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# Spectrum Sensing of DVB-T2 Signals using a Low Computational Noise Power Estimation

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**Abstract**— Cognitive radio is a promising technology that answers the spectrum scarcity problem arising from the proliferation of wireless networks and mobile services. In this paper, spectrum sensing of digital video broadcasting-second generation terrestrial (DVB-T2) signals in AWGN, WRAN and COST207 multi-path fading environment are considered. ED is known to achieve an increased performance among low computational complexity detectors, but it is susceptible to noise uncertainty. Taking into consideration the edge pilot and scattered pilot periodicity in DVB-T2 signals, a low computational noise power estimator is proposed. Analytical forms for the detector are derived. Simulation results show 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 multi-path fading environments. Moreover, based on this algorithm a practical sensing scheme for cognitive radio networks is proposed.

**Keywords** — energy detection; spectrum sensing; cognitive radio; DVB-T2; DTT; noise estimation.

## I. INTRODUCTION

Cognitive radio (CR) is a promising technology that answers the spectrum scarcity problem arising from the proliferation of wireless networks and mobile services. CR has been designed for autonomous reconfiguration by learning from and ultimately adapting to the continuously changing radio environment [1]. CR capabilities, along with some adjustments in its operation will be a key technology for fifth generation (5G) heterogeneous network deployment. Currently the number of different wireless devices and technologies available, the exponential 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. This 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. Accordingly, 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 whereby unlicensed secondary users (SUs) are allowed to opportunistically share the spectrum of licensed primary users (PUs) without causing any adverse impact on the PU transmission [2].

In CR, spectrum sensing is a key enabling technology. The main techniques of sensing are commonly classified into matched filter detection, energy detection (ED), feature

detection and waveform based detection [3]. Of these algorithms, ED, is commonly used as a spectrum sensing technique due to its low computational complexity and hardware simplicity [4-6]. Significantly, prior information regarding the primary signal is not required. 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 [7]. The standard is intended for cognitive operation in the digital TV bands, among others. As a result of 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) [7]. CR operations in the DVB-T spectrum have also been deliberated in other research literature [8-10]. In Cognitive radio network's (CRN), DVB-T signals are one of the most important signal type used by PUs in TV bands [11].

In [12-15], intended at sensing signals such as DVB-T2 signals, correlation-based detectors were proposed by exploiting correlation cyclic prefix or pilot. The exact noise power in these detectors are replaced with the power of received signal which is impracticable because the received signal is highly likely to contain the PUs signals. The detectors are robust to noise uncertainty, but in comparison with the prefect ED, they exhibit considerable performance degradation in the region of 8 ~ 14dB [16]. In this paper, by taking into consideration the periodicity in DVB-T2 signals, a low computation 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 in different channel is proposed, the signification of carrier frequency offsets (CFO's) is eliminated in the noise power estimate and frequency synchronization is not required. Simulation results show that the proposed noise power estimator enables ED in different channel conditions is able to obtain a reliable detection performance. The rest of this paper is organized as follows: In Section II, the characteristics of DVB-T2 signals are presented. Section III presents the system model. Next, Neyman-Pearson time-domain symbol cross-correlation detection is introduced in section IV. In Section V, the proposed low computational ED noise power estimation for DVB-T2 signals is presented. Section VI, presents simulation results concerning the performance of the proposed algorithms in various multi-path fading environments. Finally, conclusions are given in section VII.

## II. DVB-T2 Signal

DVB-T2 supports SD, HD, UHD, mobile TV, radio, or any combination thereof. DVB-T2 signals are more resilient against certain types of interference than DVB-T [11]. Since its publication in 1997, over 70 countries have deployed DVB-T services and 69 countries have now adopted or deployed DVB-T2 [11]. DVB-T2 is the world's most advanced DTT system, offering more robustness, flexibility and 50% more efficiency than any other DTT system [11]. This well-established standard benefits from massive economies of scale and very low receiver prices. Owing 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 and the ability to reuse existing reception antennas [14]. From a spectrum sensing point of view, important DVB-T2 parameters 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). 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.

## III. System Model

The primary signal is modeled as a Gaussian process, which is a prevalent assumption in the literature on spectrum sensing. In [4-7] [13], 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,  $\Delta_f$  denotes the unknown CFO,  $n(k)$  denotes the circularly symmetric complex Gaussian white noise with mean zero and variance  $\sigma_n^2$ , 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 [11]. 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

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

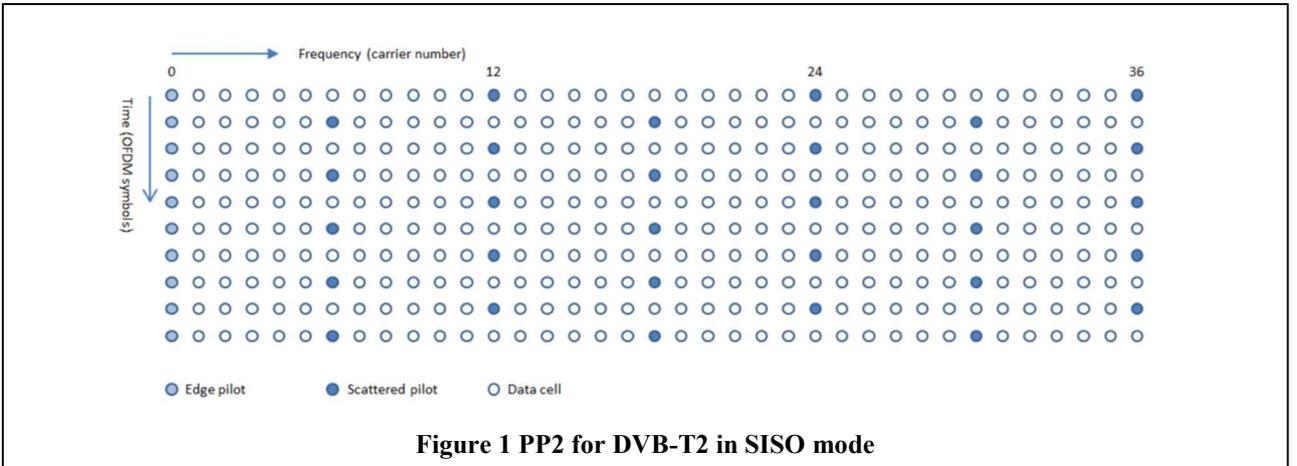


Figure 1 PP2 for DVB-T2 in SISO mode

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  $\sigma_{ep}^2$ ,  $\sigma_p^2$  and  $\sigma_d^2$  denote the power of  $x_{ep}(k)$ ,  $x_p(k)$  and  $x_d(k)$ , 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  $\alpha$  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  $\eta$  and  $\Delta_f$ , therefore achieving reliable detection using ED.

#### IV. NEYMAN-PEARSON TIME-DOMAIN SYMBOL CROSS-CORRELATION DETECTION

The test statistic of ED is given by [6]

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

As the precise noise power  $\sigma_w^2$  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  $\sigma_{wu}^2$  to replace the exact noise power. This bounded worst behavior (BWB) model causes the noise uncertainty [6], 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  $\sigma_w^2$  is also required. As in many other correlation-based detectors, the unknown  $\sigma_w^2$  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].

#### V. LOW COMPUTATIONAL ED NOISE POWER ESTIMATION

##### 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  $\sigma_r^2$  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).

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

According to (16) and (13),  $\alpha_1$  and  $\alpha_2$  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,  $\alpha_1$  and  $\alpha_2$  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  $\alpha_1$  can be cancelled by using the phase of  $\alpha_2$  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  $\rho_\eta$  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. The calculation of  $\hat{\sigma}_w^2$  with (8) requires approximately  $K$  multiplication, which can be shared by the calculation of  $\alpha_1$  and  $\alpha_2$ . Thus, the calculation of  $\rho_\eta$  with (20) requires about  $K-N$  multiplication. The complexity of ED with noise power estimator is only about  $2K-N$  multiplication.

## VI. SIMULATION RESULTS

### Multi-path fading environment:

Recalling the definition of SNR, the form does not depend on the type of fading. Therefore, the low computational ED noise power estimation could apply for multi-path fading. In the simulation, the detection performance is verified in AWGN channel, the multi-path channels of the WRAN with the profile B and the COST207 channels.

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 [11] with a 2K mode and  $\frac{1}{4}$  guard interval as shown in table 1. 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  $-0.8\pi$ . 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.

Table 1. DVB-T2 Parameters

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

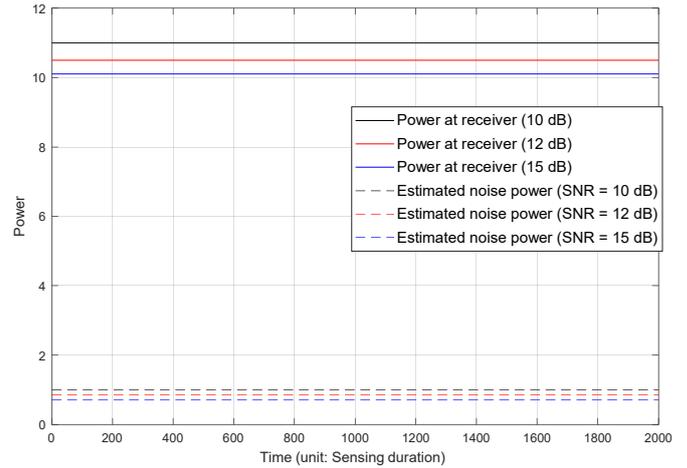


Figure 2. Power of PU plus noise at the SU receiver and estimated noise power when  $Q = 100$  for SNR = 10 dB, 12 dB, 15 dB.

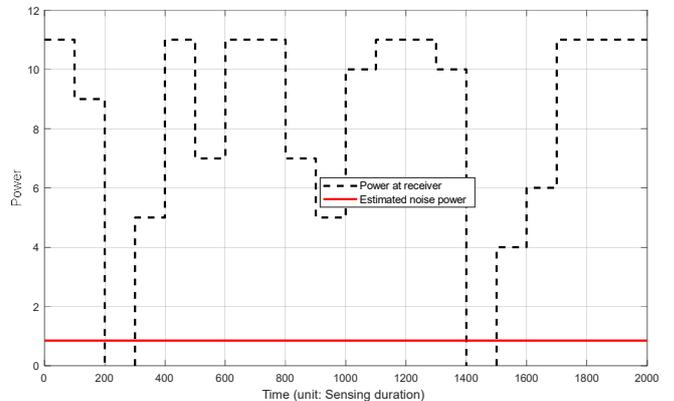


Figure 3. Power of PU plus noise at the SU receiver and estimated noise power when  $Q = 100$  for varying SNR.

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 300<sup>th</sup> to the 400<sup>th</sup> sensing duration. The results illustrate that the noise power estimator gives an increased performance when the PU is present or absent.

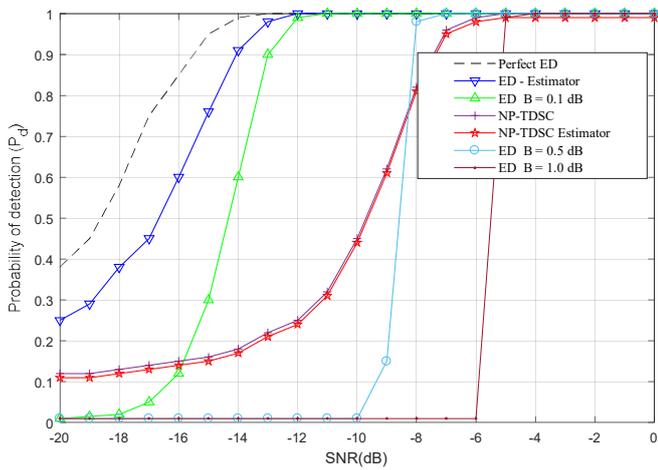


Figure 4. Probability of detection ( $P_d$ ) of ED and NP-TDSC vs SNR (dB) when  $Q = 100$  for AWGN. Note  $\beta$  is denoted  $B$ .

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  $\sim 16$  dB, while the NP-TDSC requires an SNR of about  $-7.2$  dB, and ED with  $\beta = 1.0$  dB requires a higher SNR.

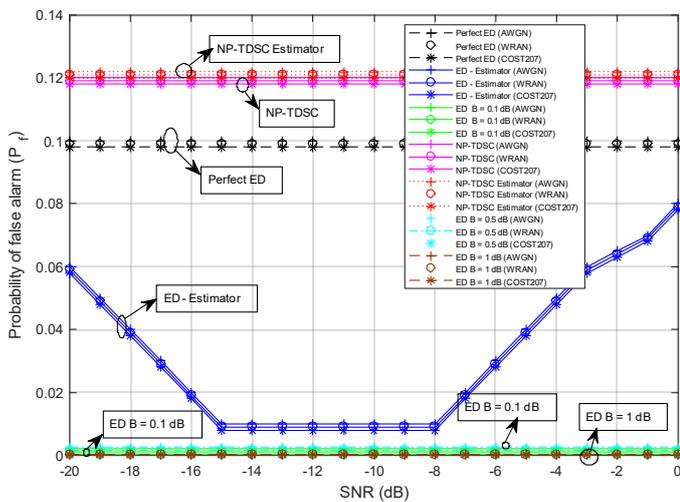


Figure 5. Probability of false alarm ( $P_f$ ) of ED and NP-TDSC vs SNR (dB) when  $Q = 100$  for AWGN, WRAN and COST207 environments.. Note  $\beta$  is denoted  $B$ .

By applying the proposed noise power estimate to ED and the NP-TDSC, leads to “ED – estimator” and “NP-TDSC-estimator”, respectively, who’s  $P_d$  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 . Figure 5 and figure 6 shows that the estimator can be applied to various multi-path environments such as AWGN, WRAN and COST207 and it yields similar results.

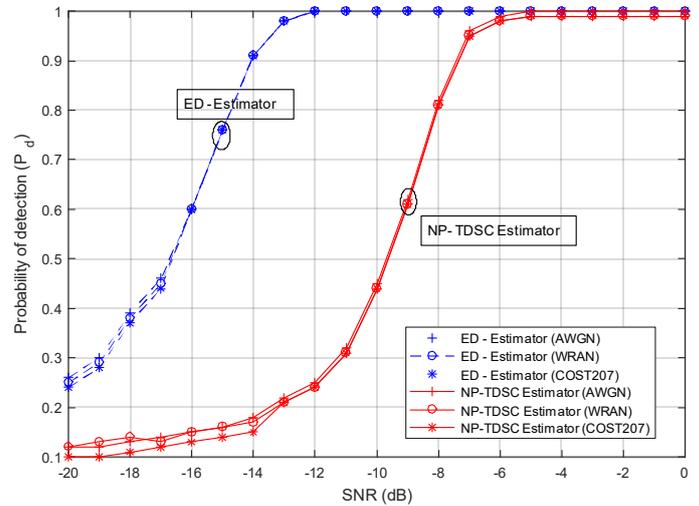


Figure 6. Probability of detection ( $P_d$ ) of ED - Estimator and NP-TDSC Estimator vs SNR (dB) when  $Q = 100$  for AWGN, WRAN and COST207 environments.

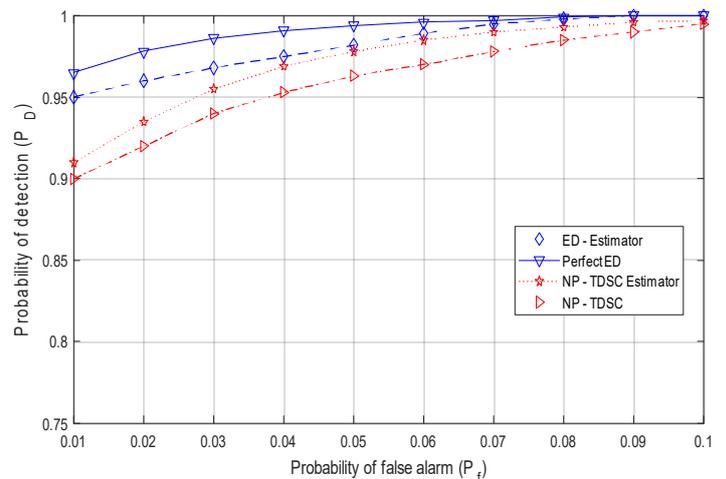


Figure 7. Probability of detection ( $P_d$ ) vs Probability of false alarm ( $P_f$ ) (ROC Curve) of ED and NP-TDSC.

In figure 7, the ROC of ED, NP-TDSC, ED – estimator and NP-TDSC- estimator are presented. Considering the recommendation of missed detection from the IEEE 802.22 standard and the false alarm for the efficiency of DVB-T practical CR systems, this paper focuses on the ROC with  $P_f$  in the range of 0.01 to 0.1. At  $-10$ dB, the performance of ED – estimator is 97.8% and 99.8% with false alarm in of 1% and 10%, respectively. The performance of NP-TDSC- estimator are 90.2% and 98.2% with false alarm in of 1% and 10%, respectively.

## VII. CONCLUSION

In this paper, ED for DVB-T2 spectrum sensing acquiring a precise estimate of the noise power in AWGN, WRAN and COST207 multi-path fading environment 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. Finally, and most important of all, in this paper a practical sensing scheme for cognitive radio networks has been proposed.

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