Design of a Narrow Transition Band Dynamic Digital Channelized Receiver without Merging Adjacent Sub-channels

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Abstract. In electronic reconnaissance, effective channelization of several input signals with unknown parameters is the key requirement of the broadband receiver. Generally, since the prototype filter must be perfect, the dynamic channelized receiver based on merging adjacent sub-channels to reconstruct non-uniform filter Banks occupies too many resources and is difficult to realize. This paper proposed a sharp wideband channelized receiver based on DFT filter banks and interpolated filter banks. The proposed receiver consists of an analysis section, channel detection and arbitration section, and a synthesis section. The analysis section divides the wideband signal into multiple sub-channels. Channel detection and arbitration section output the center frequencies and the bandwidths of the input signals. In the synthesis section, after being converted to baseband, the input signals are filtered and decimated to achieve the low rate signals. The proposed method released the design of the prototype filter from the requirement of perfect reconstruction. So it is computationally efficient. Besides, time-multiplexed and interpolated filter techniques can be used to further reduce the hardware complexity. Simulation results verify the validity of the proposed method.

1. Introduction

Digital channelized receiver is widely used in the interception of wideband RF signal as the excellent characteristics including high resolution, high dynamic range, wide instantaneous bandwidth, and simultaneous signal processing. In the complicated electromagnetic environment, the number of signals, bandwidths, and center frequencies are all unknown. Normally, it is called a blind signal. For blind signal receiving, the receiver is required to have the ability of dynamic reception.

A common way to implement a channelized receiver is to feed the received signal into a filter bank (FB). In the literatures, various types of filter banks are presented for the design of the channelized receiver. These FBs can be generally classified into two categories: uniform FB and non-uniform FB.

The channelized receiver based on a discrete Fourier transform filter bank (DFTFB) can process the wideband signal efficiently [1]. However, it cannot extract signals with different bandwidths since the bandwidth of sub-channels is fixed. To extract the blind signal with different bandwidths, the uniformed FB must be designed. In [2], a new non-uniform FB by merging the adjacent sub-channels of a uniform FB is proposed. It consists of an analysis section and a synthesis section. The analysis section usually divides the signal into uniform sub-channels using cosine modulated filter Banks (CMFB) or complex exponentially modulated filter Banks (CEMFB) or DFTFB. The synthesis section employs the coefficient decimation method (CDM) to design the non-uniform FB with different bandwidths. This method can theoretically realize dynamic channelization. Based on this, many
approaches have been proposed [3-7]. Current methods for implement the dynamic channelized receiver are afflicted by one or more of the following shortcomings:

1) The prototype filter of CMFB/CEMFB must be perfectly reconstructed, which makes the design difficult [4-6].

2) CMFB/CEMFB operates at the sampling rate of ADC which brings many difficulties for real-time implementation [4-6].

3) The constituent filters in CMFB/CEMFB do not have a linear phase [4-6].

4) The method does not give a method to detect the existence of signal [4, 6].

5) The synthesis filter is obtained by decimation which cannot ensure high resolution as the transition bandwidth increases with the decimation rates. Meanwhile, the order of the synthesis filter is high. As a result, it is difficult to implement filter operation [3-7].

6) The hardware complexity is high, especially in the case of sharp filters [4-6].

Since the non-uniform FB is obtained by merging the adjacent sub-channels, the prototype filter of analysis FB must be perfectly reconstructed. Otherwise, the reconstruction error will occur. This leads to a high order of the filter. Near perfect reconstruction FB (NPRFB) can reduce the hardware complexity. However, NPRFB brings spectrum aliasing and amplitude distortion.

This paper proposed an efficient dynamic channelized receiver which can extract multiple blind signals without merging sub-channels. Firstly, the input signal is divided into even sub-channels and odd sub-channels by a pair of complementary periodic filters. Secondly, the even sub-channels and odd sub-channels are filtered by two uniform FBs. The uniform FBs can be implemented using polyphase architecture to reduce the hardware complexity. Thirdly, energy-based channel detection and arbitration are performed. The center frequencies and bandwidths of the input signal will be outputted. Finally, the input signal is converted down to baseband according to the center frequencies. Then the baseband signals are filtered by unified low pass filters (LPFs). The unified LPFs are designed with the same transition bandwidth, so the same order can be achieved. The proposed method can extract the signal in baseband instead of merging the adjacent sub-channels. So the design of the prototype filter is released from the requirement of perfect reconstruction. This can reduce the hardware complexity significantly. Meanwhile, the synthesis section consists of digital down-convert and filtering operation, which is simpler than the structure of merged sub-channels and has less transmission distortion.

The remainder of this paper is organized as follows. In Section 2, the architecture of the proposed digital channelized receiver is presented. Some simulation results are shown in Section 3, followed by the conclusions in Section 4.

2. Proposed architecture of dynamic channelized receiver

The most widely used architecture for the dynamic channelized receiver consists of an analysis section and synthesis section [2-7]. Many previous works focus on the design of analysis section and achieve remarkable performance. But few of them detail the channel detection and arbitration, and the design of synthesis filters.

This paper proposed a simple architecture for the dynamic channelized receiver. The overall structure of the channelized receiver is shown in Fig.1. $H_o(z^k)$ represents the $K$-time interpolated half filter, $z^{-K(N_o-1)/2}$ represents the input signal is delayed $-K(N_o-1)/2$ sampling clocks. Here, $2K$ is the number of sub-channels and $N_o$ is the order of $H_o(z^k)$. $H_{pd}(z)$ is a low pass filter. $T_j$ is the arbitration period, that is, the center frequencies and bandwidths of the input signal are output every $T_j$ sampling clocks. Register block 1 and register block 2 are one-to-one, registering the center frequencies and bandwidths respectively. $\omega_{d,q}$ is the radian frequency, $G_{d}(z) \ (q=0,1,2,\ldots,Q-1, \ Q \ is \ the \ number \ of \ signals)$ is the synthesis filter and $M_q$ is the decimation rate. The detailed analysis process of these three parts is as follows. The outputs of the dynamic channelized can be described as

$$y_q(m) = (s(n)e^{-j\omega_{d,q}}) \ast g(n)\big|_{n=M_qm}, \ q = 0, 1, \ldots, Q-1$$ (1)
2.1 Analysis section

The purpose of the analysis section is to divide the wideband into multiple sub-channels. The number of sub-channel is $2K$ which can be determined as [5]

$$2K = 2^{\log_2 \left( \frac{G_{\text{min}}}{G_{\text{max}}} \right)}$$

(2)

where, $G_{\text{min}}$ is the minimum guard-band of all sub-channel signals which is given beforehand.

Designing a half band filter $H_a(z)$, $H_c(z)$ is the complementary filter of $H_a(z)$. $H_a(z^K)$ and $H_c(z^K)$ are the $K$-time interpolation of $H_a(z)$ and $H_c(z)$ respectively. $H_a(z^K)$ and $H_c(z^K)$ are period filters, and $H_c(z^K)$ can be described by $H_a(z^K)$ as

$$H_c(z^K) = z^{-K(N-1)/2} - H_a(z^K)$$

(3)

The frequency responses of $H_a(z^K)$, $H_c(z^K)$ and $H_{\text{ana}}(z)$ are shown in Fig.2. Where the solid line represents $H_a(z^K)$, the dashed line represents $H_c(z^K)$, and the dot dashed line represents $H_{\text{ana}}(z)$. $\omega_{aP}$ and $\omega_{aS}$ are the passband cutoff frequency and stopband initial frequency of $H_a(z)$ respectively, $\omega_{aP} + \omega_{aS} = \pi$. 

Figure 1. Proposed architecture of channelized receiver
The whole bandwidth is decomposed to even and odd sub-channels as Fig.2. Then $\omega_{a,k}$ and $\omega_{c,k}$ can be described as

$$\omega_{a,k} = \frac{2\pi k}{K}, \omega_{c,k} = \frac{2\pi}{K}(k + \frac{1}{2}), k = 0, 1, 2, \ldots, K - 1$$

Based on the above analysis, only the half filter $H_a(z)$ and the masking filter $H_{ma}(z)$ need to be designed. Assume the transition bandwidth of sub-channel is $\omega_{trans}$, that is, $1/K$ times of that of $H_a(z)$. Then $\omega_{ap}$ can be written as

$$\omega_{ap} = \frac{\pi - K\omega_{trans}}{2}$$

$H_{ma}(z)$ is a LPF which used to masking the unwanted band. From fig.2, the passband cutoff frequency and stopband initial frequency of $H_{ma}(z)$ can be written as

$$\omega_{map} = \frac{\omega_{ap}}{2} = \frac{1}{2}\left(\frac{\pi}{K} - \omega_{trans}\right)$$

$$\omega_{mas} = \frac{2\pi}{K} - \frac{\omega_{ap}}{2} = \frac{3\pi}{2} - \frac{\omega_{trans}}{2}$$

After uniform filter bank processing, the analysis section can output $2K$ signals. The even sub-channels and odd sub-channels are processed through two uniform FBs respectively. This process is the same as DFTFB, so the efficient polyphase FB architecture can be employed. The efficient architecture of even sub-channels is shown in Fig.3, where $F$ is the decimated factor. The architecture is maximally decimated architecture when $F=1$, non-maximally decimated architecture when $F$ is other values. The detailed deduction is omitted. The efficient architecture of odd sub-channels is similar to that of even sub-channels, so it is not presented in this paper.

![Figure 3. Efficient architecture of even sub-channels](image-url)
It should be noted that the time-multiplexed technique is particularly suitable for polyphase filter banks because the sub-channels have the same structure and work at low clock rates. Assume the multiplexed rate is $v$, then even and odd sub-channels can be divided into $\lceil K / v \rceil$ groups respectively. $\lceil \cdot \rceil$ represents the round up. Each group operates at the clock rate of $vf/K$. The value of $v$ is chosen according to the maximum core clock rate of the hardware platform. Hence, the hardware complexity can be reduced by a factor of $v$.

The idea for designing the architecture of the analysis section is motivated by the interpolated frequency impulse filter (IFIR), which is equal to replace the prototype filter with a period model filter and a masking filter [8]. Hence, the narrow transition band can be achieved with lower hard complexity than the tradition FIR filter.

2.2 Channel detection and arbitration
Suppose $Q$ signals are input to the receiver simultaneously. Let $C^l_q$ and $C^u_q$ denotes the lower and upper number of the included sub-channels of the $q$th sub-channel signal. The values of $C^l_q$ and $C^u_q$ can be confirmed through channel detection and arbitration, and output once every $T_J$ clock cycle. $T_J$ is set to the maximum repetition period of the pulse signal or the maximum modulated period of the continuous wave (CW) signal. Then the center frequency and bandwidth of $q$th sub-channel signal can be written as

$$\omega_{d,q} = \frac{C^l_q + C^u_q}{2}, q = 0, 1, 2, \cdots, Q - 1$$

(7)

$$B_{d,q} = \frac{C^u_q - C^l_q}{2}, q = 0, 1, 2, \cdots, Q - 1$$

(8)

The value of $\omega_{d,q}$ can be used to converted the $q$th sub-channel signal into baseband. The value of $B_{d,q}$ can be used as the address to select the proper filter coefficients for synthesis filter. For convenience, $B_{d,q}$ is written as $B_{d,q} = (C^u_q - C^l_q)$.

2.3 Synthesis section
In the synthesis section, the input signals are converted to baseband based on the value of center frequencies computed above at first. Then a unified LPF is designed to extract the signal with different bandwidths. The LPF with different bandwidths is designed with the same transition band. Since the order of the filter is proportional to the transition bandwidth, the LPFs can be designed with the same order. The coefficients of LPF for the $q$th sub-channel signal is selected based on the value of $B_{d,q}$. Finally, the data rates of extracted signals are reduced by $M_q$ times. The value of $M_q$ can be determined by $B_{d,q}$ as

$$M_q = \frac{2K}{B_{d,q}}$$

(9)

For convenience, $M_q$ is written as $M_q = 2^{\lceil \log_2(2K/B_{d,q}) \rceil}$

From the above analysis, it can be seen that the design of unified LPF is the key to the synthesis section. Assume $\theta_{q,p}$, $\theta_q$, and $\theta_{q,\text{trans}}$ are the passband cutoff frequency, stopband initial frequency and transition bandwidth of $G_q(z)$. Set the all values of $\theta_{q,\text{trans}}$ to $G_{\text{min}}$ to suppress the out-of-band signal as much as possible. Supposing the maximum input signal bandwidth is $I$ times of the sub-channel bandwidth. Then $I$ LPFs with different bandwidths and the same transition bandwidth need to be designed. For each sub-channel of the synthesis section, the LPF can be selected from the $I$ unified filters based on the value of $B_{d,q}$.
Direct design for such filters will consume many resources. IFIR is an efficient method to reduce complexity. The structure of the synthesis section based on IFIR is shown in Fig. 4. Where, \( G_{q,p}(z^L) \) is the model period filter and \( G_{q,mu}(z) \) is the masking filter. \( G_{q,p}(z^L) \) is \( L \)-time interpolation of \( G_{q,p}(z) \). For convenience, \( G_d(z) \) and \( G_{mu}(z) \) are used instead of \( G_{q,p}(z) \) and \( G_{q,mu}(z) \), \( \theta_p \) and \( \theta_s \) are used instead of \( \theta_{q,p} \) and \( \theta_{q,s} \). The design for \( G_d(z) \) and \( G_{mu}(z) \) is similar to that of \( H_d(z) \) and \( H_{mu}(z) \) which can be described as

\[
\begin{align*}
\theta_{ap} &= L\theta_p, \theta_{as} = L\theta_s \\
\theta_{map} &= \theta_p, \theta_{mas} = 2\pi / L - \theta_s
\end{align*}
\]

where, \( L \) is the interpolated factor. The transition bandwidths of \( G_d(z) \) and \( G_{mu}(z) \) are

\[
\begin{align*}
\theta_{atran} &= L\theta_{trans} \\
\theta_{matran} &= 2\pi / L - \theta_{trans} - 2\theta_p
\end{align*}
\]

Since the IFIR with different bandwidth should be designed uniformly with the same order, \( \theta_{atran} \) and \( \theta_{matran} \) must be constant. Hence, \( L \) must be a constant. And \( \theta_{matran} \) is set to the minimum value, corresponding to the maximum of \( \theta_p \) to adapt to the different band width situations. Supposing \( \theta_{max} \) is the maximum of \( \theta_p \), \( \theta_{max} = \pi l / K \). The value of the optimum value of \( L \) which provides minimum hardware complexity is given by [8]

\[
L_{opt} = \left[ \frac{2\pi}{\theta_{trans} + 2\theta_{max} + 2\pi \theta_{matran}} \right]
\]

where, \([\cdot]\) stands for the rounding integer.

![Figure 4. Architecture of synthesis section based on IFIR](image)

### 3. System simulation and analysis

#### 3.1 Dynamic channelized simulations

In this section, numerical simulations are conducted to confirm the validity of the proposed dynamic channelized receiver. Before the simulation, it is assumed that the input signal satisfies the following limitations: the minimum guard-band \( \Omega_{min} = \pi / 32 \), the maximum bandwidth of input signal is \( \pi / 8 \), the maximum number of signal \( Qs \) and 4, the transition bandwidth of sub-channel \( \omega_{trans} = \pi / 800 \). Hence, the number of sub-channels \( 2K \) is 32, the optimum interpolated factor \( L_{opt} = 5 \). Assume the bandwidth of the receiver is 800MHz, then the sub-channel bandwidth is 25MHz. Based on the Eq. (5) and Eq. (6), the half-band filter \( H_p(z) \) has a passband cutoff frequency of 384MHz, a stopband initial frequency of 416MHz. The masking filter \( H_{mu}(z) \) has a passband cutoff frequency of 12MHz, a stopband initial frequency...
frequency of 37MHz. The unified model filter $G_a(z)$ has passband cutoff frequencies of 62.5MHz, 125MHz, 187.5MHz and 250MHz, stopband initial frequency of 187.5MHz, 250MHz, 312.5MHz and 375MHz. The unified masking filter $G_{ma}(z)$ has passband cutoff frequencies of 12.5MHz, 25MHz, 37.5MHz and 50MHz, stopband initial frequency of 187.5MHz, 250MHz, 312.5MHz and 375MHz, stopband initial frequency of 182.5MHz, 195MHz, 207.5MHz and 220MHz.

The input complex signals are as follow.

(1) Sine signal 1: carrier frequency is 37MHz and pulse repetition interval (PRI) is 80us;
(2) LFM pulse signal 2: carrier frequency is 68MHz, bandwidth is 10MHz and PRI is 100us;
(3) LFM pulse signal 3: carrier frequency is 345MHz, bandwidth is 60MHz and PRI is 110us;
(4) LFMCW signal 4: carrier frequency is 550MHz, bandwidth is 100MHz and modulated period is 120us.

According to the channel division shown in Fig.2, it is known that signal 1 and signal 2 occupy 1 sub-channel which is 1st and 3rd sub-channel, signal 3 occupies 3 sub-channels: from 13th to 15th sub-channels, signal 4 occupies 5 channels: from 20th to 24th sub-channels. To validate the ability to process the dynamic signals, the time-frequency diagram is used to represent the information of the input signal, as shown in Fig.5. During the interception process, the number of input signals is changed from 3 to 4. The output of the channel detection and arbitration section is shown in Fig.6. The results show that the channel detection and arbitration section can outputs the lower and upper number of the included sub-channels of sub-channel signals accurately.

For convenience, the decimation factors for the four signals are set to 32, 32, 16 and 8. The amplitude-frequency responses of the extracted signals are shown in Fig.7. From the simulation results, it can be seen that the proposed dynamic channelized receiver can extract multiple signals with unknown parameters.
3.2 Computation complexity analysis and comparison

In this paper, the complexity of the channelized receiver architectures in [3] and [5] are compared with the proposed channelized receiver architecture in the same situation. The filters of the three architectures have the same passband bandwidth, stopband bandwidth, passband ripple, and stopband ripple. Since multipliers resources are limited for hardware platforms, the number of multipliers used is regarded as the measure of computational complexity. The computational complexity for implementing the entire architecture is set to be $C$ which consists of the analysis section and synthesis section.

Through analysis of fig.1, it can be seen that analysis section needs $(N_{a}+1)/2+4N_{ma}/v$ complex multiplications and $K\times\log K$ complex multiplications. Here $N_{ma}$ is the order of $H_{ma}(z)$. Considering half of the coefficients of half filter $H_{a}(z)$ is zero, the multipliers needed in analysis section is: $(N_{a}+1)/2+4N_{ma}/v+2K\times\log K$. Similarly, the synthesis section needs $2Q\times N_{fir}$ multipliers. Here $N_{fir}$ is the total order of $G_{a}(z)$ and $G_{ma}(z)$. Hence, the computational complexity of the proposed dynamic channelized receiver is

$$C_1 = (N_{a}+1)/2 + 4N_{ma}/v + 2K\times\log K + 2Q\times N_{fir}$$

(13)

In [3], the analysis section needs $(N_{a}+1)/2+4N_{ma}+2K\times\log K$ multipliers. The synthesis section needs $2\times\sum_{q=0}^{Q-1}(N_{q} + L_{q}/2)\times\log L_{q}$ multipliers. Here, $N_{q}$ is the order of filter for $q$th sub-channel signal. $L_{q}$ is the number of sub-channel which $q$th sub-channel signal occupies. Hence, the computational complexity of the dynamic channelized receiver proposed in [3] is

$$C_2 = (N_{a}+1)/2 + 4N_{ma} + 2K\times\log K + \sum_{q=0}^{Q-1}(2N_{q} + L_{q} \times \log L_{q})$$

(14)
In [5], the analysis section directly filters the input signal with a CMFB which need $N_d \times 2K$ multipliers. Here $N_d$ is the order of prototype filter. The synthesis section is similar to that of [3] which requires the same number of multipliers. Hence, the computational complexity of the dynamic channelized receiver proposed in [5] is

$$C_3 = N_d \times 2K + \sum_{q=0}^{Q-1} (2N_q + L_q \times \log_2 L_q)$$  \hspace{1cm} (15)$$

Supposing the passband ripple and stopband ripple are all 0.001dB. Then we can get $N_a=167$, $N_{ma}=210$, $N_{ga}=42$, $N_{gma}=31$, $N_{f0}=N_{ga}+N_{gma}=73$. Here, $N_{ga}$ and $N_{gma}$ are the order of $G_a(z)$ and $G_{ma}(z)$ respectively. Signal 1 and signal 2 can be output from the synthesis section in [3], so the values of $N_0$, $N_1$, $L_0$ and $L_1$ are all zero. The filter order of signal 3 is $N_2=653$, and the interpolation factor $L_2=4$. Similarly, the value of $N_3$ and $L_3$ are 1304 and 8 respectively. The order of prototype filter in [5] is $N_d=5211$.

The complexities of the three methods are compared as shown in Table 1 below.

| Design Method     | The number of multipliers needed | saving       |
|-------------------|----------------------------------|--------------|
| Ref [3]           | 167336                           | 99.3%        |
| Ref [5]           | 5222                            | 79.0%        |
| Proposed method $v=8$ | 1096                            |              |

4. Conclusion

This paper proposed a new dynamic channelized architecture without merging the adjacent sub-channels. The narrow transition band is achieved by employing IFIR with low hardware complexity. A time-multiplexed technique is used in the analysis section which reduces the complexity further. The architecture is very simple. The complexity is very low compared with others. So it is easy to implement. Simulation results show that the dynamic channelized structure is effective.

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