Spectrum Sensing for TV White Space in North America

Hou-Shin Chen, *Member, IEEE,* and Wen Gao, *Member, IEEE*

***Abstract*—Spectrum sensing for the use of TV spectrum in North America is addressed in this paper. A unified signature-based spectrum sensing algorithm is designed for the primary licensed signals including NTSC analog TV and ATSC digital TV signals. In addition, an autocorrelation-based spectrum sensing algorithm is proposed for the secondary licensed signals such as wireless microphone signals. To our knowledge, the spectrum sensing algorithms presented in this paper are the first one which can reliably determine the availability of a TV channel for unlicensed radio transmissions in the presence of a strong adjacent channel interferer and the target signal strength is as low as -114 dBm.**

***Index Terms*—Spectrum Sensing, Cognitive Radio, TV White Space, ATSC, NTSC, Wireless Microphone, FM**

I. Introduction

IN NOVEMBER 2008, the Federal Communications Commission (FCC) approved the use of unlicensed radio transmitters in the broadcast television spectrum at locations where that spectrum is not being used by licensed services (this unused TV spectrum is often termed "white space"), under certain rules [1]. This approval brings the ever increasing interest in cognitive radio technologies to a new level in academia and industry.

Cognitive Radio was introduced in [2] to implement negotiated or opportunistic spectrum sharing to provide a viable solution to the problem of demand for more wireless spectrum. In order to use TV white space, FCC regulations require that white space devices will be able to sense, at levels as low as -114 dBm, licensed incumbent signals such as TV signals (digital and analog), wireless microphone signals, and signals of other services that operate in the TV bands on intermittent basis [1]. The noise power in a 6 MHz TV channel under normal temperature is about -96 dBm assuming that the noise figure of a sensing device is 10 dB. Thus, the sensing requirement set by FCC is about -18 dB in terms of signal-to-noise power ratio (SNR), which is a rather difficult task.

Power detection, or energy detection, was widely used to determine the presence of signals without prior knowledge of signals. However, power detectors do not function well when SNR is low. Under low SNR conditions, accurate noise power levels and large numbers of data samples are needed to achieve a good sensing performance [6]. However, an accurate noise power level is difficult to obtain because it can be affected by several factors, e.g., temperature and system calibration. The lack of knowledge about the noise power is termed noise uncertainty. The amount of noise uncertainty can be as large as ± 1 dB. When the noise uncertainty is as large as 1 dB, a power detector fails when the SNR is below -3.3 dB even with a very long sensing time [7]. Another blind detection method is eigenvalue-based spectrum sensing algorithm which makes use of the property that the eigenvalues of the sample covariance matrix of an AWGN noise are approximately the same when the number of collected samples are large enough [12]. Although some eigenvalue-based detectors can overcome the problem of noise uncertainty, both of the power detector and eigenvalue-based detector have a disadvantage that they cannot distinguish interference signals from licensed signals. Consequently, both methods generate a high false alarm rate due to an interference signal.

TV signals (analog and digital) and wireless microphone signals are primary and secondary licensed signals in the TV broadcast bands which demand protection when CR communication systems are deployed. The process of converting from analog TV to digital TV transmissions in United States was finished in June 2009. After the transition, full service TV stations broadcast only ATSC Digital TV (DTV) signals [5]. A variety of sensing algorithms for ATSC DTV signals have either been reported in the literature, or proposed to IEEE 802.22 Group [8]–[13]. Some of them claim to achieve the sensing requirement set by FCC [10]–[13]. In addition, several white space prototypes developed by different companies have been tested by FCC. **The FCC test report [3] reveals that all these white space prototypes (spectrum sensors) have a high false alarm rate in Two-Signal (Adjacent Channel Interference such as an adjacent TV signal) test model especially when a strong DTV signal exists in the lower adjacent channel**. It is stated in the FCC test report [3] that "*although different algorithms are utilized among the prototype devices to detect the presence of a DTV signal, they all appear to share a similarity in that they all sample the spectrum in proximity of an anticipated DTV pilot signal.*" The DTV pilot tone is only 310 kHz away from the edge of the lower adjacent channel as shown in Fig. 1. When there is a strong DTV signal in the lower channel, this pilot tone signal is severely impaired by the adjacent channel interference, as shown in Fig. 2. As a result, the white space devices, which try to utilize the pilot tone signal, fail to indicate that an empty channel is free to use when a strong TV signal exists in the adjacent channel. Spectrum sensing of ATSC DTV signals draws more attention than NTSC analog TV signals [4] and wireless microphone signals. However, although full service TV stations have been converted to ATSC DTV stations,

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H. Chen and W. Gao are with Technicolor Research & Innovation, Princeton, Princeton, NJ 08540 USA (e-mail: houshin.chen@gmail.com, wgao@ieee.org).
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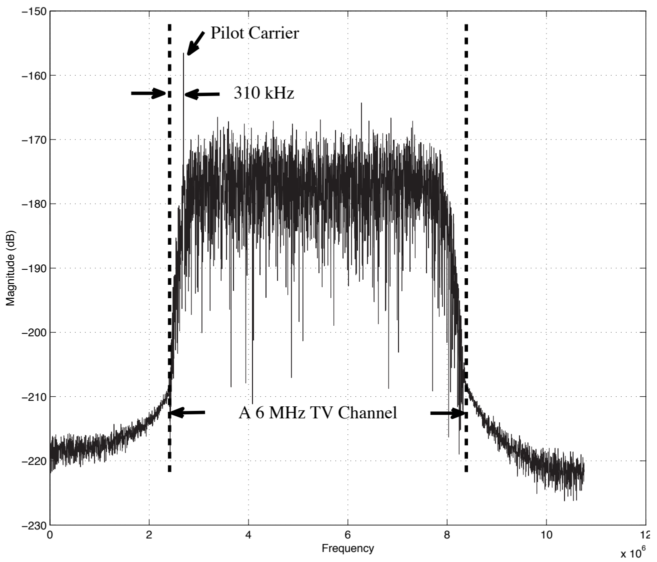

Fig. 1. Frequency response of a pure ATSC DTV signal

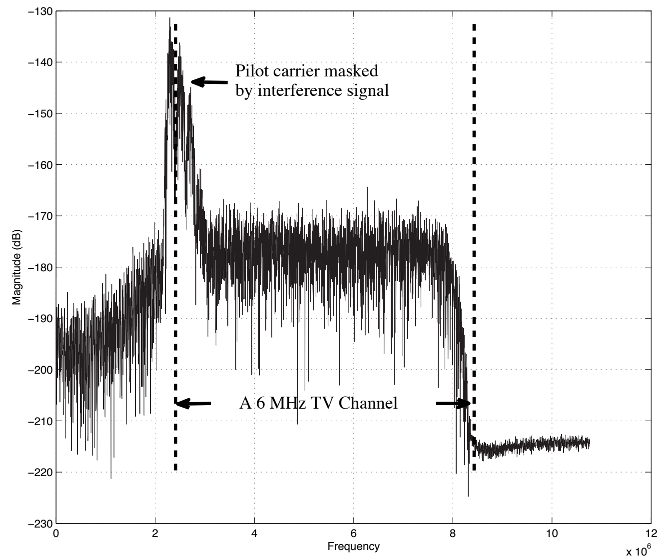

Fig. 2. Frequency response of a pure ATSC DTV signal with a lower adjacent channel interference.

the digital transition is not required for low power NTSC stations or translators. In addition, wireless microphones are commonly used both within the studio and during sports or news gathering events. There are approximately 35k to 70k licensed wireless microphone devices operating in the United States. Thus, spectrum sensing of the NTSC analog TV and wireless microphone signals is important as well. **The FCC report [3] also reveals that the white space device fails to distinguish the presence of wireless microphone or NTSC signals from interference signals in the Two-Signal test model**.

Motivated by the need to reliably detect incumbent licensed signals operating in the TV bands in North America, spectrum sensing algorithms for ATSC DTV signals, NTSC analog TV signals and FM wireless microphone signals are proposed in this paper. These algorithms are designed to reliably sense the target signals under two tough conditions including low SNR and strong adjacent channel interference. To our knowledge, the spectrum sensing algorithms presented in this paper are the first one which can reliably determine the availability of a TV channel for unlicensed radio transmissions in the presence of a strong adjacent channel interferer and the SNR is as low as -18 dB. This paper is organized as follows: a unified spectrum sensing algorithm for NTSC and ATSC TV signals is described in Section II. Spectrum sensing for FM wireless microphone signals is considered in Section III. In Section IV, the satisfactory performance of the sensing algorithms under strong adjacent channel interference is demonstrated using computer simulations. Finally, conclusions are given in Section V.

II. Spectrum Sensing for TV Broadcast Signals

In this Section, a unified spectrum sensing algorithm is described for both NTSC analog and ATSC digital TV signals.

A. ATSC DTV Signals

First, we briefly describe the structure of ATSC DTV signals [5]. ATSC DTV signals consist of consecutive data segments as shown in Fig. 3. A complete data segment has 832 symbols: 4 symbols for data segment SYNC, and 828 data symbols. The two-level data segment SYNC employs a 1001 pattern ([5 -5 -5 5], after mapping to 8-PAM symbols) and the data symbols are eight-level PAM (8-PAM) symbols. The Vestigial Sideband (VSB) modulation is applied to improve bandwidth efficiency. An 8-PAM with VSB modulation is also called an 8-VSB modulation. Note that before VSB modulation, a constant of 1.25 is added to each symbol for the purpose of creating a pilot tone as shown in Fig. 1. As mentioned before, this pilot tone is widely used to perform spectrum sensing in variety of algorithms. However, algorithms utilizing this pilot carrier severely suffer from adjacent channel interference. Thus, instead of utilizing the pilot tone, we use data segment SYNC to perform spectrum sensing. There are two reasons to utilize data segment SYNC for spectrum sensing. First, as shown in Fig. 4, the signal power of the data segment SYNC spreads over the whole 6 MHz TV channel so that unlike the pilot tone, the data segment SYNC signal will not be fully shaded by the adjacent channel interference. Second, because the time difference between any two data segment SYNC for a given sensing time is at most tens of milli-seconds, it is reasonable to assume that they encounter the same channel effects including timing offset, frequency offset, and multi-path fading effect. Thus, the correlation of two data segment SYNC generates a constant term. The interference and noise signals do not exhibit this property. As a result, the correlation of two data segment SYNC elements is used as a basic approach to perform spectrum sensing for ATSC DTV signals. Let $y[n]$ be the received complex baseband signal, and

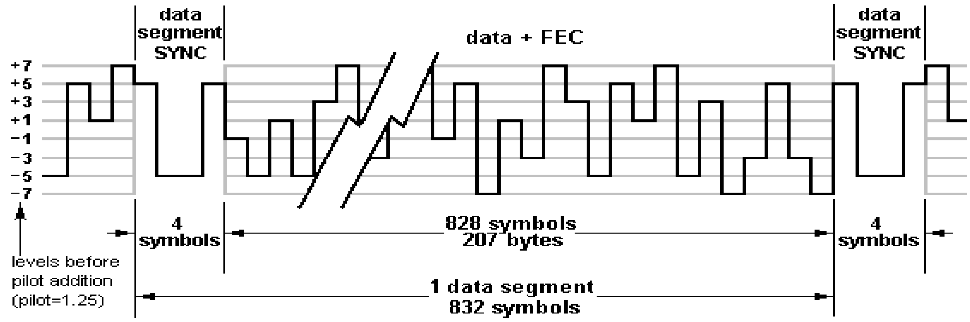

Fig. 3. ATSC DTV data signal segment

define the accumulated SYNC correlation function to be

$$C(m,s)=\frac{1}{KN_s}\sum_{n=0}^{N_s-1}\sum_{k=0}^{K-1}y[m+k+nL]y^*[m+k+(n+s)L] \quad (1)$$

where m is the starting sample index and s is correlation delay in terms of data segments. Here, the parameter $K=4$ is the number of symbols in a data segment SYNC and $L=832$ is the number of symbols in a data segment. In addition, the parameter N_s is the total number of two data segment SYNC element pairs which are separated by s data segments for a given sensing time. For example, if a sensing time of 10 data segments is used, the parameters of N_s will be $N_1=9$, $N_2=8$, etc. Note that although symbol timing information is lacking, the absolute value of $C(m,s)$ is maximized when the starting sample index is the first symbol of a data segment SYNC. Consequently, the decision statistic for the SYNC correlation-based spectrum sensing algorithm is given by

$$T_C(s)=\max_{0\le m\le L-1}|C(m,s)| \quad (2)$$

Note that s must be a non-zero positive integer and $T_C(1)$ is used to perform spectrum sensing in [9]. An important observation is that the sensing performance of using only $T_C(1)$ in [9] is -108 dBm for a sensing time of 92.5 ms, which is not good enough to meet the sensing requirement of -114 dBm set by FCC [1]. In order to improve sensing performance, the accumulated SYNC correlation functions $C(m,s)$ are computed for different value of s for a given sensing time and are combined together. Since the noise embedded in each of the accumulated SYNC correlation functions, $C(m,s)$, for different values of s is not highly correlated, it is expected that a performance gain can be achieved by combining various accumulated SYNC functions $C(m,s)$. However, $C(m,s)$ corresponding to different s cannot be linearly (coherently) combined due to different phase terms caused by the carrier frequency offset. Assume that δ_f is the frequency offset normalized with respect to the sampling frequency which is symbol rate here. The phase term embedded in $y[n]$ is equal to $e^{j2\pi\delta_f n+\phi}$, where ϕ is a random initial phase of the signal. As a result, $C(m,s)$ contains a phase term $e^{-j2\pi\delta_f Ls}$, which is a function of s. In order to solve this problem, let's further define a function,

$$Q(m,s,s+d)=C(m,s)C^*(m,s+d) \quad (3)$$

which is the conjugate product of two accumulated SYNC correlation functions. Then the phase term embedded in

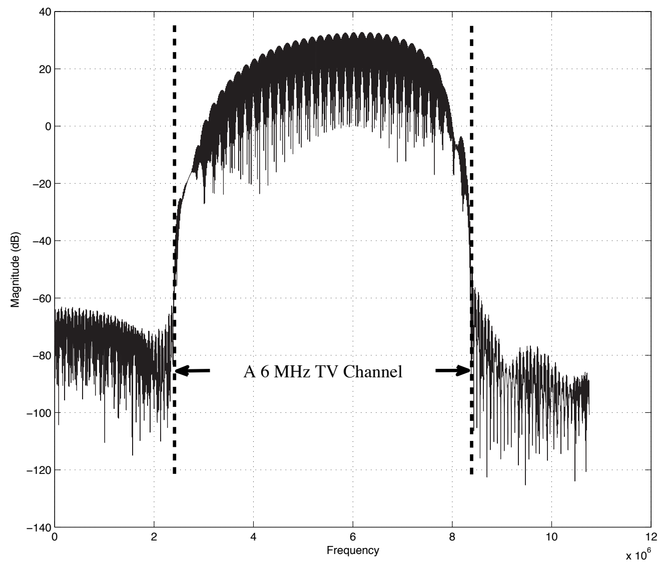

Fig. 4. Frequency response of ATSC data segment SYNC signal.

$Q(m,s,s+d)$ becomes $e^{j2\pi\delta_f Ld}$ which is a function of d, and hence, we can linearly combine $Q(m,s,s+d)$ for different s. Based on the observation above, the decision statistic of the SYNC-correlation-based spectrum sensor can be given by

$$T_Q=\max_{0\le m\le L-1}|\sum_{s=1}^{N_Q}a_sQ(m,s,s+1)| \quad (4)$$

where a_s are combining coefficients and N_Q is the number of $Q(\cdot)$ summed. Note that in (4), d is chosen to be 1 to efficiently utilize all accumulated SYNC correlation functions obtained in a period of sensing time. Following the same procedure in [14], the combining coefficients can be found to be $a_s=N_sN_{s+1}$.

B. NTSC Analog TV Signal

The North American NTSC TV signal is an analog signal consisting of picture frames at a rate of $30/1.001\approx 29.97$ frames/s and a total of 525 scan lines per frame [4]. The NTSC TV signal uses interlaced scan so that each frame is scanned in two fields and each field contains half of the scan lines in a frame. As shown in Fig. 5, the first 19 lines of each field

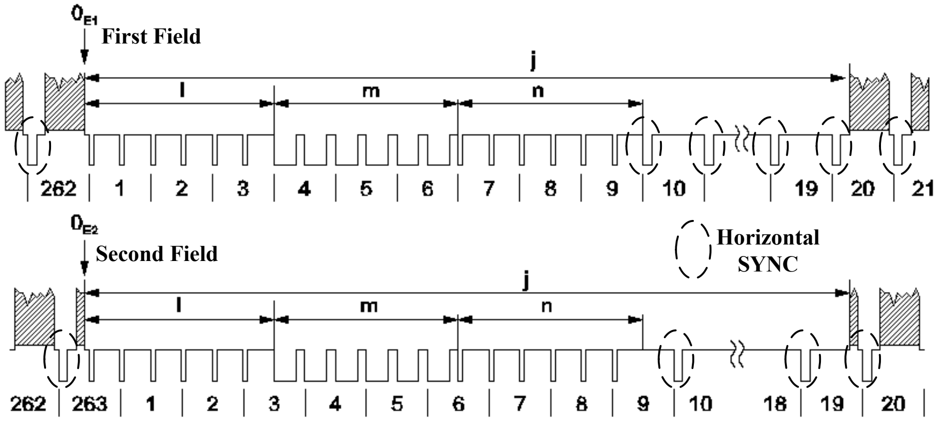

Fig. 5. NTSC Field SYNC signal.

are vertical blanking interval (field SYNC) and they are not displayed. The remaining 487 scan lines in a frame have a Horizontal SYNC signal in the beginning of a line. In fact, only the first 9 lines of vertical blanking interval do not contain a Horizontal SYNC signal. It is clear that the Horizontal SYNC of NTSC signals is analogous to the data segment SYNC of ATSC signals. Consequently, the spectrum sensing algorithm described in the previous Section can be applied to NTSC TV signals by modifying parameters. For NTSC TV signals, the parameters K is the number of samples within a Horizontal SYNC, L is the number of samples in a line and N_s is the total number of two Horizontal SYNC element pairs which are separated by s lines for a given sensing time. Note that when the accumulated SYNC correlation functions $C(m, s)$ are used to perform spectrum sensing for NTSC TV signals, the sampling rate should be carefully chosen such that each line contains an integer number of samples. Furthermore, since there are 9 consecutive lines which do not have Horizontal SYNC in each field, the sensing time should not be less than 11 lines. One point worth mentioning is that the signal interval of the Horizontal SYNC signal is about 4.7 microseconds and the interval of a line is 63.55 microsecond. The density of the Horizontal SYNC signal is about $4.7/63.55 \approx 7.4\%$ [4]. On the other hand, the density of the data segment SYNC of ATSC signals is $4/832 \approx 0.48\%$. Thus, the Horizontal SYNC contained in NTSC signals is a much stronger feature than the data segment SYNC contained in ATSC signals. It can be expected that the sensing performance of NTSC signals will be much better than that of ATSC signals for the same sensing time.

C. Interference Alleviation Scheme

In the FCC report [3], all white space devices submitted by different companies cannot accurately indicate that an empty channel is free to use in the Two-Signal (Adjacent Channel Interference) test model, especially when a strong DTV signal exists in the lower adjacent channel. Fig. 1 and Fig. 2 show the spectrum of a pure ATSC DTV signal and an ATSC DTV signal with an interference due to leakage from a strong lower adjacent channel DTV signal, respectively. The center frequency of the signals in these two figures is 5.38 MHz. In Fig. 1, the signal power is at a level of -114 dBm. The amplitude of the pilot tone is about 13 dB higher and it can be used to perform spectrum sensing even SNR is as low as -20 dB. This is the reason that the pilot tone signal is widely used in different spectrum sensing algorithms. It is stated in the FCC report [3] that "*although different algorithms are utilized among the prototype devices to detect the presence of a DTV signal, they all appear to share a similarity in that they all sample the spectrum in proximity of an anticipated DTV pilot signal*." In the Two-Signal model [3], the signal power of the adjacent channel can be as high as -28 dBm. In Fig. 2, the in-band DTV signal power is -114 dBm and the lower adjacent DTV signal power is -28 dBm. It is obvious that the pilot tone signal is completely shaded by the interference. It is very difficult for a pilot-tone-based spectrum sensor to discriminate between a DTV pilot tone signal or an interference signal. This explains why the spectrum sensors, which sample the spectrum in proximity of an anticipated DTV pilot signal, have a high false alarm rate in the Two-Signal test model. Fig. 4 plots the ATSC data segment SYNC signal in the frequency domain. We can see that the data segment SYNC signal spreads mainly from 3 MHz to 8 MHz. Thus, instead of a regular 6 MHz receiver filter, a narrower filter having e.g., 5 MHz, which serves as an interference alleviation filter can be applied to suppress the power of the adjacent channel interference. Note that the interference alleviation filter can be implemented in the analog or digital domain. By using the interference alleviation filter, the interference signal is greatly suppressed but the data segment SYNC signal is mostly preserved. Then, the SYNC-correlation-based spectrum sensing algorithm described in previous sections can be implemented. In addition, the proposed interference alleviation scheme is not limited to SYNC-correlation-based spectrum sensing algorithms for ATSC and NTSC TV signals. It can be applied to spectrum sensing algorithms of other types of signals e.g., OFDM signals, provided that the feature used for spectrum sensing is a wide band signal. Note that in practice, we may encounter random spikes (tone signals) in the environment. The sensing performance of the described algorithms will degrade depending on the number of spikes and their power levels.

D. Threshold Setting for a Specific Probability of False Alarm

1) AWGN Only: Consider that if only AWGN is present and the variance of the noise after interference filtering, $\tilde{w}[n]$

is $\tilde{\sigma}^2$, then the accumulated SYNC correlation function is

$$C(m,s) = \frac{1}{KN_s}\sum_{n=0}^{N_s-1}\sum_{k=0}^{K-1}\tilde{w}[m{+}k{+}nL]\tilde{w}^*[m{+}k{+}(n{+}s)L]. \tag{5}$$

According to the Central Limit Theorem [23], when KN_s is sufficiently large, $C(m,s)$ will approach a circularly symmetric complex Gaussian variable with zero-mean and a variance of $\frac{\tilde{\sigma}^4}{KN_s}$. Let

$$\begin{aligned} T_Q(m) &= \sum_{s=1}^{N_Q} N_s N_{s+1} Q(m,s,s+1) \\ &= \sum_{s=1}^{N_Q} N_s N_{s+1} C(m,s) C^*(m,s+1) \end{aligned} \tag{6}$$

and apply the Central Limit Theorem again, when N_Q is sufficiently large, $T_Q(m)$ will approach a circularly symmetric complex Gaussian variable with zero-mean and a variance of $\sigma^2_{T_Q}$. where

$$\sigma^2_{T_Q} = \sum_{s=1}^{N_Q} \frac{N_s N_{s+1}\tilde{\sigma}^8}{K^2}. \tag{7}$$

Thus, the amplitude, $|T_Q(m)|$ is Rayleigh distributed with the following probability density function (pdf)

$$f_{|T_Q(m)|}(t:H_0) = \begin{cases} \frac{2t}{\sigma^2_{T_Q}} e^{-\frac{t^2}{\sigma^2_{T_Q}}}, & t \geq 0 \\ 0, & t < 0 \end{cases} \tag{8}$$

where H_0 is a hypothesis corresponding to the presence of noise only. According to (4), the decision statistic T_Q is the maximum of $\{|T_Q(m)|\}_{m=0}^{L-1}$. The random variables $\{|T_Q(m)|\}_{m=0}^{L-1}$ follow a joint Rayleigh distribution. The joint Rayleigh distribution for more than four random variables with arbitrary covariance matrix is still an open research problem [22]. Thus, we will not try to derive the exact probability distribution of the random variables $\{|T_Q(m)|\}_{m=0}^{L-1}$. Instead, we assume that the random variables are independent in order to calculate an approximate threshold. By making the assumption that the $\{|T_Q(m)|\}_{m=0}^{L-1}$ are independent, from [23], we find that the cumulative probability distribution (cdf) of T_Q is

$$F_{T_Q}(t:H_0) = \left(\int_0^t \frac{2u}{\sigma^2_{T_Q}} e^{-\frac{u^2}{\sigma^2_{T_Q}}} du\right)^L. \tag{9}$$

Then, for a particular probability of false alarm (P_{FA}), the corresponding threshold γ_{T_Q} can be found by

$$P_{FA} = 1 - F_{T_Q}(\gamma_{T_Q}:H_0). \tag{10}$$

Finally, after some straightforward calculation, we have

$$\gamma_{T_Q} = \mu_{T_Q}\left(\sigma^2_{T_Q}\ln\frac{1}{1-(1-P_{FA})^{1/L}}\right)^{1/2} \tag{11}$$

where μ_{T_Q} is an heuristic adjusting factor added artificially to account for the independence assumption mentioned above. According to computer simulations, we find that when a sensing time of 1200 data segments (92 ms) is used and T_Q is computed by combining accumulated SYNC correlation functions $C(m,s)$ for $s = 1,2,...,1000$, i.e., $N_Q = 999$, the heuristic adjusting factor μ is around 0.7.

2) Interference Plus AWGN: Although the adjacent channel interference is greatly suppressed by interference filtering, we cannot ignore it all the time. Thus, the threshold obtained from (11) can not be used. If the interference power after interference filtering is known, we can assume that the statistical behavior of the interference is Gaussian and get a reference threshold from (11) by adjusting $\sigma^2_{T_Q}$. Unfortunately, the interference power is usually unknown and the threshold can only be found heuristically. In practice, we can obtain the threshold corresponding to a specific false alarm probability by assuming the maximum possible interference. Thus the threshold will work for all the possible scenarios. In this paper, we set the interference signal power in the lower adjacent channel to -28 dBm, the maximum level used in [3], and determine the threshold by a heuristic method. For example, we can run 10000 simulations in which only noise and interference exist. The resulting statistics are recorded and sorted in decreasing order. For $P_{FA} = 1\%$, the 101^{th} value of the 10000 sorted statistics is the required threshold.

III. Spectrum Sensing for Wireless Microphone Signals

In the United States, wireless microphones are low-power secondary licensed signals operating in the locally unused TV bands regulated by FCC Radio Broadcast Rules in Title 47 Codes of Federal Regulations (CFR), Part 74 (47cfr74). There are four main regulations for wireless microphone usage: (1) The wireless microphones are allowed to operate in unused VHF or UHF TV bands listed in 47cfr74. (2) The frequency selection shall be offset from the upper or lower band limits by 25 kHz or an integral multiple thereof. (3) One or more adjacent 25 kHz segments within the assignable frequencies may be combined to form a channel whose maximum bandwidth shall not exceed 200 kHz. (4) The maximum transmitter power is 50 mW in VHF bands and 250 mW in UHF bands. In other countries, wireless microphone operations are regulated by different agencies, but with technical characteristics similar to those that apply in the United States. Most of the wireless microphone devices use analog Frequency Modulation (FM) although other types of modulations are permitted. There are a few digital and hybrid analog/digital systems on the market. It is difficult to use signal features e.g., pilot symbols, to sense digital wireless microphone signals because there is no common transmission standard for digital wireless microphones. Blind spectrum sensing method, e.g., Eigenvalue-Based algorithms [12], can be applied to sense a wireless microphone signal regardless of its modulation type. Another method is to look for the spectrum peak [16]. The bandwidth of wireless microphone signals is less than 200 kHz which is small compared with a TV band of 6 MHz. As a result, the power of wireless microphone signals is very concentrated while the noise power is uniformly distributed over the whole 6 MHz band. However, both methods fail when a strong adjacent channel interference is presented and the wireless microphone operates in the channel edge. The problem of sensing wireless microphone signal with the presence of adjacent channel interference is much more difficult than that of TV signals. The center frequency of a wireless microphone may be only 50 kHz from the adjacent channel edge in the Two-Signal test model [3]. Thus, the wireless microphone signal is fully masked by the adjacent channel interference. In this paper, a spectrum sensing algorithm for wireless microphone signals using FM modulation is described. The proposed method can determine the presence of FM-based wireless microphone signals even with strong adjacent channel interference. The problem of sensing digital wireless microphone signals with strong channel interference is still an open research area.

A. Autocorrelation Function of FM Signals

Frequency modulation is an analog modulation scheme. The frequency of the sinusoidal carrier wave is varied in accordance with the baseband signal. The FM signal $x(t)$ can be described by [20]

$$x(t) = A_c \cos\left[2\pi f_c t + 2\pi\Delta f \int_0^t m(u)du + \theta\right] \tag{12}$$

where θ is a random phase uniformly distributed on $(0, 2\pi)$ and $m(t)$ is the transmitted voice signal. It is zero-mean and its amplitude $|m(t)| \leq 1$. The parameters A_c and f_c are carrier amplitude and carrier frequency, respectively. The constant Δf is the frequency deviation of an FM modulator, representing the maximum departure of the instantaneous frequency of the FM signal from the carrier frequency f_c. In addition, it can be shown that the autocorrelation function of the FM signal, $x(t)$ is given by

$$\begin{aligned} R_x(\tau) &= E[x(t+\tau)x(t)] \\ &= \frac{A_c^2}{2} E\left[\cos\left(2\pi f_c \tau + 2\pi\Delta f \int_t^{t+\tau} m(u)du\right)\right] \end{aligned} \tag{13}$$

where the first expectation is over θ and $m(t)$ while the second expectation is over $m(t)$. The integral term inside the cosine function has a maximum value of $2\pi\Delta f\tau$. Several wireless microphone simulation models are described in [21] and the maximum frequency deviation suggested is 32.6 kHz. The carrier frequency f_c is in the order of MHz. For example, let's have $f_c = 3.26$ MHz which is 100 times of the maximum value of Δf. For a period of $0 \leq \tau \leq 10$ μs, the phase variation caused by the carrier frequency is 65.2π (32.6 cycles) while only a maximum of about 0.6π is contributed by the integral term at $\tau = 10$ μs. Therefore, when $f_c \gg \Delta f$ and the correlation delay τ is small, the phase variation is dominated by the carrier frequency and the contribution of the integral term can be ignored. Based on the observations above, we have

$$R_x(\tau) \simeq \frac{A_c^2}{2}\cos(2\pi f_c \tau) \tag{14}$$

given that $f_c \gg \Delta f$ and τ is small.

B. Spectrum Sensing Algorithms Without Interference

Assume that the received signal is $r(t)$, given as

$$r(t) = x(t) + w(t) \tag{15}$$

where $w(t)$ is an additive white Gaussian noise (AWGN). The signal $r(t)$ is sampled at a sampling frequency of f_s by an Analog-to-Digital Converter (ADC) to obtain $r[n] = r(n/f_s)$. The autocorrelation function is computed by

$$R_r[m] = \frac{1}{N_r}\sum_{n=0}^{N_r-1} r[n+m]\cdot r[n] \tag{16}$$

where N_r is the number of samples used to compute $R_r[m]$. Note that the estimated autocorrelation function given in (16) is computed by averaging over the same number of lag products. That means not all available signal samples are utilized to compute estimated autocorrelation function. By doing so, the sample autocorrelation function, $R_r[m]$ will have the same variance for different correlation delay m. The formulation of the threshold setting given in Section III-D will be simplified. Note that the accuracy of the estimated autocorrelation function is not affected because N_r is much larger than the largest correlation delay. Since the FM signal $x(t)$ and noise $w(t)$ are both zero-mean and they are independent, the autocorrelation function of the received signal $r[n]$ consists of the sum of the autocorrelation functions of these two signals,

$$\begin{aligned} R_r[m] &= R_x[m] + R_w[m] \\ &\simeq \frac{A_c^2}{2}\cos(2\pi f_c m/f_s) + R_w[m]. \end{aligned} \tag{17}$$

Note that ideally, the auto correlation function of the noise, $R_w[m]$, is zero for $m \neq 0$. In practice, although the values of $R_w[m]$ are not zero for $m \neq 0$, it is relatively small compared to the value of $R_w[0]$. If the carrier frequency of the FM signal is known, the optimal detector is a matched filter [19], i.e., the decision statistic of the optimal detector is given by

$$T_R(f_c) = \frac{1}{M}\sum_{m=1}^{M} R_r[m]\cdot\cos(2\pi f_c m/f_s). \tag{18}$$

where M is the number of $R_r[m]$ values computed for different m. However, the carrier frequency of a wireless microphone device can be any frequency within a TV channel as long as the carrier frequency offsets from TV channel edge is a multiple of 25 kHz. Assume that the received signal occupies a band from P MHz to $(P+6)$ MHz. The wireless microphone devices can select $f_0 = P$ MHz+50 kHz, $f_1 = P$ MHz+75 kHz,..., $f_{N_f-1} = (P+6)$ MHz$-$50 kHz, as its carrier frequency which is indicated in [3]. There are totally $N_f = 1+(6\text{ MHz}-100\text{ kHz})/(25\text{ kHz}) = 237$ possible carrier frequencies. As a result, the decision statistic of the optimal FM signal sensor is given by [19]

$$T_R = \max_{0\leq n\leq N_f-1} T_R(f_n). \tag{19}$$

C. Spectrum Sensing Algorithms with Interference

When the received signal $r(t)$ contains an interference signal $i(t)$, we have

$$r(t) = x(t) + i(t) + w(t). \tag{20}$$

Because the FM signal $x(t)$, interference signal $i(t)$, and noise $w(t)$ are zero-mean and mutually independent, the autocorrelation function of the sampled received signal $r[n]$

consists of the sum of the autocorrelation functions of these three signals,

$$\begin{aligned} R_r[m] &= R_x[m] + R_i[m] + R_w[m] \\ &\simeq \frac{A_c^2}{2}\cos(2\pi f_c m/f_s) + R_i[m] + R_w[m]. \end{aligned} \tag{21}$$

An important observation is that the autocorrelation function of the interference signal, $R_i[m]$, has significantly large values when the correlation delay m is small, and decrease when the correlation delay m increases. The sinusoid embedded in $R_r[m]$ is destroyed by $R_i[m]$ when the correlation delay m is small. It can be expected that for sufficiently large m, say $m \geq D$, $R_r[m]$ will reveal the sinusoidal property of the FM signal. Note that D depends on the statistic property of the adjacent channel interference and it is determined heuristically. Then, the decision statistic of an FM signal spectrum sensor is modified as

$$T_R(f_c) = \frac{1}{M}\sum_{m=D}^{M+D-1} R_r[m] \cdot \cos(2\pi f_c m/f_s). \tag{22}$$

for a known carrier frequency f_c. However, the approximation made in (14) is no longer accurate due to an increased correlation delay. In this situation, the spectrum sensing algorithm given in (22) only works for small frequency deviations and fails for large frequency deviations. To solve this problem, let's consider a higher order statistic given by

$$Z_x(\lambda) = \frac{1}{T}\int_{\tau=T_D}^{T_D+T} R_x(\tau+\lambda)R_x(\tau)d\tau. \tag{23}$$

The integration starts from T_D and the integration time is T. The function, $Z_x(\lambda)$ consists of two terms, given by (24) in the next page. When T is large enough, the second term approaches zero. As a result,

$$Z_x(\lambda) \simeq \frac{A_c^4}{4}\cos(2\pi f_c \lambda) \tag{25}$$

given that $f_c \gg \Delta f$ and λ is small for the same reason in obtaining (14). The higher order statistic is computed by

$$Z_r[k] = \frac{1}{M-k}\sum_{n=D}^{D+M-k-1} R_r[n+k]R_r[n] \tag{26}$$

and

$$Z_r[k] \simeq \frac{A_c^4}{4}\cos(2\pi f_c k/f_s) + Z_i[k] + Z_w[k]. \tag{27}$$

The decision statistic of the optimal FM detector is then given by

$$T_Z = \max_{0 \leq n \leq N_f-1} T_Z(f_n) \tag{28}$$

where

$$T_Z(f_n) = \frac{1}{K}\sum_{k=1}^{K} Z_r[k] \cdot \cos(2\pi f_n k/f_s). \tag{29}$$

Note that the autocorrelation-based sensing method for FM wireless microphone signals is actually a tone detector. Thus, it is sensitive to the narrow-band interference, e.g., random spikes in the environment. When implementing this method to sense FM wireless microphone signals in practice, we need to take care of the narrow-band interference.

D. Threshold Setting for A Specific Probability of False Alarm

1) AWGN Only: Consider that there is only AWGN noise present and the autocorrelation-based spectrum sensor specified in (19) is used to perform spectrum sensing. By substituting $r[n]$ with $w[n]$ in (16), we have

$$R_w[m] = \frac{1}{N_r}\sum_{n=0}^{N_r-1} w[n+m] \cdot w[n] \tag{30}$$

Assume that the noise variance is $E(w^2[n]) = \sigma^2$ and from the Central Limit Theorem, when N_r is sufficiently large, $R_w[m]$ is approaching to a zero-mean Gaussian random variable with a variance of $\frac{\sigma^4}{N_r}$. The random variables $\{R_w[m]\}_{m=D}^{M+D-1}$ are independently and identically distributed (i.i.d.) Gaussian Random variables. Since the addition of Gaussian random variables is still Gaussian, the random variables $\{T_R(f_n)\}_{n=0}^{N_f-1}$ are identically distributed zero-mean Gaussian random variables with a variance of $\sigma_{T_R}^2 = \frac{\sigma^4}{2MN_r}$. However, they are not independent. Although the covariance matrix of the random variables $\{T_R(f_n)\}_{n=0}^{N_f-1}$ can be obtained and hence their joint distribution function, it is still difficult to compute a reference threshold. Therefore, the independent approach used in II-D1 is adopted. We first assume $\{R_w[m]\}_{m=D}^{M+D-1}$ are independent to obtain the cumulative distribution function of the decision statistic defined in (19), given as

$$F_{T_R}(x : H_0) = \left(\int_{-\infty}^{x} \frac{1}{\sqrt{2\pi}\sigma_{T_R}} e^{-\frac{u^2}{2\sigma_{T_R}^2}} du\right)^{N_f}. \tag{31}$$

Then, for a particular P_{FA}, the corresponding threshold γ_{T_R} can be found by

$$P_{FA} = 1 - F_T(\gamma_{T_R} : H_0). \tag{32}$$

After some straightforward calculation, we have

$$\gamma_{T_R} = \mu_{T_R}\sigma_{T_R} \cdot Q^{-1}(1-(1-P_{FA})^{1/N_f}) \tag{33}$$

where $Q^{-1}(\cdot)$ is the inverse function of the function

$$Q(x) = \int_{x}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{1}{2}u^2} du. \tag{34}$$

Note that μ_{T_R} is an heuristic adjusting factor introduced to account for the independence assumption made in (31).

2) Interference Plus AWGN: When the adjacent channel interference is taken into consideration, the higher-order-statistic-based spectrum sensor specified in (28) is used to perform spectrum sensing. Because the statistic of the interference signal is unknown, the threshold can only be determined by a heuristic method described in Section II-D2.

IV. SIMULATION RESULTS

A. Simulation Model for TV Signals

Fig. 6 illustrates the simulation model for the TV signals. The TV signals which are converted to a lower center Intermediate Frequency (IF) are sampled at a rate of f_s to generate discrete signals $x[n]$. According to the Two-Signal (Adjacent Channel Interference) model [3], an interference signal from the lower TV channel is added. The signal power of the lower adjacent channel is -28 dBm. The additive white Gaussian noise (AWGN) $w[n]$ is then added to form the experimental

$$
\begin{aligned}
Z_x(\lambda) &= \frac{A_c^4}{4T}\int_{\tau=T_D}^{T_D+T}\cos\left(2\pi f_c\lambda + 2\pi\Delta f\int_{t+\tau}^{t+\tau+\lambda} m(u)du\right)d\tau \\
&+ \frac{A_c^4}{4T}\int_{\tau=T_D}^{T_D+T}\cos\left(4\pi f_c\tau + 2\pi f_c\lambda + 2\pi\Delta f\int_{t}^{t+\tau} m(u)du + 2\pi\Delta f\int_{t}^{t+\tau+\lambda} m(u)du\right)d\tau
\end{aligned} \tag{24}
$$

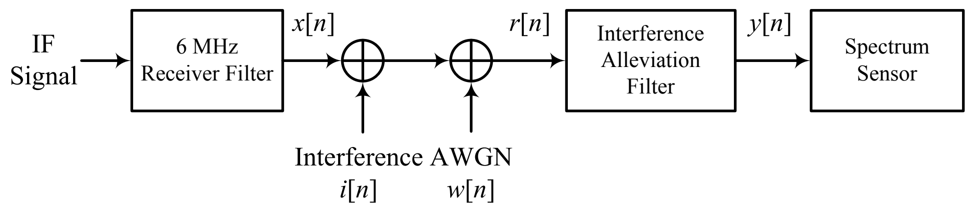

Fig. 6. Spectrum sensing simulation model for TV signals.

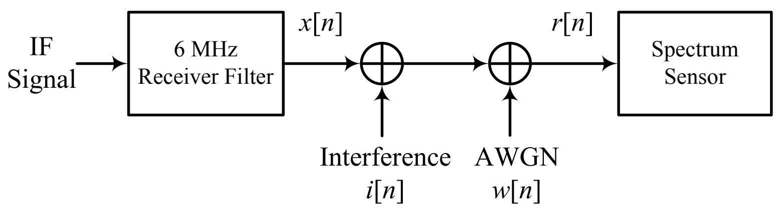

Fig. 7. Spectrum sensing simulation model for wireless microphone signals.

received signal $r[n]$. We will further assume that $w[n]$ is zero mean and the noise power spectral density (PSD) is $N_0 = -174 + 10 = -164$ dBm/Hz where -174 is thermal noise power spectral density under normal temperature condition and 10 is noise figure of the receiver [18]. Therefore, the noise power is $N_0 B = -164$ dBm/Hz·6MHz ≈ -96 dBm. Then, a single-sided complex filter centered at frequency f_{IF} is used to alleviate interference and generate complex signals $y[n]$. The bandwidth of this interference alleviation filter is less than 6 MHz, e.g., 5 MHz. Note that $y[n]$ is equivalent to a baseband signal with a frequency offset of f_{IF}. Since the proposed algorithm is immune of the frequency offset, the spectrum sensing algorithm is applied to $y[n]$. For ATSC DTV signals, field capture data files [17] are used to test the proposed sensing algorithms. The field captured signals have $f_{IF} = 5.38$ MHz and $f_s = 21.524476$ MHz [17]. Because an 8 MHz IF filter was used when capturing the signals, some captured signals include adjacent channel interference. Thus, a 6 MHz receiver filter, as shown in Fig. 6, is used to eliminate the interference. Furthermore, because $f_s = 21.524476$ MHz is 2 times of the ATSC symbol rate, $y[n]$ is decimated by 2 and then applied with the spectrum sensing algorithm. For NTSC analog TV signals, the parameters are slightly different. The IF and sampling frequencies are $f_{IF} = 5.6875$ MHz and $f_s = 15.75$ MHz, respectively. The sampling frequency $f_s = 15.75$ MHz is chosen so that each line contains 1001 samples.

B. Simulation Model for Wireless Microphone Signals

In order to reduce the implementation complexity, the sampling frequency used for ATSC DTV signals is adopted.

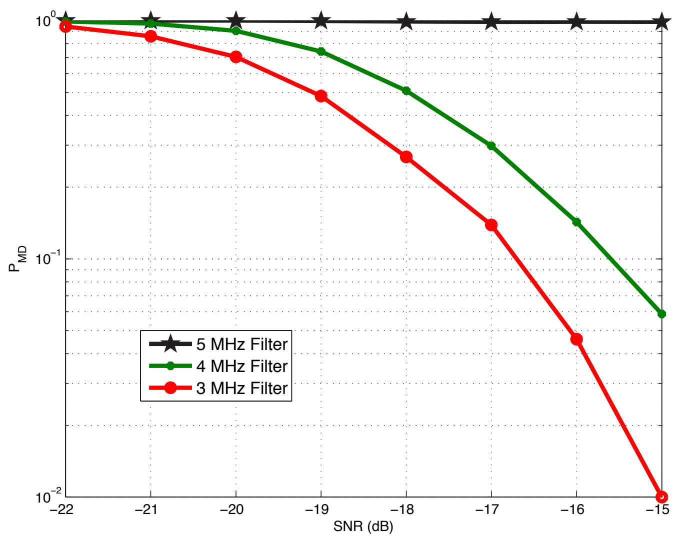

Fig. 8. Sensing performance comparison of ATSC DTV signals for alleviation filters with different bandwidth (interference signal power = -78 dBm)

Thus, the simulation model of wireless microphone signals as shown in Fig. 7 is similar to simulation model of the TV signals except that there is no interference filtering operation. The spectrum sensing algorithm is applied to the signal $r[n]$. Furthermore, the FM wireless microphone signals are generated by (12) using voice signal model in [24] and a frequency deviation factor $\Delta f = \pm 32.6$ kHz which is the maximal one suggested in [21].

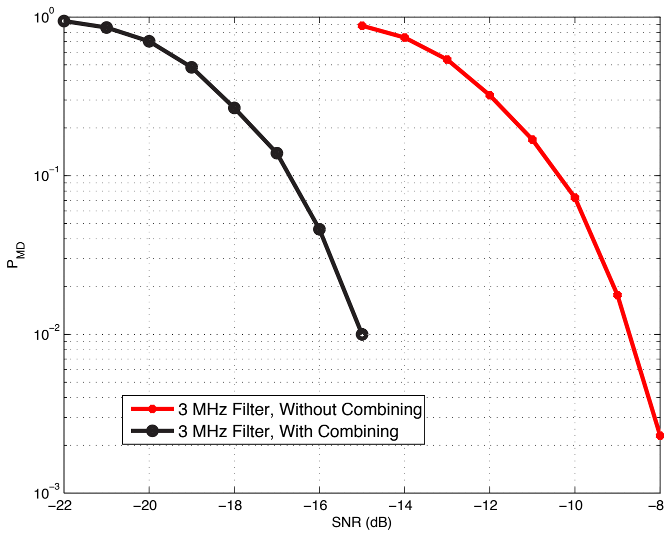

Fig. 9. Sensing performance comparison of (2) and (4) for ATSC DTV signals (interference signal power = -78 dBm)

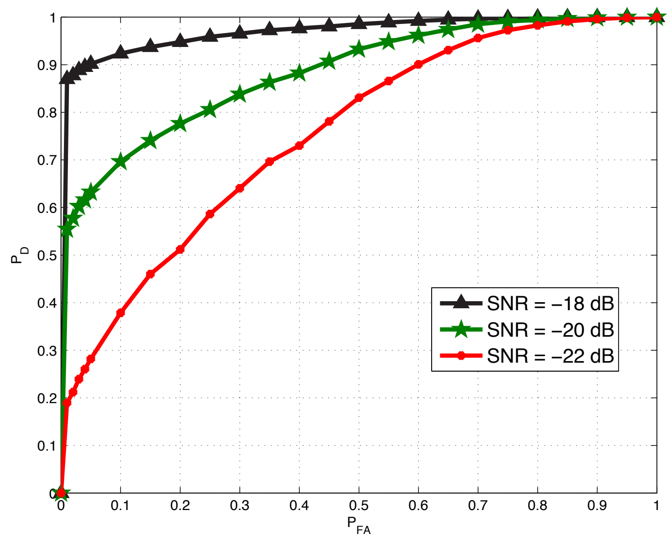

Fig. 11. Family of ROC curves for ATSC DTV signals at different levels of SNR (interference signal power = -83 dBm, filter BW = 5 MHz)

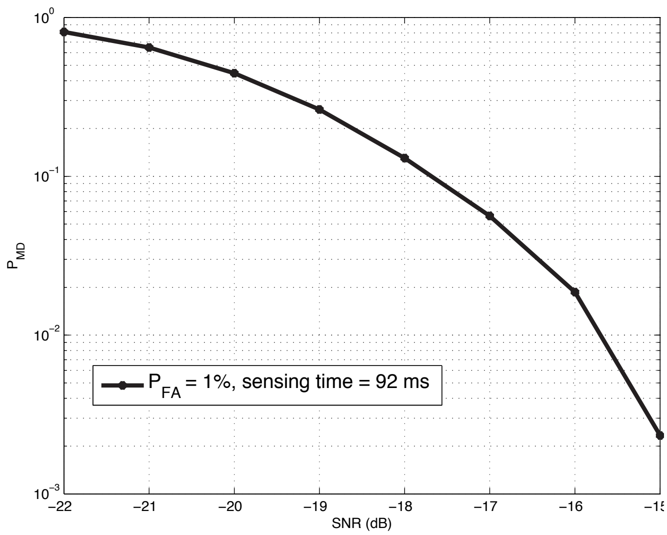

Fig. 10. Sensing Performance of ATSC DTV signals (interference signal power = -83 dBm, filter BW = 5 MHz)

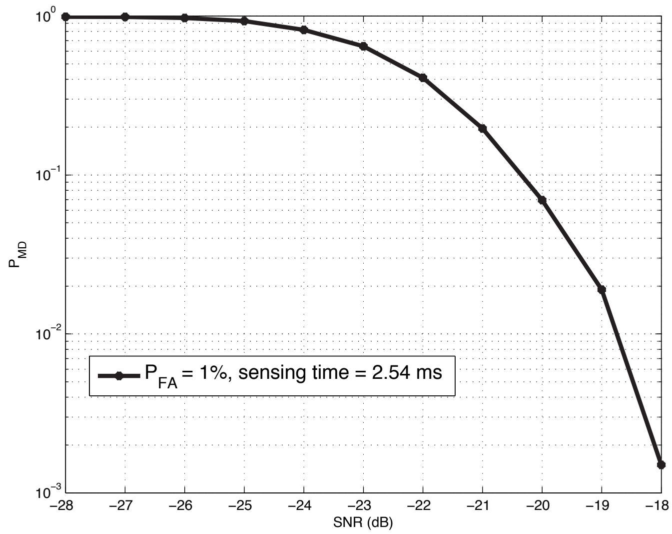

Fig. 12. Sensing Performance of NTSC signals (interference signal power = -83 dBm, filter BW = 5 MHz)

C. Spectrum Sensing Performances

For ATSC DTV signals, 12 referenced field capture data files [17] are used to evaluate the proposed sensing algorithms. Furthermore, the sensing time is 92 ms which is equal to 1200 data segments and the decision statistic, T_Q, is computed by combining accumulated SYNC correlation functions $C(m,s)$ for $s = 1, 2, ..., 1000$, i.e., $N_Q = 999$. For NTSC analog TV signals, there is no field capture data available. The required test data is generated by sampling the NTSC format video in the Lab. The sensing time is 2.54 ms which is equal to 40 lines and the decision statistic, T_Q, is computed by combining accumulated SYNC correlation functions $C(m,s)$ for $s = 1, 2, ..., 20$. Figure 8 shows the sensing performance for ATSC DTV signals with interference alleviation filters of different bandwidth for $P_{FA} = 1\%$. The units of the x-axis and y-axis are SNR and probability of missdetection (P_{MD}), respectively. In this simulation, the total out-of-band emission (interference) power is -78 dBm which is 50 dB lower than the original signal power. It is a very strong interference. The interference-plus-noise power ratio (SINR) is an extremely low value of -36 dB. We can see from the figure that the interference alleviation filter can effectively reduce the interference. This figure demonstrates that the proposed sensing algorithm for ATSC or NTSC signals can handle very strong adjacent channel interference by adjusting the bandwidth of the interference alleviation filter. Figure 9 compares the sensing performances of $T_C(1)$ given in (2) and T_Q in (4) for a sensing time of 92 ms and $P_{FA} = 1\%$. The interference

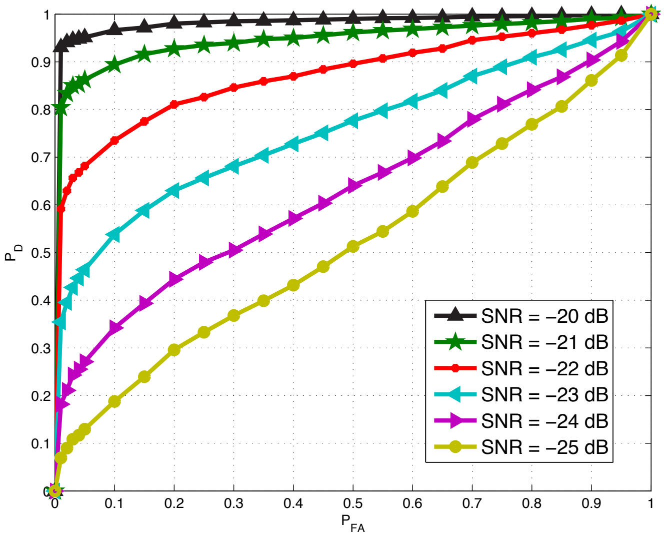

Fig. 13. Family of ROC curves for NTSC signals at different levels of SNR (interference signal power = -83 dBm, filter BW = 5 MHz)

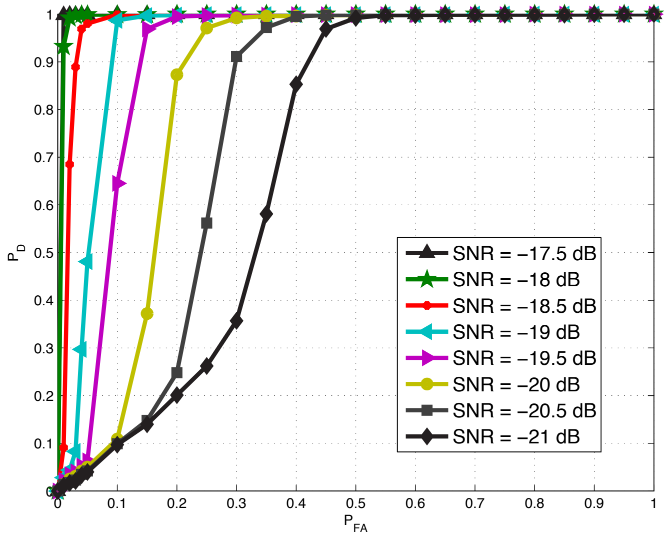

Fig. 15. Family of ROC curves for FM wireless microphone signals at different levels of SNR (interference signal power = -83 dBm)

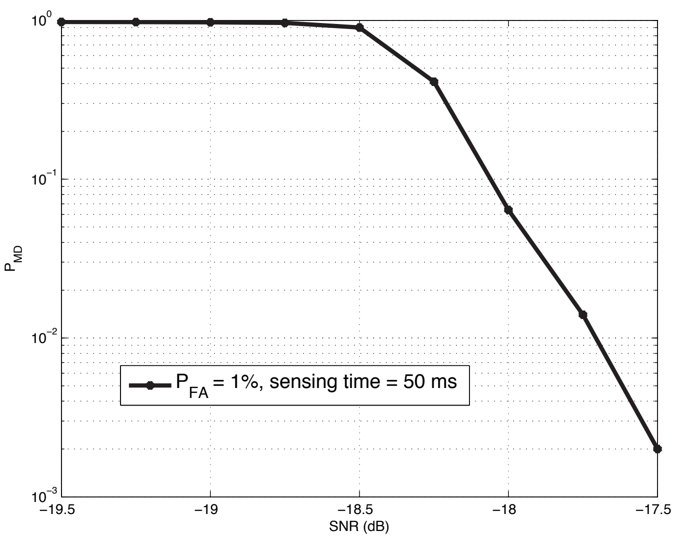

Fig. 14. Sensing performance of FM wireless microphone signals (interference signal power = -83 dBm)

power is -78 dBm and a 3 MHz interference alleviation filter is used. We can see that by using T_Q to combine multiple $T_C(s)$ for different value of s, the performance improvement is conspicuous. The required SNR to achieve $PMD = 0.1$ for T_Q is about 6 dB less than using $T_C(1)$ as a decision statistic. Figures 10 and 12 show the sensing performance of ATSC DTV signals and NTSC analog TV signals for $P_{FA} = 1\%$. Note that the interference power is -83 dBm which is 55 dB less than the original signal power. In this level of interference, a 5 MHz filter is able to handle interference. The sensitivity of -114 dBm set by FCC is about -31 dB in terms of SINR. To achieve $P_{MD} = 0.1$, the required SNR is -18 dB for ATSC DTV signals. For NTSC signals in the same situation, it only takes a sensing time of 2.54 ms and a SNR of -20.5 dB to achieve $P_{MD} = 0.1$. Thus, spectrum sensing for NTSC signals is much easier than that of ATSC signals, which is predicted in our discussion in Section II-B. A family of Receiver Operating Characteristic (ROC) curves at different levels of SNR is plotted to summarize the sensing performance of a detector. They are given in Fig.s' 11 and 13 for the proposed SYNC-correlation-based spectrum sensors of ATSC DTV and NTSC analog TV signals, respectively. A reliable spectrum sensor should achieve a low P_{MD} with respect to a low P_{FA}. The sensing requirement defined in IEEE 802.22 WRAN Standard is that a spectrum sensor should detect the incumbent signal with $P_D = 0.9$ ($P_{MD} = 0.1$) subject to $P_{FA} = 0.1$ when SNR is -20 dB. Furthermore, the sensing threshold specified by FCC R&O [1] is -114 dBm (-18 dB of SNR). Thus, we say a spectrum sensor is reliable if it can achieve $P_{FA} = 0.1$ and $P_{MD} = 0.1$ when SNR is -18 dB. From the ROC curves, the proposed ATSC and NTSC spectrum sensors achieve $P_{MD} = 0.1$ when SNR is -18 and -20 dB, respectively. However, in our simulations, we choose a much smaller false alarm probability.

Fig. 14 shows the sensing performance of a higher order statistic detector specified in (28) for FM wireless microphone signals with interference and $P_{FA} = 1\%$ with a sensing time of 50 ms ($N_f = 1076000$). The parameters D, M and K are 200, 500, and 200. The requires SNR to achieve $P_{MD} < 0.1$ is about -18 dB. The ROC curve is plotted in Fig. 15 for wireless microphone signals. From the ROC curves, the proposed FM wireless microphone spectrum sensor is reliable when SNR is -18 dB.

V. Conclusions

Spectrum sensing for TV signals including ATSC DTV and NTSC analog TV signals as well as FM wireless microphone signals are presented in this paper. Simulation results reveal that the proposed spectrum sensors can determine the availability of a TV channel for unlicensed radio transmissions when the SINR is as low as -31 dB assuming a 5 MHz interference alleviation filter is used for sensing ATSC and NTSC signals. In our knowledge, the sensing algorithms presented are the first known spectrum sensing algorithms for TV and wireless microphone signals which achieve the sensing requirement specified by FCC in the presence of a strong adjacent channel interference.

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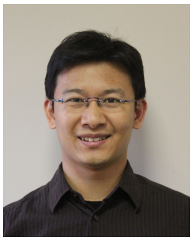
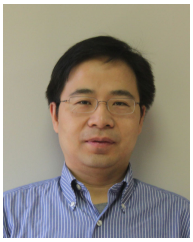


Hou-Shin Chen (M'08) received his B.S. and M.S. degree in electrical engineering from National Taiwan University, Taiwan in 2002 and 2004, respectively. He received the Ph.D. degree in electrical and computer engineering from Rutgers University, USA in 2008. He joined Corporate Research of Technicolor Inc. in Princeton in March 2008 as a Technical Staff Member. His current research interests include cognitive radio communication systems and detection and estimation theory.



Wen Gao (M'01)received his Ph.D. and M.S. degree in electrical engineering from Purdue University in 2001 and 1998 respectively. He joined Corporate Research of Technicolor Inc. in Princeton in July 2001 as a Technical Staff Member. He became a Senior Technical Staff Member in 2003 and Communications Lab manager in 2007. His current research interests include cognitive radio, MIMO cooperative network and error control coding. He also has extensive experiences in system design and hardware development.