Novel Filter Banks Based Wireless Microphone Detection in IEEE 802.22 WRAN

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Abstract— **The primary task in any cognitive radio (CR) network is to dynamically explore the spectrum and determine spectrum band that can be used without causing interference to other users. In this paper, we propose a multistage filter banks to sense the spectrum band. Compared with the conventional filter banks, our proposed method has low complexity but high detection precision. The proposed scheme has been analyzed and tested through simulations on wireless microphone detection in IEEE 802.22 WRAN.**

Keywords: **cognitive radio, multistage, filter banks, IEEE 802.22 WRAN, wireless microphone**

I. INTRODUCTION

Cognitive radio (CR) is a revolutionary intelligence technology which can maximize the utilization of the spectrum band by allowing the second users (SU) access the spectrum holes without causing interference to primary users [1-2]. IEEE 802.22 WRAN is the first worldwide Wireless Regional Area Networks (WRAN) standard based on cognitive radios (CR) [3] [4]. In IEEE 802.22, second users can operate in the licensed VHF/UHF bands. An efficient and reliable spectrum sensing and detection scheme is a key feature that IEEE 802.22 CR air-interface must support.

Application of filter banks techniques in spectrum sensing in CR has been studied in [7]. In reference [8], ployphase DFT filter banks are proposed for spectrum detection, in which the TV channel is divided into 6 subbands and the filter banks presented in [8] are used to detect 5 continuous TV channels. Furthermore, reference [9] uses DFT filter bank (DFTFB) to detect the WM. If WM spans two filters, reference [9] uses the cyclostationary property to detect the precise frequency. But the complexity is very high when using DFTFB, not to mention adding the cyclostationary property detection. Multiple-stage coefficient decimation filter bank (MS-CDFB) is proposed in [10]. Although the MS-CDFB approach can reduce the complexity, it is suitable for detecting wide-band spectrum only. In this paper, we propose a multistage filter banks which is very suitable for narrow band users detection to sense the spectrum bands. Compared with conventional filter banks, our proposed scheme has low complexity but high detection precision.

The rest of this paper is organized as follows. Section II describes the spectrum sensing using the conventional DFTFB. Our proposed MS-DFTFB is presented in Section III-A, which is followed by the relevant computational complexity analysis in Section III-B. Simulations of MS-DFTFB on wireless microphone detection in IEEE 802.22 WRAN will be given in section IV. Final conclusions are discussed in Section V .

II. SPECTRUM SENSING USING CONVENTIONAL FILTER BANKS

The filter banks system often involves two processes: separation of the frequency components and recombination of the components to recover the original signal. The separation process is known as analysis filter banks and the recombination process is known as synthesis filter banks. Analysis and synthesis filters often appear in pairs and satisfy the perfect reconstruction condition which guarantees the perfect reconstruction of the signal. In the area of wireless communications and networks, we usually utilize the analysis filter banks to sense the spectrum environment. Filter banks are often implemented based on a prototype filter [11]. The prototype filter is a low pass filter, which is also used as the 0th band of the filter banks. As shown in Fig. 2, the other bands can be formed through modulation of the prototype filter. M band filter banks can be formed as

$$h_k = h(n)e^{j\frac{2\pi}{M}kn}, \quad k = 0,1,\ldots M-1, \tag{1}$$

where $h(n)$ is the prototype filter of the filter banks.

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Filter banks can use polyphase structure to reduce complexity. The i^{th} band of M band filter banks can be organized as

$$H_i(z)=\sum_{l=0}^{M-1}(zW_M^i)^l E_l(z^M), \tag{2}$$

where $W_M^i = e^{-j2\pi i/M}$, $E_l(z)$ is the l^{th} polyphase component of $H(z)$, which is defined as

$$E_l(z)=\sum_{n=0}^{N-1}h(Mn+l)z^{-n}. \tag{3}$$

Fig. 3 expresses a typical polyphase DFT filter banks structure (DFTFB) [7]. Here, the M bands of the filter banks share the same structure. $y_i(n),\ i=0,1,...,M-1$ are the outputs of the M filters. By computing the energy of $y_i(n)$, we can acquire the presence of the primary user (PU). The computational complexity of the polyphase structure is equivalent to the realization of the prototype filter and one M point IDFT.

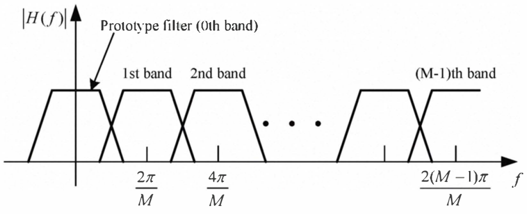


Fig. 2. M band filter banks.

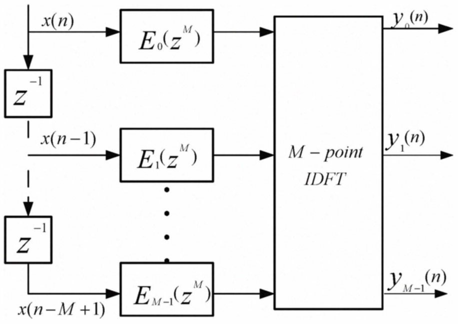


Fig. 3. Polyphase structure of M band DFT filter banks.

III-A. PROPOSED MULTISTAGE FILTER BANKS

The detection precision is directly affected by the parameter M, which is the number of subbands of the M band filter banks. If we want higher detection precision, we need to increase the value of M, which will cause the increasing of computational complexity. By dividing the DFTFB into several stages, the computational complexity can be reduced while the detection precision will increase. The main process of our proposed method is as following, which is named multistage DFT filter banks (MS-DFTFB):

Stage 1: we first use a M_1 band DFTFB to sense the spectrum that we interest in. If there are narrow band users existing in the spectrum band and the detection result is not precise enough, we carry out stage 2;

Stage 2: we use a M_2 band DFTFB to detect the narrow band user we interest in (we use $x_{in}(n)$ to denote our target signal below.) based on the result of stage 1. We divide the subband which has detected $x_{in}(n)$ in stage 1 into M_2 subbands. These M_2 subbands can build up a M_2 DFTFB, which has much narrower subband comparing with M_1 DFTFB. If the narrow band signal spans two or more subbands, for example SU2, as presented in Fig. 4, we divide these adjacent subbands into M_2 subbands and then build up a M_2 band DFTFB.

If the detection precision is still not good enough, we can carry out *stage 3*, *stage 4*, ... *stage N*, until the detection precision achieve our requirement. The process is just as same as stage 1 to stage 2.

Before we carry out our analysis on MS-DFTFB, we need to make some useful definitions. Assume the bandwidth of the spectrum we want to sense is W_0 and the frequency of the spectrum band ranges from 0 to W_0 . (In fact, by using a down conversion, we can change any spectrum band into baseband.) We define S_i as the number of subbands that $x_{in}(n)$ spans in the M_i band DFTFB of stage i, $i=1,2,...,N-1$. In Fig. 4, for example, $S_1=1$ for SU1 and $S_1=2$ for SU2. N is the number of total stages of MS-DFTFB. Obviously, the bandwidth of subband in M_1 band DFTFB is W_0/M_1 . The bandwidth of the subband in *ith* stage's M_i band DFTFB (W_{sub}^i) is given by equation (4):

$$W_{sub}^i = S_{i-1}W_{i-1}/(M_{i-1}\times M_i),\ i=2,3,...,N \ . \tag{4}$$

The center frequency of $x_{in}(n)$ (f_{cx}^i) is decided by the subbands that detected $x_{in}(n)$ in *ith* stage, which are assumed *kth* to the $(k+S_i-1)th$ subbands of the M_i band DFTFB, as presented in equation (5):

$$f_{cx}^i = f_k + \frac{W_i}{2\pi M_i}\times\frac{S_i}{2} \quad i=1,2,...,N, \tag{5}$$

where f_k is the start frequency of the *kth* subband.

The detailed method of dividing the interested subband(s) of M_{i-1} ($i>1$) band DFTFB into M_i subbands and build up a M_i band DFTFB is as following (*ith* stage):

We assume the lowest index of the subbands that have detected $x_{in}(n)$ is k, which means the $x_{in}(n)$ spans from the kth to the $(k+S_{i-1}-1)th$ subbands of the M_{i-1} band DFTFB. Then we divide these S_{i-1} subbands into M_i subbands and build up a M_i band DFTFB. We assume the types of the prototype filters of M_i $i=1,2,...,N$ are same except the bandwidth. The initial start frequency of M_i band DFTFB is 0 as shown in Fig. 2. In order to sense the kth to $(k+S_{i-1}-1)th$ subbands of the M_{i-1} band DFTFB, we need to modulate the center frequency of M_i band DFTFB onto the center frequency of those subbands of M_{i-1} band DFTFB. This can be realized by multiplying a modulation component after each polyphase component $E_i(z)$, $i=0,1,...M_i-1$ in Fig. 3. In *ith* stage, the modulation component (C_m^i) is defined as $C_m^i = \exp(j(2k-1)W_{sub}^{i-1}/2)$.

The whole structure of our proposed scheme is presented in Fig. 5(a). There are N stages and each stage (except stage 1) has two input flows. One is the detection information, which is used to set the modulation component. The other is the SU signal ($x_{in}(n)$) we want to detect from the antenna. Fig. 5(b) is the structure of the M_i band DFTFB in *ith* stage. In Fig. 5(b), $\exp(j(2k-1)W_{sub}^{i-1}/2)$ is the modulation component of the *ith* stage.

III-B. COMPLEXITY AND DETECTION PRECISION ANALYSIS

As multiplication is the most complex operation in filtering, we use the number of complex multiplication to estimate the complexity. Corresponding to the conventional DFTFB mentioned above, the number of complex multiplication is given by (6)

$$N_{DFTFB} = length(h_0(n)) + \frac{M}{2}\log_2 M, \tag{6}$$

where $length(h_0(n))$ is the points of the prototype filter and M is the number of total subbands. We choose $M=2^l$, l is an integer which guarantees basic radix-2 IFFT can be used in the structure of DFTFB as shown in Fig. 3. In order to avoid aliasing in the M band DFTFB, it is necessary to ensure the points of prototype filter be bigger than M. In this paper, we choose the length of prototype filter α times as the total bands of DFTFB, namely $length(h_0(n)) = \alpha M, \alpha \ge 1$ .While the number of complex multiplications our proposed MS-DFTFB needs is $N_{MS-DFTFB} = \sum_{i=1}^{N} N_i$ $i=1,2,...,N$, where N_i is the number of multiplications that *ith* stage needs as equation (7):

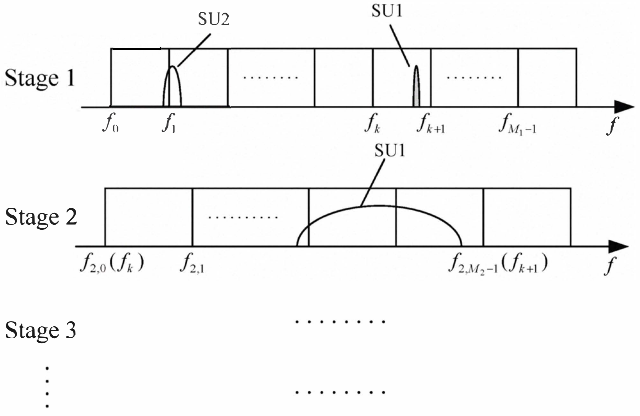
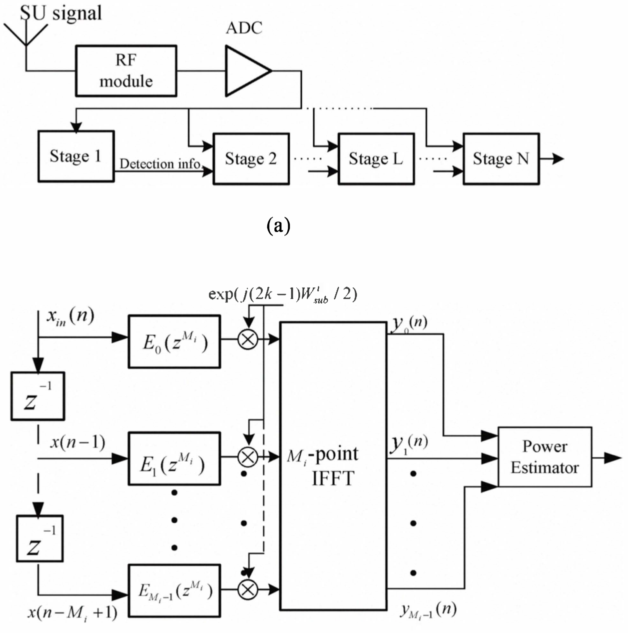


Fig.4 The scheme of proposed multistage DFTFB



(a)

(b)

Fig. 5(a) Brief structure of proposed multistage DFT filter banks. Fig. 5(b) Structure of DFT filter banks with modulation component in stage L.

$$N_i = \begin{cases} \alpha M_1 + \frac{M_1}{2}\log_2 M_1 & i=1 \\ 2\alpha M_i + \frac{M_i}{2}\log_2 M_i & i=2,3,...,N \end{cases}. \tag{7}$$

Comparing with equation (6) and (7), there is a factor 2 multiplying αM_i , this is because the modulation of the prototype filter cost αM_i complex multiplications in *ith* stage.

For the conventional DFTFB, the detection precision is directly affected by the bandwidth of subband. The detection precision will increase while the number of bands, which is M, is increasing. But the increasing of M will also lead to the addition of complex multiplications as presented in equation (6). Obviously, the MS-DFTFB with N stages is equivalent to a $M_E = \prod_{i=1}^{N} M_i / \prod_{i=1}^{N-1} S_i$ band DFTFB. The detection precision of N stages MS-DFTFB is equal to a conventional M_E band DFTFB. But the complex computations the two methods need are different, which will be analyzed in section IV in the scenario of wireless microphone detection in IEEE 802.22 WRAN.

IV . WIRELESS MICROPHONE DETECTION USING MS-DFTFB

In 802.22 WRAN, at any time when there is wireless microphone (WM) appearing in TV channel, the whole channel of 6 MHz bandwidth should be evacuated immediately for interference avoidance. In order to avoid interference to WM when other users using the adjacent TV channel, we need to know the precise location of WM. Usually, the bandwidth of WM is 200 KHz or less. Conventional DFTFB divide the whole 6 MHz TV channel into 30 or more subbands. By detecting the energy of each subband, we can locate the occupied frequency of WM. We can prove that the largest detection error (LDE) of a N stages DFTFB is $W_{sub}^{N} / 2$. In our simulations, we use LDE to estimate the detection precision.

Table.1 and Table. 2 are the simulation results of our proposed MS-DFTFB in WM detection. In our simulations, we use a sequence of WM with random center frequency and a fixed bandwidth of 200 KHZ. In Table. 1 and 2, N_{cm} is the number of complexity multiplications that MS-DFTFB needs. While N_{cm}^{c} is the number of complications that conventional DFTFB whose total number of subbands is a nearest power-of-two integer less than $M_1 \times M_2$ ($M_1 \times M_2 \times M_3$) needs. And we choose the parameter α equal to 4. Table. 1 shows the performances of 2 stages DFTFB. From Table. 1, we can find that the 2 stages DFTFB has higher detection precision but lower computational complexity comparing with the conventional DFTFB. For example, when $M_1 = 8$, $M_2 = 8$, the 2 stages DFTFB is equal to a 54 band DFTFB, but the N_{cm} this 2 stages DFTFB needs is 120, less than the 32 band conventional DFTFB. The LDE, however, is 55.5 KHz, smaller than the 32 band DFTFB's. $\overline{M_E}$ of 2 stages DFTFB of 8×16 and 16×8 are different because the value of $\overline{S_1}$ is determined by M_1. In Table. 1, $\overline{M_E}$ =108 when $M_1 \times M_2 = 8 \times 16$ while $\overline{M_E}$ =86 at the parameters $M_1 = 16$, $M_2 = 8$. This means the LDE of former scheme is higher but the N_{cm} is bigger too. This is because the first stage do not need the modulation process which can reduce the complex multiplications. Table. 2 is the simulation results of 3 stages DFTFB. Comparing with Table. 1 and Table. 2, we can find that the performance of 3 stages DFTFB is better than 2 stages DFTFB regardless the detection time delay.

Table. 1 Simulation results of 2 stages DFTFB

$M_1 \times M_2$	8×4	8×8	8×16	16×8
$\overline{S_1}$	1.17	1.17	1.17	1.48
$\overline{M_E}$	27	54	108	86
N_{cm}	80	120	204	172
N_{cm}^{c}	96(16)	208(32)	448(64)	448(64)
$LDE(KHz)$	111	55.5	27.5	35
$LDE^{c}(KHz)$	187.5	93.7	47	47

Table. 2 Simulation results of 3 stages DFTFB

$M_1 \times M_2 \times M_3$	$4 \times 4 \times 8$	$4 \times 8 \times 8$	$8 \times 8 \times 4$	$8 \times 8 \times 8$
$\overline{S_1 \times S_2}$	1.64	2.1	2.4	2.4
$\overline{M_E}$	78	116	106	213
N_{cm}	132	172	156	196
N_{cm}^{c}	448(64)	448(64)	448(64)	960(128)
$LDE(KHz)$	38	25	28	14
$LDE^{c}(KHz)$	47	47	47	23.5

V. CONCLUSIONS

In this paper, we proposed a novel filter banks named multistage DFT filter banks (MS-DFTF). Comparing with the conventional DFTFB, our proposed MS-DFTFB has higher detection precision but lower computational complexity. The proposed scheme has been analyzed and tested through simulations on wireless microphone detection in IEEE 802.22 WRAN.

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