Multiphase Parallel Demodulation for Remote Sensing Satellite Data Transmission—Filter Bank Based on WOLA Structure

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Abstract—With the rapid increase of data generated by remote sensing satellites, the demand of real-time demodulation of ultrahigh speed remote sensing data by digital receivers has become a research hotspot and difficulty. Aiming at the international research gap of parallel demodulation of remote sensing satellite data using multichannel channelization technology, a parallel matched filtering architecture of filter banks based on the weighted overlap-add (WOLA) structure is designed. This structure has the following advantages. First, the decimation factor is flexible and adjustable when the signal is channelized, which is not restricted by the number of channels. Second, the signal is divided into several sub-bands according to the frequency spectrum, which is beneficial to the parallel analysis and processing of the signal. This article completes the parallel matched filtering processing and provides ideas for the subsequent processing operation. Third, the structural channelization decomposition has strong expandability and can be flexibly applied to signal transmission of different systems. Aiming at the characteristics of high speed and wide bandwidth of remote sensing data transmission, this article designs the channelization method of filter bank based on WOLA structure and deduces and realizes the sub-band matching filtering operation. Taking QPSK modulation as an example, the simulation results show that parallel matched filtering has almost no loss in bit error performance. The research results of this article have great reference value for designing multiphase parallel receivers of high-speed remote sensing data in the future.

Index Terms—Nonmaximum decimation filter bank, parallel demodulation, perfect reconstruction, remote sensing data transmission.

I. INTRODUCTION

W ith the vigorous development of Earth observation technology, the amount of data generated by remote sensing satellites has increased rapidly. Statistics show that the daily remote sensing data generated by some countries has reached the order of TB, and the transmission rate of remote sensing satellites has also exceeded Gbps [1]. In the future, remote sensing satellites will develop in the direction of higher resolution, faster transmission rate, and stronger on-board processing capability, which undoubtedly puts forward higher requirements for ground receiving equipment.

The progress of digital chip processing capability provides a broad space for the development of software radio. All-digital receiver has gradually replaced the traditional analog receiving and demodulation mode with its high-efficiency and flexible signal processing capability and has become the mainstream of broadband wireless communication receiver at present. However, large bandwidth remote sensing satellite data transmission will inevitably lead to extremely high data sampling rate at the receiver. In this high-speed environment, even a simple filtering operation will lead to the saturation of the hardware processing module, which cannot meet the requirements of real-time demodulation. Therefore, the high parallelism of demodulation module of remote sensing satellite data ground receiver has become an inevitable development trend.

A common method of multicarrier parallel demodulation is orthogonal frequency division multiplexing (OFDM), which uses multiple mutually orthogonal carriers to realize parallel demodulation. However, the peak-to-average ratio characteristic of the OFDM system requires higher transmission power of remote sensing satellites, and its special frequency reuse mechanism makes it highly dependent on synchronization accuracy, which is more sensitive in the high Doppler frequency shift environment of satellite–ground links. In order to find a high-speed digital demodulation technology more suitable for remote sensing signal reception, relevant scholars have done a lot of fruitful research on the technology of single-carrier high-speed parallel receiver. The earliest time-domain parallel demodulation method based on channelization decomposition was put forward by Jet Propulsion Labs (JPL) of NASA in 1994 [2], which is called PRX. This method uses a filter bank technology to decompose broadband signal into multiple parallel data for matching filtering and synchronous calculation, effectively reducing the calculation amount of single signal. In 1997 [3], JPL improved the PRX structure and put forward the parallel demodulation technology of the APRX structure. This technology performs demodulation operation in the frequency domain, which greatly improves the hardware consumption compared with the PRX structure. Therefore, the research of single-carrier parallel demodulation technology by later generations, without exception, takes the
APRX structure as the basic framework for improvement [4], [5], [6], [7], [8], [9].

However, the system still has the following defects.

1) Because the matched filtering operation is carried out in the frequency domain, it is affected by discrete Fourier transform (DFT) frequency resolution when filtering out high-frequency components, which brings some errors.

2) The order of matched filter is influenced by the number of channelization. When the number of channel division is large, the order of matched filter may be large.

3) Frequency-domain timing synchronization cannot complete the difference filtering operation, resulting in that the signal sampling rate must be an integer multiple of the information symbol rate, and the flexibility is not good.

Statistics show that in the research results of high-speed demodulation technology based on the APRX architecture, the relationship between sampling rate and symbol rate is fixed at four times, so only fixed-rate demodulation can be realized, which obviously does not conform to the development orientation of universality of all-digital receivers. On the other hand, with the rapid development of digital processing chip technology, the hardware cost for the developing parallel demodulation module is greatly reduced, which provides a good foundation for the research and development of parallel demodulation architecture based on polyphase filter banks. At the same time, this architecture realizes parallel demodulation in time domain and can interpolate the best decision time, which naturally solves the problem that the sampling rate is not an integer multiple of the symbol rate. Therefore, with its excellent BER performance and high flexibility, it caters to the original development intention of software-defined communication system. The schematic diagram is shown in Fig. 1. Therefore, the author believes that the channelized architecture of polyphase filter banks will surely set off a new wave of research on parallel demodulation technology of remote sensing signal in the future.

This article aims to design a parallel matched filtering algorithm based on the multirate filter bank architecture, which can complete the matched filtering calculation and subsequent processing in parallel and synchronously at a low processing rate, so as to meet the challenge brought by the increasing data transmission rate to the real-time processing capability of digital receivers.

The rest of this article is organized as follows. Section II designs the overall architecture of parallel demodulation based on multirate digital filter bank channelization technology and analyzes its perfect reconstruction conditions. Section III deduces the parallel matched filtering technology and designs its implementation structure. Section IV carries out performance simulation analysis for the system designed above. Finally, Section V concludes this article.

II. PARALLEL DEMODULATION ARCHITECTURE BASED ON MULTIRATE DIGITAL FILTER BANKS

A. System Composition

The core content of parallel demodulation architecture based on polyphase filter banks mainly include two parts: 1) wideband signal channelization technology that meets the condition of complete reconstruction; and 2) parallel matched filtering technology. The system block diagram is shown in Fig. 2.

The receive intermediate frequency signal is digitally sampled, and that operation of downconversion to baseband is complete in the digital domain. Serial-to-serial conversion of high-speed data stream is realized by channelization decomposition of polyphase filter bank. After parallel matched filtering, carrier and timing synchronization, the system is processed by logic control unit for parallel-to-serial conversion, and finally the demodulation result is output. It should be noted that the decoding and decision process of information symbols is also carried out in parallel in sub-bands, and each sub-band outputs the decision results correctly and orderly through the predesigned logic relation operation.

B. Principle of Operation

The channelization technology based on the polyphase filter bank theory can divide a high-speed data stream into multiple parallel low-speed data in the frequency domain, which is beneficial to the parallel calculation of digital processing module. Among them, filter banks based on the polyphase DFT structure are popular among researchers as a common parallel structure. However, this architecture requires that the number of channels...
must be an integer multiple of the decimation factor, which reduces its flexibility. When designing an all-digital receiver, we hope to get a channelization technology with a flexible and variable decimation rate, so as to improve the applicability of the receiver. The channelization method of filter banks based on weighted overlap-add (WOLA) structure is a generalized form of DFT filter banks, and the selection of decimation factors is independent of the number of channels, so it can be flexibly set [9]. In this article, the channelization technology based on WOLA structure filter bank will be used to realize parallel matched filtering operation. By reasonably selecting the size of decimation factor, the demodulation has high bit error performance, and at the same time, the processing rate is reduced to the maximum.

The theoretical basis of this method is complex exponential modulation filter bank technology, which is mainly composed of analysis filter bank and synthesis filter bank, as shown in Fig. 3. After passing through the analysis filter bank, the signal is divided into several sub-bands in different frequency domains, and the synthesis filter bank reconstructs the signals of different sub-bands into original signals.

1) Polyphase Analysis Channelizer: The channelization decomposition technology based on WOLA structure filter banks is derived as follows. As shown in Fig. 3, the analysis process of the kth subchannel can be described by the formula as follows:

\[
X_k(m) = \left[ x(n)e^{-j\omega_k n} \right] * h(n) |_{n=mR} = \sum_{i=-\infty}^{\infty} h(i) e^{-j\omega_k i}
\]

where \( x(n) \) is the input signal, \( \omega_k = \frac{2\pi k}{K}, k = 0, 1, \ldots, K-1 \), represents the center frequency of the frequency band where the kth channel is located, \( h(n) \) is the prototype low-pass filter, and \( R \) is the decimation multiple. Let \( q = i - mR \), (1) can be further deformed as

\[
X_k(m) = \sum_{q=-\infty}^{\infty} h(-q) x(q + mR) e^{-j\omega_k (q + mR)}
\]

\[
e^{-j\omega_k mR} \sum_{q=-\infty}^{\infty} h(-q) x(q + mR) e^{-j\omega_k q}.
\]

We define functions

\[
x_m(q) = h(-q) x(q + mR)
\]

Following (3) and (4), we further rewrite (2) as

\[
X_k(m) = e^{-j\omega_k mR} \sum_{q=-\infty}^{\infty} x_m(q) e^{-j\omega_k q}
\]

\[
= e^{-j\omega_k mR} \tilde{X}_k(m),
\]

Observing (4), it is easy to find that \( \tilde{X}_k(m) \) is actually the DFT of \( x_m(q) \). Further define \( q = p + lK \), where \( K \) is the number of divided channels, \( p = 0, 1, \ldots, K-1 \), and \( l \) is an integer. Bring it into (5) to get

\[
\tilde{X}_k(m) = \sum_{p=0}^{K-1} \sum_{l=-\infty}^{\infty} x_m(p + lK) e^{-j\omega_k (p + lK)}
\]

\[
= \sum_{p=0}^{K-1} \left[ \sum_{l=-\infty}^{\infty} x_m(p + lK) e^{-j\omega_k p} \right] e^{-j\omega_k lK}.
\]

Define functions

\[
\tilde{x}_m(p) = \sum_{l=-\infty}^{\infty} x_m(p + lK)
\]

\[
= \sum_{l=-\infty}^{\infty} x(mR + p + lK) h(-p - lK).
\]

We can simplify (6) to

\[
\tilde{X}_k(m) = \sum_{p=0}^{K-1} \tilde{x}_m(p) e^{-j\omega_k p}.
\]

According to the abovementioned derivation process, the realization model of channelization decomposition method of filter banks based on WOLA structure can be obtained as shown in Fig. 4.

As can be seen from the figure, the channelization decomposition model based on WOLA structure adopts the form of data stream segmentation, and every \( R \) consecutive data points are input into the system as a division. For every \( R \) data input,
the subchannel at the output generates one data, thus realizing
\( R \) times decimation. Among them, the complex modulation
operation process of formula (5) is realized by cyclic shift in the
time domain, and it can be obtained from the signal processing
theory that the complex modulation operation in the frequency
domain is equivalent to the cyclic shift operation in the time
domain, that is,
\[
Y(\omega) = X(\omega) e^{-j\omega k_0} \iff y(k) = x((k-k_0)_{\text{mod} K}). (9)
\]

Therefore, in the process of implementation, the order of FFT
operation and product operation can be adjusted, cyclic shift
operation can be carried out in advance, and the number of shift
samples is the modulus of \(-Rm\) and \(K\), so as to efficiently
realize frequency-domain weighting operation.

2) Polyphase Synthesis Channelizer: The process of synthe-
sizing the signals of \( K \) subchannels into single-channel signals
can be described by
\[
\hat{x}(n) = \frac{1}{K} \sum_{k=0}^{K-1} Y_k(n) \ast f(n) \times e^{j\omega_k n}
\]
\[
= \frac{1}{K} \sum_{k=0}^{K-1} \sum_{m=-\infty}^{\infty} \hat{X}_k(m) f(n - mR) e^{j\omega_k n} (10)
\]
where \( f(n) \) is the synthesis filter. And,
\[
Y_k(n) = \begin{cases} \hat{X}_k(m), & n = mR \\ 0, & \text{else} \end{cases} . (11)
\]

Note that in the actual system, the number of points of the
synthesis filter is finite length \( L \), so the value range of \( m \) in (10)
must satisfy \( 0 \leq n - mR \leq L - 1 \). Let \( n = r + M R \), where
\( r = 0, 1, \ldots, R - 1 \) and \( M \) is an integer. Therefore, the value
range of available \( m \) is \( M - \text{ceiling}(L/R) \leq m \leq M \), where
\( \text{ceiling}(\cdot) \) indicates the rounding in the direction of \(+\infty\). Further
set \( \alpha = \text{ceiling}(L/R) \), then (10) can be rewritten as (12), where
\( \hat{X}_k(m) = \hat{X}_k(m)e^{j\omega_k mR} \). Define functions [Equ. (12) shown at
the bottom of this page]
\[
U_m (r + (M - m) R) = \frac{1}{K} \sum_{k=0}^{K-1} \hat{X}_k(m) e^{j\omega_k (r + (M - m) R)} , (13)
\]
\[
\hat{x}(r + MR) = \frac{1}{K} \sum_{k=0}^{K-1} \sum_{m=M-\alpha}^{M} \hat{X}_k(m) f(r + MR - mR) e^{j\omega_k (r + MR)}
\]
\[
= \sum_{m=M-\alpha}^{M} f(r + (M - m) R) \frac{1}{K} \sum_{k=0}^{K-1} \hat{X}_k(m) e^{j\omega_k mR} e^{j\omega_k (r + (M - m) R)}
\]
\[
= \sum_{m=M-\alpha}^{M} f(r + (M - m) R) \frac{1}{K} \sum_{k=0}^{K-1} \hat{X}_k(m) e^{j\omega_k (r + (M - m) R)} . (12)
\]

Fig. 5. Parallel implementation model of synthesis filter banks with WOLA
structure.

Since \( e^{j\omega_k (r + (M - m) R)} \) is a periodic function with a period of
\( K \), (13) can be rewritten as
\[
U_m (r + (M - m) R) = \frac{1}{K} \sum_{k=0}^{K-1} \hat{X}_k(m) e^{j\omega_k (r + (M - m) R)_{\text{mod} K}}
\]
\[
= U_m ((r + (M - m) R)_{\text{mod} K}) . (14)
\]

According to (14), (12) can be transformed into
\[
\hat{x}(r + MR) = \sum_{m=M-\alpha}^{M} \left[ f(r + (M - m) R) \frac{1}{U_m ((r + (M - m) R)_{\text{mod} K})} \right]
\]
\[
= f(r)U_M(r) + f(r + R)U_{M-1}((r + R)_{\text{mod} K})
\]
\[
+ \cdots + f(r + \alpha R)U_{M-\alpha}((r + \alpha R)_{\text{mod} K}) . (15)
\]

According to the abovementioned derivation process, the par-
allel implementation model of the multiphase synthesis method
with the WOLA structure is designed, as shown in Fig. 5, in
which the weighting and superposition operations of data are

carried out in sub-bands.

1) Sub-band Data Weighting: First, the \( L \)-point synthesis fil-
ter is polyphase decimated to obtain \( K \) groups of weighted
data points with the length of \( L/K \), and then these \( L/K \)
data points are multiplied by the output samples, respect-
ively, and each output sample point can obtain \( L/K \)
weighted data.

2) Sub-band Data Addition: After polyphase decimation, the
shift register of length \( L \) is distributed to each sub-band
as the buffer data for addition operation. It should be

noted that since the decimation factor \( R \) is allowed not
to divide \( K \) evenly, the \( R \) sub-bands of the output data do
K
K
K
K
K
1

\cos
F
×
⩽
H
⩽
|
1+
A
is even, there are
β
H
β
is the result of shifting raised cosine
×
=2
|
β
is the cutoff frequency of filter and
(19)
<β
\cos
1
=1
1
1
+2
is the
−
ω
1
1
×
1
1
to the right-hand side by
(16)
π/K
42x50
thus reducing the interference of aliasing parts in sub-bands.

practical application. By using the form of nonmaximum deci-
completely eliminated, but it can be suppressed to be ignored in
sub-band part is set to zero, the aliasing error generated at the
boundary of the reserved sub-band cannot be filtered out by the
synthesis filter, resulting in error interference. For this kind of
aliasing error caused by partial sub-band reconstruction. Assuming that the prototype filter
has perfect reconstruction characteristics, when the maximum
decimation filter is applied, the schematic diagram of aliasing
effect generated by the reconstructed partial sub-bands is shown
in Fig. 6.

not form a fixed proportional relationship with the total
number of channels \( K \), which will cause data transmission
of buffered data among sub-bands.

C. Perfect Reconstruction Condition of Discarding Partial
Sub-bands

Generally, the distortion sources of filter banks mainly in-
clude aliasing, amplitude distortion, and phase distortion. In
the universal application of channelization of full-band signal
reconstruction, prototype filters that meet the perfect re-
construction characteristics can be used to eliminate all the
abovementioned three kinds of distortions. However, in this
article, after channelization, only the data transmission signals
in the baseband part are concerned, so there is no need to
reconstruct the interference signals in the higher frequency band.
This model reconstructs only a part of the required channels,
which has the effect of the low-pass filter, but it also introduces
special aliasing interference. Assuming that the prototype filter
has perfect reconstruction characteristics, when the maximum
decimation filter is applied, the schematic diagram of aliasing
effect generated by the reconstructed partial sub-bands is shown in
Fig. 6.

From the analysis in the figure, it can be seen that when the
maximum decimation filter is applied, because the abandoned
sub-band part is set to zero, the aliasing error generated at the
boundary of the reserved sub-band cannot be filtered out by the
synthesis filter, resulting in error interference. For this kind of
aliasing error caused by partial band reconstruction, it is difficult
to eliminate only by using perfect reconstruction characteristics.
Therefore, in this article, a nonmaximum decimation filter bank
based on WOLA structure is used to realize partial sub-band
reconstruction to filter out aliasing. It should be pointed out that
due to the nonideal characteristics of the filter, aliasing cannot be
completely eliminated, but it can be suppressed to be ignored in
practical application. By using the form of nonmaximum deci-
mation filter banks, the interval of spectrum images is increased,
thus reducing the interference of aliasing parts in sub-bands to
useful signals. And, in this case, the transition band of the filter
need not be very narrow, which relaxes the design requirements
of the prototype filter.

We continue to consider the perfect reconstruction condition
of the prototype filter under the condition of nonmaximum
decimation filter bank. In order to eliminate amplitude and phase
distortions, the designed prototype filter must have linear phase
characteristics and satisfy the following requirements:

\[
A(z) = \frac{1}{K} \sum_{k=0}^{K-1} H_k(z) F_k(z) = 1
\]

(16)

where \( A(z) \) is the distortion function of the system and \( H_k(z) \)
and \( F_k(z) \) are the system functions of the analysis and synthesis
filters in the sub-bands, respectively.

The filter that meets the abovementioned conditions is called
a Nyquist filter, and the raised cosine filter is a Nyquist filter
with good performance. In this article, the prototype filter is
designed by adding a window function to the ideal low-pass
filter. The window function is obtained by intercepting the square
root raised cosine function, and the raised cosine function is
expressed as

\[
H^2(e^{j\omega}) = \begin{cases}
1, & \frac{\omega}{\omega_c} \leq 1 - \beta \\
\frac{1}{2} + \frac{1}{2} \cos \left( \frac{\pi}{\omega_c} \left( \frac{\omega}{\omega} - (1 - \beta) \right) \right), & 1 - \beta < \frac{\omega}{\omega_c} < 1 + \beta \\
0, & \frac{\omega}{\omega_c} \geq 1 + \beta
\end{cases}
\]

(17)

where \( \omega_c \) is the cutoff frequency of filter and \( 0 < \beta \leq 1 \) is the
roll-off coefficient of filter.

The following examples verify the power complementary
characteristics of the window function, and we assume \( \beta = 1 \),
\( \omega_c = 2\pi/K \). \( H^2_k(e^{j\omega}) \) is the result of shifting raised cosine
function \( H^2(e^{j\omega}) \) to the right-hand side by \( 2\pi k/K \). Then, the
following equation is obtained:

\[
A' \left( e^{j\omega} \right) = \frac{1}{K} \sum_{k=0}^{K-1} H^2_k \left( e^{j\omega} \right)
\]

\[
= \frac{1}{K} \sum_{k=0}^{K-1} \left[ \frac{1}{2} + \frac{1}{2} \cos \left( \frac{\pi}{\omega_c} \left( \frac{\omega}{\omega} - \frac{2\pi k}{K} \right) \right) \right]
\]

\[
= \frac{1}{K} \times \frac{K}{2} + \frac{1}{2K} \cos \left( \frac{K}{2} \omega \right) \times \sum_{k=0}^{K-1} (-1)^k.
\]

(18)

When \( K \) is even, there are

\[
A' \left( e^{j\omega} \right) = \frac{1}{2} + \frac{1}{2K} \times 0 = \frac{1}{2}.
\]

(19)

This satisfies the power complementary condition.

In the actual design of the prototype filter, the order of the filter
with limited length cannot completely eliminate the aliasing and
amplitude distortion errors, but the error level can be far lower
than the thermal noise level by reasonably selecting the roll-off
coefficient \( \beta \) and the filter length \( L \), so as to ignore its influence.
III. PARALLEL DERIVATION OF DIGITAL MATCHED FILTERING

The parallel matched filter calculation can be derived from the serial matched filter, as shown in Fig. 7, which is a schematic diagram of matched filter calculation after the signal passes through the complex exponential modulation filter bank. When the analysis/synthesis filter meets the perfect reconstruction condition, it can realize the signal demodulation function.

According to the theory of multirate signal processing, after the signal $X(z)$ is decimated by $M$ times or the signal $Y(z)$ is interpolated by $M$ times, it can be expressed by (20) and (21) in the Z-transform domain as follows:

$$Y(z) = \frac{1}{M} \sum_{k=0}^{M-1} X(z^{1/M} W_M^k)$$  \hspace{1cm} (20)$$

$$X(z) = Y(z^{1/M}) $$  \hspace{1cm} (21)$$

where $W_M = e^{-j2\pi/M}$. From this, the Z-transform domain expression of each point signal in the complex exponential modulation filter bank of Fig. 7 can be derived as

$$R_k(z) = R(z)H_k(z)$$  \hspace{1cm} (22)$$

$$U_k(z) = \frac{1}{M} \sum_{i=0}^{M-1} R_k(z^{1/M} W_M^i)$$  \hspace{1cm} (23)$$

$$V_k(z) = U_k(z^{1/M}).$$  \hspace{1cm} (24)$$

Combining the results of (22)–(24), we can describe the system shown in Fig. 7 as follows:

$$Y(z) = \frac{1}{M} \sum_{k=0}^{K-1} [F_k(z)S_{M,k}(z)] \cdot Q(z)$$  \hspace{1cm} (25)$$

where $S_{M,k}(z) = \sum_{i=0}^{M-1} R(z^{1/M} W_M^i) H_k(z^{1/M} W_M^i)$. $K$ is the parallel number of channels, and $M$ is the decimation factor. In this article, nonmaximum decimation filter banks are adopted, i.e., $M < K$. In order to realize the parallel matched filter calculation, (25) is rewritten as

$$Y(z) = \frac{1}{M} \sum_{k=0}^{K-1} [Q(z)F_k(z)S_{M,k}(z)]$$

$$= \frac{1}{M} \sum_{k=0}^{K-1} [P_k(z)S_{M,k}(z)]$$  \hspace{1cm} (26)$$

where $P_k(z) = Q(z)F_k(z)$. We define $F_k'(z)$ to satisfy

$$P_k(z) = F_k'(z) \cdot \sum_{i=0}^{M-1} P_k(z^{1/M} W_M^i).$$  \hspace{1cm} (27)$$

Then, the following equation can be obtained:

$$Y(z) = \frac{1}{M} \sum_{k=0}^{K-1} \left[ \sum_{i=0}^{M-1} P_k(z^{1/M} W_M^i) \right] S_{M,k}(z^{1/M}) F_k'(z).$$  \hspace{1cm} (28)$$

According to the general identity of multirate convolution

$$X(z^M) = X_1(z^M)X_2(z^M),$$

the interpolation operation in (28) can be interchanged with the filtering calculation, so (28) can be further rewritten as

$$Y(z) = \frac{1}{M} \sum_{k=0}^{K-1} \left[ \sum_{i=0}^{M-1} P_k(z^{1/M} W_M^i) \right] S_{M,k}(z^{1/M}) F_k'(z).$$  \hspace{1cm} (29)$$

Equation (29) can transfer the matched filtering calculation to sub-band, and its implementation block diagram is shown in Fig. 8. The convolution–decimation and interpolation–convolution operations in the figure can be reduced by using the theory of multirate filter banks, and the part in the dashed box can be precalculated according to the prior information of demodulation, which reduces the difficulty of real-time demodulation of data transmission signals. In this article, the analysis filter bank based on WOLA structure is used to reduce the speed of baseband signal channelization and sub-band matched filter calculation, and the synthesis filter bank based on WOLA structure is used to reduce the speed of demodulation signal parallel-to-serial conversion.

According to the abovementioned derivation process, it can be seen that the focus of designing parallel matched filter lies in designing analysis/synthesis filters $H_k(z)$ and $F_k(z)$ that meet the perfect reconstruction conditions, and constructing $F_k'(z)$ that meets (27). The perfect reconstruction condition of filter
Fig. 9. Frequency spectrum change of signal decimation and interpolation.

Fig. 10. Frequency response model of the synthetic filter.

Fig. 11. Implementation block diagram of parallel matched filtering.

bank and the design of prototype filter have been discussed in Section II. We next discuss the implementation for the filter $F'_k(z)$.

Observing (27), we can find that $\sum_{i=1}^{M-1} P_k(zW_i)$ is to decimate $P_k(z)$ first and then interpolate it, and its spectrum diagram is shown in Fig. 9.

For (27) to hold, the filter $F'_k(z)$ needs to remove the image part of $P_k(z)$ after decimation and interpolation and only keep the baseband part. Because the nonmaximum decimation filter bank is adopted, the passband bandwidth of the signal is less than the interval of spectral images, so there is a certain interval between adjacent images, which greatly reduces the design difficulty of the synthesis filter $F'_k(z)$. Fig. 10 shows the frequency response model of $F'_k(z)$. To ensure that the image part is adequately suppressed, the stop band attenuation of filter $F'_k(z)$ is usually less than $-60$ dB.

Based on the abovementioned analysis, we get the block diagram of parallel matched filtering technology of filter banks based on WOLA structure, as shown in Fig. 11, and the concrete implementation process can be divided into the following four steps.

a) Calculate each sub-band matched filter in advance according to the prior knowledge of the data transmission signal and store it in the buffer.
b) When the system receives the baseband data transmission signal, it parallelizes the data into $K$ channels by using the analysis filter bank of WOLA structure.
c) The matched filter calculation is performed in the sub-band.
d) The synthesis filter bank of WOLA structure is used for parallel-to-serial conversion, and the data transmission signal is judged and sent to the back-end operation module.

IV. SIMULATION RESULTS

This section simulates the satellite-to-ground transmission system of remote sensing satellites commonly used in the world at present, simulates the multiphase parallel demodulation technology of the receiver, and verifies the effectiveness of the abovementioned design method. X-band QPSK modulation signal is used for transmission and the code rate is 160 MHz, which can realize 320 Mbps remote sensing data transmission.

The simulation receiving module of this article first downconverts the radio frequency signal to 720-MHz intermediate frequency, and then uses the 960-MHz data acquisition card for A/D conversion to obtain the digital signal. The configuration of simulation parameters is listed in Table I.

The matched filter module of polyphase parallel demodulation is designed as follows. Because the passband of synthesis filter $F'_k(z)$ needs to include the transition band of the analysis filter $H(z)$, the choice of roll-off factor of the root raised cosine filter will be limited. In the channelization design of this article, the roll-off factor $\beta$ of the prototype filter must meet the constraint of

$$ (1 + \alpha) \frac{1}{K} + \Delta \omega \leq \frac{1}{R} $$(30)

where $\Delta \omega$ is the transition band bandwidth of the synthetic filter $F'_k(z)$. When the order of the filter used in this article is 1152, the simulation shows that its normalized transition bandwidth is about $2.6 \times 10^{-3}$, so the required roll-off factor of the prototype filter can be calculated as

$$ \alpha \leq K \left( \frac{1}{R} - \Delta \omega \right) - 1 \approx 0.61. $$

(31)

$\beta = 0.6$ is selected in this simulation, and the unit impulse response and system function of the prototype filter are designed, as shown in Fig. 12.
When the simulated signal-to-noise ratio is 6 dB, the demodulated constellation can be obtained by using the parallel matched filtering method proposed in this article, as shown in Fig. 13.

The demodulation error rate of this method is simulated in the following, assuming that the channel is Gaussian and there is no synchronization error. When the carrier-to-noise ratio of the transmission signal $E_b/N_0 \in [6, 10]$, it is compared with the theoretical error rate curve, and the result is shown in Fig. 14.

When the bit error rate $E_b/N_0$ of simulation data is fixed to 10 dB. Demodulating with the method proposed in this article, the obtained BER varies with the order $L$ of the prototype filter, as shown in Fig. 15.

When the number of channels is fixed at $K = 64$ and the decimation factor $R$ is changed. According to formula (30), the roll-off factor beta of the prototype filter is calculated. Table II lists some simulation parameters in this case. And, the variation of simulation error rate with $R$ is shown in Fig. 16.

From the experimental analysis, it can be seen that the demodulation error rate decreases with the increase of prototype filter order $L$ and increases with the increase of decimation factor $R$. In practical application, the corresponding parameters can be changed according to the actual needs to obtain more suitable performance.

From the abovementioned analysis, it can be found that the parallel matched filtering algorithm adopted in this article is close to the theoretical limit in demodulation error performance and has obvious advantages in demodulation speed under the condition of large bandwidth and high bit rate of remote sensing satellite data, so it has a wide application space.
Fig. 16. Error rate curve changes with respect to $R$.

V. CONCLUSION

Aiming at the high processing rate of remote sensing satellite data reception, this article used the channelization technology of filter banks based on WOLA structure to execute the matched filtering operation in parallel, which greatly improved the operation efficiency and reduced the requirement of hardware processing speed for real-time demodulation. The polyphase filter bank structure that meets the perfect reconstruction characteristics was designed, and its oversampling factor was flexible and variable, which reduced the difficulty of designing prototype analysis/synthesis filter. A parallel implementation structure of synthetic filter banks based on WOLA structure was designed, which can carry out sampling decision and subsequent synchronous estimation calculation in sub-band. The simulation results showed that the parallel matched filtering technology of filter banks based on WOLA structure approached the theoretical limit in demodulation error performance, and there was no performance loss. In addition, the parallel architecture proposed in this article has good scalability, is suitable for flexible modulation system and symbol rate, and has a wide application prospect in different remote sensing data transmission systems.

The next work of this article is mainly focused on the following two aspects.

1) The analog demodulation experiment of higher order modulation mode and wider bandwidth signal is carried out to verify the scalability of the method.

2) Improve timing and carrier synchronization technology based on the polyphase filter bank demodulation architecture to optimize demodulation performance.

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