9, the perfect ED requires an SNR of about ~ 16 dB, while the NP-TDSC requires an SNR of about -7.2 dB, and ED with $\beta = 1.0$ dB requires a higher SNR.

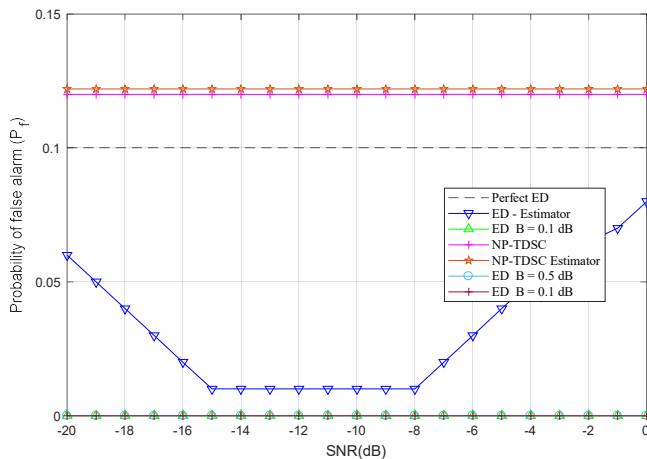


Figure 5. Probability of false alarm (P_f) of ED and NP-TDSC vs SNR (dB) when $Q = 100$. Note β is denoted B .

By applying the proposed noise power estimate to ED and the NP-TDSC, leads to “ED – estimator” and “NP-TDSC-estimator”, respectively, whose P_d and P_f are presented in figure 4 and 5. The performance of NP-TDSC- estimator is similar to

that of NP-TDSC for the same noise power. For ED – estimator, an SNR of -13 dB is required to achieve the detection probability of 0.9. It is shown that “ED–estimator” significantly outperforms NP-TDSC with the same noise power and the ED with low noise uncertainty of $\beta = 0.1$ dB.

VI. CONCLUSION

In this paper, ED for DVB-T2 spectrum sensing acquiring a precise estimate of the noise power in multipath channels has been proposed. Results have shown that ED with noise power estimator considerably outperform traditional correlation-based detectors and the ED with BWB noise model.

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