

Designing simultaneous multichannel receivers based on fast filter bank

Hao Jinguang Pei Wenjiang Wang Kai Xia Yili

(School of Information Science and Engineering, Southeast University, Nanjing 210096, China)

Abstract: A scheme to design a simultaneous multichannel receiver is proposed to process multichannel signals in parallel, which is achieved by exploiting the attractive characteristics of a fast filter bank (FFB), such as cascaded structure, high frequency selectivity and low computational complexity. Based on the minimization of the objective function, quantified in terms of the total number of multiplications required, subject to prescribed allowable ripples in the passband and stopband, the impulse response coefficients of the prototype filter in each stage are obtained to meet the requirements of the overall specifications for each channel at the receiver side. Simulations and experimental results on the frequency modulation (FM) broadcast multichannel signal receiving system with the FM range from 88 to 108 MHz, built upon the proposed FFB structure, are performed to verify its performance. Those results indicate that the proposed scheme is efficient in FM audio indexing applications and has a lower computational complexity, which is approximately 66.4% of the weighted overlap and add (WOLA) filter banks based solution.

Key words: fast filter bank (FFB); low complexity; sparse coefficients; modular instrument

doi: 10.3969/j.issn.1003-7985.2015.04.005

Software defined radio (SDR) is a promising technique for wireless communications, which provides a general purpose hardware to support different air interface standards^[1]. The SDR receiver uses a channelized receiver to extract multiple narrowband channels from a wide-band digital signal. In past decades, different methods or techniques have been developed to implement channelized receivers in the literature^[2-5]. However, in some specific applications, it is required that the channelized receivers

have to capture and process signals in parallel coming from a large number of channels, such as frequency modulation (FM) broadcast receivers used in audio indexing applications^[6], and this issue has not been addressed in the existing literature. To this end, a structure to build up FM broadcast receivers based on weighted overlap and add (WOLA) filter banks is proposed in Ref. [6]. Unfortunately, the number of multiplications required by the WOLA filter banks-based solution is large, and as expected this is more evident when the number of channels increases. The computational burden limits its use in practical applications.

To alleviate the above problem, a low complexity scheme to design channelized receivers for simultaneous multichannel signal processing is proposed based on a fast filter bank (FFB) to exploit its attractive properties, such as cascaded structure, high frequency selectivity, and comparable computational complexity as FFT/IFFT. In order to obtain the optimum coefficients of the prototype filter within FBB to achieve acceptable ripple magnitude tolerance in each subfilter, also with moderate computational complexity, we utilize an optimization technique to minimize the total number of multiplications required by the proposed structure. This is subject to predefined allowable ripples in the passband and stopband.

Simulations in Matlab are conducted to verify the suitability of the proposed FFB based simultaneous multichannel receiver for FM broadcasting applications. Moreover, a hardware platform based on modular instruments is set up to implement the proposed scheme with a field-programmable gate array (FPGA) module and demodulate multichannel monophonic FM radio broadcast signals coming from different channels. Simulations and experimental results illustrate that the proposed scheme is applicable in the context of simultaneous multichannel signal processing with less multiplications required by per channel than that of conventional methods, such as the WOLA filter banks-based solution.

1 Review of FFB

FFB was first introduced as a generalized form of the sliding fast Fourier transform (FFT) filter bank^[7]. This class of filter bank has attractive characteristics, such as cascaded structure, high frequency selectivity, and comparable complexity of FFT due to its sparse coefficients. FFB has been widely used in music signal processing, spectrum

Received 2015-03-26.

Biographies: Hao Jinguang (1977—), male, graduate; Pei Wenjiang (corresponding author), male, doctor, professor, wjpei@seu.edu.cn.

Foundation items: The National Natural Science Foundation of China (No. 61201173, 61271058, 61401094), the Specialized Research Fund for the Doctoral Program of Higher Education of China (No. 20110092110008), the Natural Science Foundation of Jiangsu Province (No. SBK201140040, BK2011060, BK20140645), the Scientific Research Foundation for the Returned Overseas Chinese Scholars, State Education Ministry of China.

Citation: Hao Jinguang, Pei Wenjiang, Wang Kai, et al. Designing simultaneous multichannel receivers based on fast filter bank[J]. Journal of Southeast University (English Edition), 2015, 31(4): 457 – 461. [doi: 10.3969/j.issn.1003-7985.2015.04.005]

sensing and multi-standard wireless receivers^[8-11].

For an L -stage FFB, we can realize $N = 2^L$ data channels. The cascaded structure of an FFB to process multichannel data is shown in part A of Fig. 1 and the structure of the subfilters within FFB is given in Fig. 2. Note that for notation simplicity in Fig. 1, indices (k, m) represent the transfer function of the m -th subfilter in the k -th stage, that is, $H^{k,m}(z)$, where $k = 0, 1, \dots, L-1$ and $m =$

$0, 1, \dots, 2^k - 1$. In general, subfilters within FFB can be modelled as two-path systems, where the upper path is termed as the original component with the transfer function $H_a^{k,m}(z)$ and the lower path is considered as the complementary component with the transfer function $H_c^{k,m}(z)$. $H_a^{k,m}(z)$ and $H_c^{k,m}(z)$ satisfy the relationship,

$$H_a^{k,m}(z) + H_c^{k,m}(z) = 2 \quad (1)$$

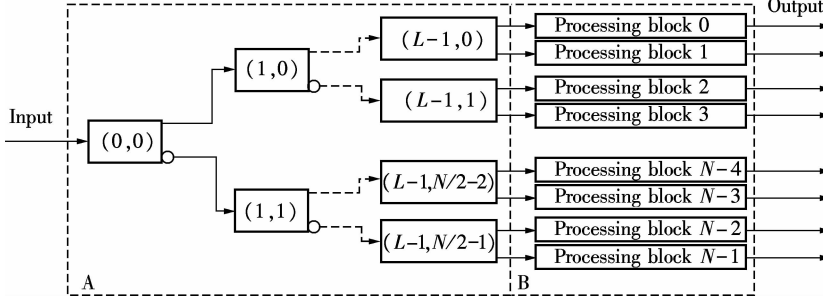


Fig. 1 The proposed scheme to process multichannel data based on FFB

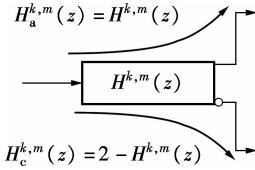


Fig. 2 The structure of the subfilter

When the transfer function $H_a(z)$ of the prototype filter for the k -th stage is obtained in hand by using the halfband filter design method^[7, 12-13], the transfer function $H^{k,m}(z)$ can be determined by replacing z in $H_a(z)$ with $W_N^{\tilde{m}} z^{L-1-k}$, where W_N stands for $e^{-j2\pi/N}$ and \tilde{m} is the bit reversed version of m in $L-1$ bits. In this way, the transfer function of a specific channel n_0 , that is, $H_{n_0}(z)$, can be expressed as

$$H_{n_0}(z) = \frac{1}{N} \prod_{k=0}^{L-1} H_{n_0}^k(z) \quad (2)$$

where $H_{n_0}^k(z)$ represents the transfer function of the corresponding subfilter in the k -th stage within the channel n_0 ; $1/N$ is the normalized factor.

2 Proposed Scheme for Simultaneous Multi-Channel Receivers

The proposed FFB based simultaneous multichannel receiver for FM broadcasting applications is shown in Fig. 1. The structure consists of two parts: part A is the cascaded channelized receiver built up based on FFB, which is used to extract narrowband signals from input and decompose the input signal into N -channel parallel filtered components. In the following, the filtered multiple components from part A are transferred into different processing blocks working in parallel in part B. In this manner, all filtered components in different frequency bands can be processed simultaneously. Note that the normalized factor $1/N$ is omitted in Fig. 1.

To achieve predefined ripple magnitude tolerance in

both the passband and stopband of each subfilter for the overall FM multichannel receiving system, we utilize an optimization technique to obtain optimal coefficients for the prototype filter in each stage. Since FFB is an N -channel uniformly spaced filter bank and the frequency response of a specific filter channel is the modulated version of that of the 0th one, it can be deduced that the magnitude frequency response of the 0th channel is identical to that of other channels. Therefore, the overall specifications of FFB for the whole spectrum from 0 to π now boil down to the design of the 0th channel, in which there are four important specification factors, including the passband edge frequency ω_p , the allowable passband ripple δ_p , the stopband edge frequency ω_s , and the allowable stopband ripple δ_s .

We further assume that the prototype filter for the k -th stage is an odd-length noncausal filter with a symmetrical structure, and design it according to the halfband filter method, so that the prototype filter has the property that its every other coefficient value is zero, giving a smaller number of nontrivial coefficients. In addition, the halfband filter method can be achieved with less specification factors, that is, the passband edge frequency ω_p^h and the allowable ripple δ_p^h in the passband only, since its stopband edge frequency ω_s^h and its stopband ripple can be determined as $\omega_s^h = \pi - \omega_p^h$ and $\delta_s^h = \delta_p^h$, respectively.

The optimal design of the prototype filter for each stage within FFB by using the halfband filter method can be achieved as follows:

Step 1 Initialization of specifications of the prototype filter for each stage with the halfband method to satisfy the constraints required by overall specifications, that is,

$$\omega_{p,k} = \max\left(\frac{N}{2}\omega_p, \pi - \frac{N}{2}\omega_s\right), \quad \delta_{p,k} = \min\left(\frac{\delta_p}{L}, \frac{\delta_s}{L}\right)$$

where $\max(\cdot)$ and $\min(\cdot)$ are the maximum and min-

imum operators, respectively; $\omega_{p,k}$ and $\delta_{p,k}$ denote the passband edge frequency and the allowable passband ripple in the passband of the prototype filter designed for the k -th stage of FFB, respectively.

In general, the length of finite impulse response (FIR) filter, Q , can be calculated as^[14]

$$Q = \frac{-20\pi\log_{10}(\delta_{p,k}\delta_{s,k}) - 13}{14.6(\omega_{s,k} - \omega_{p,k})} + 1 \quad (3)$$

Within a halfband filter, due to the fact that $\delta_{s,k} = \delta_{p,k}$ and $\omega_{s,k} = \pi - \omega_{p,k}$, the evaluated length in Eq. (3) can be rewritten as

$$Q = \frac{-40\pi\log_{10}(\delta_{p,k}) - 13}{14.6(\pi - 2\omega_{p,k})} + 1 \quad (4)$$

From Eq. (4), we can observe that the minimization of the filter length Q and hence the total number of multiplications is achieved by making the passband edge frequency $\omega_{p,k}$ as small as possible and the ripple in the passband $\delta_{p,k}$ as large as possible for the prototype filter in each stage.

For the notation simplicity, we use a vector $\boldsymbol{\mu} = \{\omega_{p,0}, \delta_{p,0}, \dots, \omega_{p,L-1}, \delta_{p,L-1}\}$ to represent the specification factors of the prototype filters in all stages and a constant vector $\boldsymbol{\nu} = \{\alpha_0, \beta_0, \dots, \alpha_{L-1}, \beta_{L-1}\}$ to represent the specification adjustments, where $L = \log_2(N)$ and $\beta_k \geq 1$. The following steps are used to decrease the total number of required multiplications.

Step 2 The coefficients $h_a^k(n)$ and the length of the prototype filter, $2M_k + 1$, in the k -th stage can be computed with

$$[h_a^k(n), 2M_k + 1] = \Theta(\omega_{p,k}, \delta_{p,k}) \quad (5)$$

where $\omega_{p,k}$ and $\delta_{p,k}$ are from vector $\boldsymbol{\mu}$; $n = -M_k, \dots, 0, \dots, M_k$ and $\Theta(\cdot)$ denotes the halfband filter method. Subsequently, the transfer function $H_a^k(z)$ and its frequency response $H_a^k(e^{j\omega})$ can be expressed with coefficients $h_a^k(n)$ as

$$\begin{aligned} H_a^k(z) &= \sum_{n=-M_k}^{M_k} h_a^k(n) z^{-n} \\ H_a^k(e^{j\omega}) &= \sum_{n=-M_k}^{M_k} h_a^k(n) e^{-jn\omega} \end{aligned} \quad (6)$$

Step 3 The transfer function of the 0th channel $H_0(z)$ and its frequency response $H_0(e^{j\omega})$ can be determined with

$$\begin{aligned} H_0(z) &= \frac{1}{N} H_a^0(z^{2^{L-1}}) H_a^1(z^{2^{L-2}}) \cdots H_a^{L-1}(z) \\ H_0(e^{j\omega}) &= \frac{1}{N} H_a^0(e^{2^{L-1}j\omega}) H_a^1(e^{2^{L-2}j\omega}) \cdots H_a^{L-1}(e^{j\omega}) \end{aligned} \quad (7)$$

and $H_0(e^{j\omega})$ should obey the constraints as

$$\begin{aligned} 1 - \delta_p &\leq |H_0(e^{j\omega})| \leq 1 + \delta_p & 0 \leq \omega \leq \omega_p \\ |H_0(e^{j\omega})| &\leq \delta_s & \omega_s \leq \omega \leq \pi \end{aligned} \quad (8)$$

Step 4 When the constraints are satisfied, the ele-

ment in $\boldsymbol{\mu}$, $\mu(l)$, is updated with $\mu(l)\nu(l)$, where $\nu(l)$ is the corresponding element in vector $\boldsymbol{\nu}$, and the procedure repeats from Step 2 to Step 4. Otherwise, the index l is updated with $l - 1$, and the process repeats from Step 2 to Step 4 until l is equal to -1 . In this way, elements in $\boldsymbol{\mu}$ are optimal for the design of the prototype filter at each stage, and the coefficients $h_a^k(n)$ and the length of the prototype filter, $2M_k + 1$, at the k -th stage can be determined accordingly.

In brief, the above iterative procedure can provide a solution to the following problem:

$$\begin{aligned} \min \quad & \sum_{k=0}^{L-1} 2^k \left\lceil \frac{M_k + 1}{2} \right\rceil \\ \text{s. t.} \quad & 1 - \delta_p \leq |H_0(e^{j\omega})| \leq 1 + \delta_p \quad 0 \leq \omega \leq \omega_p \\ & |H_0(e^{j\omega})| \leq \delta_s \quad \omega_s \leq \omega \leq \pi \end{aligned} \quad (9)$$

From Refs. [7, 13], it is known that the required number of multiplications of a halfband filter is about one-quarter of its length. Due to this property, the required number of multiplications for the subfilter in the k -th stage is $\lfloor (M_k + 1)/2 \rfloor$, where $\lfloor x \rfloor$ represents the maximum integer no more than x . Therefore, the total number of multiplications of the proposed scheme M in all stages is

$$M = \sum_{k=0}^{L-1} 2^k \left\lceil \frac{M_k + 1}{2} \right\rceil \quad (10)$$

3 Simulations and Experiments

We conducted simulations and built up experiments to verify the performance of the proposed FFB based simultaneous multichannel receiver in FM broadcasting applications. The hardware platform based on modular instruments for the experiment is shown in Fig. 3, which consists of an embedded controller, a vector signal analyzer (VSA), an FPGA module, antenna and a sound card. In this experiment, monophonic FM radio broadcast signals are chosen as the input. The FM signals were first down-converted into baseband by the VSA, and then were processed by the proposed FFB based simultaneous multichannel receiver implemented in FPGA module, and were finally transferred into the embedded controller for the verification of the audio signals.

In order to cover the full FM frequency range from 88 to 108 MHz, we designed a channelized receiver with the number of channels $N = 128$ based on a FFB with $L = \log_2(N) = 7$ stages. The center carrier frequency of VSA and its sampling frequency were set to be 98.1 and 25.6 MHz, respectively. In this experiment, we aimed to demodulate and separate the audio signals coming from two different FM broadcast channels simultaneously, that is, a news channel and a music channel in Jiangsu province, whose center carrier frequency was 93.7 and 97.5 MHz, respectively.

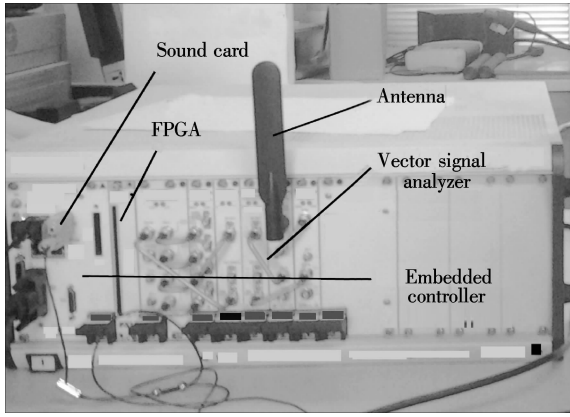


Fig. 3 The hardware platform based on modular instruments

Tab. 1 The coefficients of prototype filters in each stage

n	$h_n^0(n)$	$h_n^1(n)$	$h_n^2(n)$	$h_n^3(n)$	$h_n^4(n)$	$h_n^5(n)$	$h_n^6(n)$
0	1.000 0	1.000 0	1.000 0	1.000 0	1.000 0	1.000 0	1.000 0
± 1	0.626 7	0.615 1	0.569 1	0.571 7	0.500 5	0.500 8	0.501 0
± 3	-0.183 9	-0.154 4	-0.069 6	-0.072 7			
± 5	0.084 7	0.049 9					
± 7	-0.039 6	-0.011 5					
± 9	0.016 3						
± 11	-0.005 2						

receiver are depicted in Fig. 4, from which it is clear that the allowable ripples in the passband and stopband, the passband edge frequency and the stopband edge frequency all met the predefined constraints.

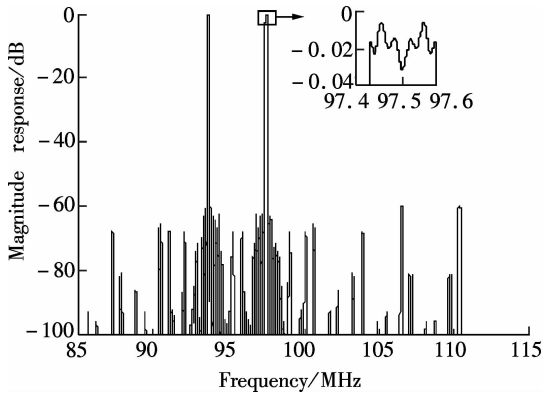


Fig. 4 Magnitude frequency responses of desired channels

In the next stage, the proposed FFB based scheme was implemented in the FPGA module of modular instruments. The hardware structure of the subfilters within FFB is depicted in Fig. 5, where p denotes the interpolation factor at each stage. Note that in this experiment, the subfilters have been used as one-input and two-output systems, where one output is the original component and the other is the complementary complement obtained by subtracting the original component from twice the input signal, as indicated by the relationship given in Eq. (1). The FM demodulation blocks used for processing FFB output signals in parallel were achieved based on the Cordic structure^[6].

For the prototype filter used in the 0th channel of the FFB, the allowable ripples in the passband and stopband, δ_p and δ_s , were set to be 0.002 3 (0.02 dB) and 0.001 (-60 dB), respectively. The passband edge frequency ω_p was $2 \cdot 0.4\pi/N$ (80 kHz) and the stopband edge frequency ω_s was $2 \cdot 0.7\pi/N$ (140 kHz). The optimal coefficients of the prototype filter in each stage of FFB, which met the above requirements, can be obtained by the optimization method, in which the nonzero ones are given in Tab. 1.

Within the channelized receiver, the data channels of interest, containing information coming from 93.7 and 97.5 MHz, were No. 234 and No. 253, respectively. The magnitude frequency responses of the proposed FFB based

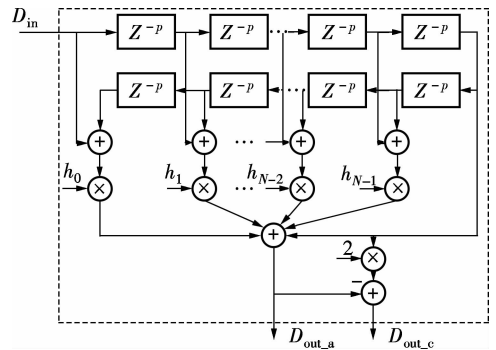


Fig. 5 The hardware structure of subfilters within FFB

The simultaneously demodulated waveforms of FM broadcasting signals of interest are illustrated in Fig. 6, and both the speech signal at 93.7 MHz and the music signal at 97.5 MHz were clear enough for the audience, indicating the suitability of the proposed FFB based multi-channel FM receiver for audio indexing application.

$$\bar{M} = 4 \cdot (6 \cdot 2^0 + 4 \cdot 2^1 + 2 \cdot 2^2 + 2 \cdot 2^3 + 1 \cdot 2^4 + 1 \cdot 2^5 + 1 \cdot 2^6) / 128 = 4.687 5 \quad (11)$$

We now consider the computational complexity, evaluated in terms of the number of real-valued multiplications of the proposed scheme. For the coefficients listed in Tab. 1, the average number of real multiplications per channel, \bar{M} , was calculated as 4.687 5 according to Eq. (11). When compared, the average number of real multiplications per channel used in the WOLA filter banks based solution is 7.062 5^[5-6], indicating that the complexity of the proposed scheme is approximately 66.4% of the WOLA filter banks based solution.

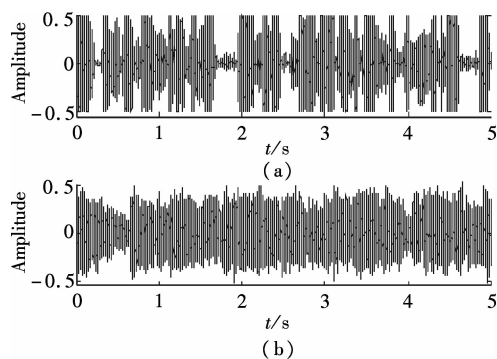


Fig. 6 The simultaneously received waveforms of two FM signals. (a) Waveform at 93.7 MHz; (b) Waveform at 97.5 MHz

4 Conclusion

A novel scheme to design low-complexity channelized receivers for simultaneous multichannel signal processing is proposed. It is achieved based on FFB by exploiting its attractive properties, such as cascaded structure, high frequency selectivity, and low computational complexity due to its sparse representation. In order to design the prototype filters used in each stage within the FFB, an optimization technique is utilized to obtain optimal coefficients for the minimization of the number of multiplications required, subject to predefined allowable ripples in the pass-band and stopband, respectively. Simulations on the frequency modulation (FM) broadcast multichannel receiving system with the FM range from 88 to 108 MHz, built upon the proposed FFB structure, are performed to verify its performance. Results show that the proposed scheme is able to simultaneously demodulate FM audio signals with the predefined ripple constraints, and it has a lower computational complexity compared with the weighted overlap and add (WOLA) filter banks based solution.

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基于快速滤波器组多通道同步接收机的设计

郝金光 裴文江 王 开 夏亦犁

(东南大学信息科学与工程学院, 南京 210096)

摘要:根据快速滤波器组所具有的级联结构、良好频率选择性以及低复杂度的特点,提出了一种多通道同步接收机的设计方法来并行处理多路接收信号。该方法将最小化系统所需乘法器的总数量作为目标函数,以通带和阻带所允许的纹波作为约束条件,设计各级原型滤波器的脉冲响应系数,以满足接收机各通道全局设计指标的要求。以工作频率为 88~108 MHz 的调频广播信号并行解调系统作为仿真和实验平台对该方法进行验证。结果表明,所提出的设计方法能够满足多通道同步信号处理的要求,且每个通道所需乘法器的数量大约为加权重叠(WOLA)滤波器组方法所需乘法器数量的 66.4%,具有较低的实现复杂度。

关键词:快速滤波器组;低复杂度;稀疏系数;模块化仪器

中图分类号: TN911.72