

SIGNAL PROCESSING TECHNIQUES FOR WIDEBAND COMMUNICATIONS SYSTEMS

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Abstract - In order to provide for a planning of the development and conceptual design of a wideband satellite system, various signal processing techniques required by the satellite payload are investigated in this paper. These techniques include polyphase filtering, Fast Fourier Transform (FFT) filtering, and tree filter bank techniques. This paper reviews and compares these techniques in a unified framework to assess their applicability to the design of a wideband satellite system.

Introduction

The present and future generation of satellites needs to operate with small earth terminals as with mobile vehicle or hand held terminals. Such satellites need to be "user oriented" in that the user terminal needs to be relatively less complex and have small power, weight and low cost requirements. Such an arrangement may be achieved at the cost of increasing the complexity of the space borne equipment and the central earth station(s). Present day technology involves such processing on board the satellite. An example of such a system is mobile messaging service via satellites. In such a system, a forward link takes messages from an earth-station to the satellite, which retransmits to the mobiles over spot beams. The return link begins at the mobiles, goes up to the satellite and terminates at the earth station. In such a system, the types of transmitter/receivers in the forward and reverse link are quite different as explained above. Thus the optimum uplink and downlink designs to and from the mobiles (and the earth stations) may be very different.

In some architectures proposed earlier, the uplink uses FDMA techniques with low cost and complexity terminals while the downlink uses TDMA technique to maximize the satellite radiated power without intermodulation noise. In such systems the small earth terminals do not need the capability of transmitting at very high burst rate and stringent satellite frame synchronization capabilities necessary for TDMA transmitter. In another architecture, uplink is based on random access technique while the downlink uses TDMA. In terms of modulation, all such architectures make use of digital techniques with inherent advantages in terms of power efficiency, flexibility, error correction and

detection coding and encryption. The feasibility of mixed mode multiple accessing techniques requires efficient means of translation between the two formats of MA techniques. Although analog techniques are in principle straightforward, in terms of implementation considerations of size, weight, cost and flexibility, direct digital techniques are expected to perform better. Digital techniques can also fully exploit advances in VLSI and ASIC technologies to achieve these objectives. Such translation may involve conversion of an FDMA signal into a TDM multiplexed signal, which may then go through digital switch to various TDMA carriers being transmitted over the spot beams. Such digital translation techniques are also useful in switching of FDMA carriers to different spot beams with out requiring arrays of analog bandpass filters and converters. There are several techniques investigated in the literature for direct digital translation and/or switching. In the following these channelization techniques are reviewed and compared in a unified framework.

Digital Channelization Schemes

Currently, there are several techniques available for channelization, namely, analytical signal approach, polyphase/Discrete Fourier Transform (DFT) approach, frequency domain filtering (or FFT filtering) approach, and tree filter bank (or multistage approach). These techniques will be described briefly below, and the readers are referred to [1-10] for more details.

The Analytical Signal (AS) Approach:

The A.S. method is a per-channel (modular in the number of channels) method that utilizes the signal's analytical property to reduce the channelizer complexity. This method allows the channelizer's filter specifications to be relaxed and hence, reduces the implementation complexity. In this method [1] depicted in Figure 1, the high rate sampled FDM signal $S(f)$ with N_c number of multiplexed channels, (after appropriate analog down conversion of the received signal to IF range) is first filtered by an analytic complex band pass filter $\bar{H}_1(fT_u)$ where

$\bar{H}_i(fT_u)$ is a frequency translated version of a prototype low pass filter $\bar{H}(fT_u)$ such that

$$\bar{H}_i(fT_u) = \bar{H}\left[\left(f - \left(i + \frac{1}{2}\right)w\right)T_u\right]; i=0,1,\dots,(N_c-1) \quad (1)$$

where T_u is the input signal sampling period and w denotes bandwidth of any one of the multiplexed channels (assumed to have equal bandwidth). The filter output is decimated by N_c in the decimator following the filter. The output of decimator is filtered by a complex low pass filter with frequency response $\bar{G}_i(fT_d)$ to yield the channelized signal in analytical form.

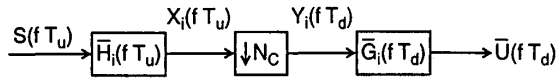


Figure 1. Analytic filtering approach of demultiplexing (complex filters)

From the sampling theorem [3], the decimator output spectrum Y in Figure 2 is related to its input X by

$$Y_i(f'_d) = \frac{1}{N_c} \sum_{j=0}^{N_c-1} X_i(f'_d - j) f'_d \equiv fT_d; T_d = N_c T_u \quad (2)$$

To eliminate any aliasing due to $j \neq 0$ terms in the above equation, it is sufficient that the transition band for filter $\bar{H}_i(fT_u)$ be limited to bandwidth w on either side of its pass band. This is illustrated in figure 2 for the channel with its index i equal to 2.

It is clear from Figure 2, that the channel signal can be recovered with no aliasing error even though filter $\bar{H}_2(fT_u)$ has a transition band of w Hz on either side. The transition band however, makes the design of the filter easier [1,2]. Clearly the output $\bar{U}(fT_d)$ is in analytic form. To obtain the corresponding real valued signal, one obtains the complex-conjugate part of the spectrum $\bar{U}(fT_d)$.

Representing analytic functions $\bar{H}_i$, $\bar{G}_i$ as sum of their respective conjugate symmetric parts H_i, G_i and anti symmetric parts H'_i, G'_i , the equivalent implementation of the analytic signal approach is given in Figure 3 below [1].

Note that the operation of multiplying by $(-1)^{in}$ (n denotes discrete time index and i is the channel number) in time domain in Figure 3 corresponds to frequency shift of w and is required for odd channels. Obviously for even channels (i even) $(-1)^{in} \equiv 1$ and no additional operation is involved.

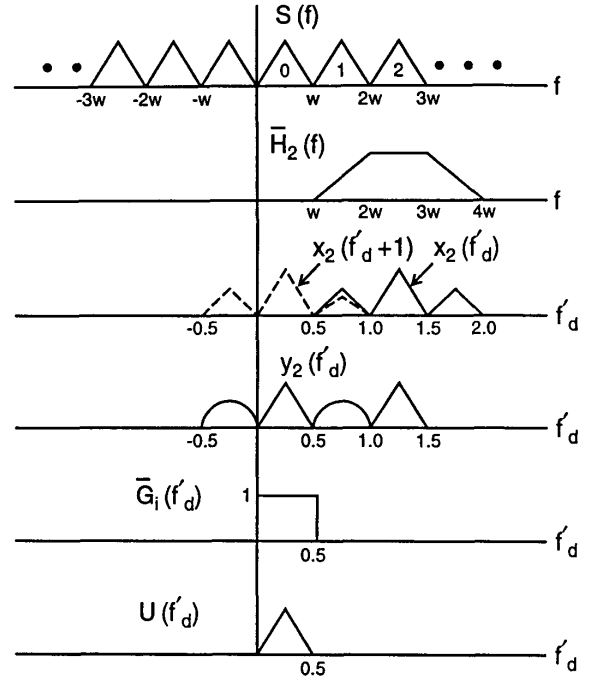


Figure 2. Signal spectra in analytic approach

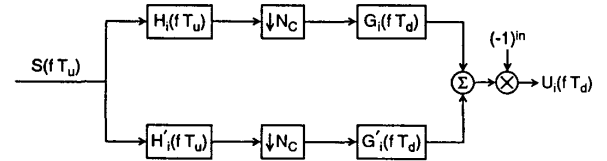


Figure 3. Demultiplexing in analytic approach (real filters)

In terms of computational complexity, the number of multiplications M_{AS} required per input channel per second is given by [2] the following equation.

$$M_{AS} = \kappa w^2 \frac{w(N_c + 4) - 2B(N_c + 2)}{(w - B)(w - 2B)} \quad (3)$$

$$\kappa = -\frac{2}{3} \log[5\delta_1\delta_2] \quad (4)$$

where δ_1 and δ_2 denote the specified in-band and out-of-band ripple respectively, and B denotes the channelizer's filtering bandwidth which in general is smaller than w to allow for guard bands.

The Polyphase/DFT (PDFT) Approach:

In the DFT filter bank [3] each channel is separately bandpass modulated by a complex modulator as

shown in Figure 4 wherein double lines refer to complex-valued signals.

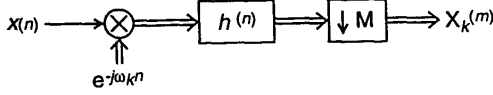


Figure 4a. Single channel of a DFT filter bank analyzer

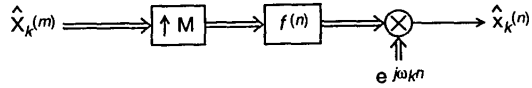


Figure 4b. Single channel of a DFT filter bank synthesizer

In the analyzer model, the input signal is demodulated by the exponential function $e^{-j\omega_k n}$, low pass filtered by the filter $h(n)$ and the resulting signal is down sampled by a factor M . The synthesizer model interpolates all the channel signals back to their high sampling rate, filters the signal by filter $f(n)$ to remove the imaging components and modulates the resulting signal by complex exponential function $e^{j\omega_k n}$ to translate it back to frequency ω_k . The output of the synthesizer is the sum of the K channel output signals, i.e.,

$$\hat{x}(n) = \sum_{k=0}^{K-1} \hat{x}_k(n) \quad (5)$$

The polyphase realization of the DFT filter bank is based on the polyphase implementation of the decimators and interpolators [4-6]. Such a realization is relatively simple for the case of critically sampled filter banks wherein $M=K$. In this case the number of independent channels N_c is also equal to K . Designs for other choices of M and K are relatively more complex. In the case of $M=K$, the center frequencies of the K frequency bands are given by

$$\omega_k = \frac{2\pi k}{K}; k=0,1,\dots,K-1; K=M$$

The analyzer-synthesizer model of figure 4 can be shown to be equivalent to the Integer band model of Figure 5.

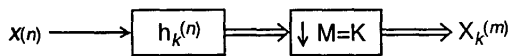


Figure 5a. Single channel of the Integer band model analyzer



Figure 5b. Single channel of the Integer band model synthesizer

In Figure 5 the filter impulse response functions are given by

$$h_k(n) = h(n)W_K^{kn}; W_K = e^{j2\pi/K} = W_M \quad (6a)$$

$$f_k(n) = f(n)W_K^{kn} \quad (6b)$$

It is apparent from Figure 5a in view of equation (6) that

$$X_k(m) = \sum_{n=-\infty}^{\infty} h(n)W_M^{kn} x(mM - n) \quad (7)$$

With change of variables $n = rM - i$, and with some appropriate manipulations, equation (7) may be deduced to equation (8) below.

$$X_k(m) = \sum_{i=0}^{M-1} \sum_{r=-\infty}^{\infty} \bar{p}_i(r) W_M^{-ki} x_i(m - r) \quad (8a)$$

$$= \sum_{i=0}^{M-1} W_M^{-ki} [\bar{p}_i(m) \otimes x_i(m)] \quad (8b)$$

where $\otimes$ denotes convolution, $\bar{p}_i(m)$ is the impulse response of the i th polyphase branch given in terms of $h(n)$ as

$$\bar{p}_i(m) = h(mM - i); i=0,1,\dots,M-1 \quad (9a)$$

and branch input signals $x_i(m)$ are given by

$$x_i(m) = x(mM + i) \quad (9b)$$

Equation (8) leads to the polyphase DFT filter bank structure of Figure 6a.

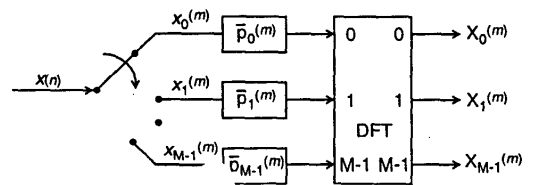


Figure 6a. Polyphase DFT filter bank analysis structure

Similarly the polyphase DFT filter bank synthesis structure is derived as shown in Figure 6b wherein

$$q_i(m) = f(mM + i); i=0,1,\dots,M-1 \quad (10)$$

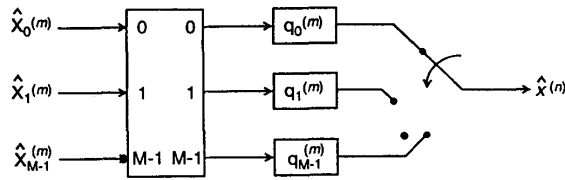


Figure 6b. Polyphase DFT filter bank synthesis structure

The polyphase implementation has the advantage of reducing the computational requirements by order K compared to direct form. In terms of polyphase branch filter design there are two broad categories of design, viz., FIR and IIR filters [7]. The FIR filters can be designed on the basis of windows (e.g. Hamming, Hanning or Kaiser), optimal equiripple linear phase design (e.g. based on Chebyshev approximation and multi-exchange Remez algorithm), Half Band filters which further reduce the computational requirements, and filter designs based on direct optimization of a criterion function. The IIR filter can be designed as in classical approach or may be based on a transformation [8] wherein the denominator is a polynomial in Z^M and thereby exploits the Interpolator/Decimator structure to minimize the computational requirements [3].

The polyphase implementation effectively allows sharing one lowpass filter among all the channels with the help of FFT transform. The total number of real multiplications required per second per channel is found to be [2]

$$M_{\text{PDFT}} = 2w \left[\frac{\left[\frac{2}{3} \log \left(\frac{1}{10\delta_1\delta_2} \right) \right]}{(w - 2B)} + 4 \log_2(N_C) \right] \quad (11)$$

where the parameters found in (11) have been defined earlier.

Frequency Domain Filtering (FDF) Approach:

Frequency domain filtering is based on the use of FFT techniques in the filtering operation. In time domain, the filtering operation consists of discrete convolution of the sampled input signal with the filter impulse response. Equivalently the result can be obtained by multiplying the Fourier transform of the input signal with filter frequency response and taking the inverse transform of the result. This is the basis of FDF techniques. Even though this approach may seem to be more indirect compared to direct convolution used in FIR filter implementation, this can be made computationally more efficient by using

FFT technique in computing the direct and inverse Fourier transform. However, the application of FFT results in a circular convolution of the input signal (segment) and the filter impulse response instead of the desired linear convolution. This problem is overcome by an appropriate modification of the straightforward FFT approach. Two such modification techniques called, "Overlap-Save Sectioning" and "Overlap-Add Sectioning", are well known in the literature [3]. For a brief discussion of these techniques, let Q denote the filter (impulse response) length and N ($N > Q$) be the sequence length selected for the FFT operation. N can of course be selected in an optimal manner so as to minimize the overall computational complexity.

In the overlap-save section method, the incoming signal is segmented into sections of length N such that the adjacent sections have an overlap of $(Q-1)$ samples. Each such section is circularly convolved with the filter response (also of length N after padding with zeros) using FFT approach. The first $(Q-1)$ samples of the result are discarded for each section and the truncated sections are concatenated to yield the desired linear convolution. In the overlap-add sectioning method, the input signal is segmented into disjoint sections of length $(N-Q)$. Each section is augmented by a sequence of zeros of length Q to yield a sequence of length N , which is circularly convolved with the augmented filter impulse response using FFT techniques. The resulting sequences are aligned in such a manner that there is an overlap of length Q between the successive sequences. During the period of no overlap, individual sequences then provide the desired response. During the periods of overlap, the two overlapping sequences are added to yield the desired output. It can be shown [3] that both the methods provide the desired linear convolution.

Figure 7 shows a simplified block diagram for demultiplexing of FDM signal using DFF approach.

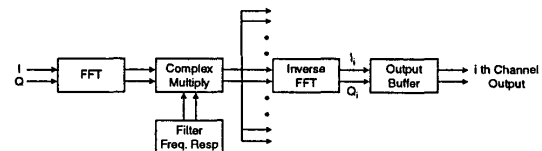


Figure 7. Demultiplexing using frequency domain filtering

A specific channelization scheme based on the FDF approach is described in [9]. This scheme uses overlap-save approach with 50 % overlap, i.e.,

$N=2Q$. The channelization scheme of [9] considers FDMA signal of 6 MHz bandwidth consisting of 300 channels of 20 KHz bandwidth each. From simulations it was determined that in the frequency domain each channel must have at least 16 sample points corresponding to a resolution of 1.25 KHz. Thus for the complete band, at least 4800 points are required. Rounding up to the nearest power of 2, an FFT size of 8192 was selected. In terms of N_c , the length for FFT operation is $N \cong 16N_c$. The number of multiplications for an FFT or IFFT of size is given by $N \log_2 N$. For the implementation requiring only one IFFT, the number of multiplications is equal to $2N[\log_2 N + 4]$. For real time operation all these operations must be performed in $(NT_s/2)$ sec, where T_s is the sampling period of the FDMA signal and the factor 2 accounts for 50 % overlap. Therefore the number of multiplications per second is given by

$$M_{\text{DFF}} = 4f_s \log_2 [16N_c + 4] \quad (12)$$

In (12) $f_s = 1/T_s$ is the sampling rate selected equal to 10.24 MHz. In practice the number of IFFTs will be determined by the number of outputs of the digital transplexer which may be connected to different spot beams in the satellite communication applications. For example in [9] there are 12 beams in the example analyzed. In this case the number of operations given by (12) is roughly multiplied by the number of these outputs.

Multistage (MS) Approach:

Multistage approach provides a means of channelization using successive stages of half-bank filters [10]. Therefore, this technique is appropriate only when $N_c = 2^L$ where L denotes the number of stages of filtering and decimating. This method provides moderate flexibility and computational efficiency, but the efficiency decreases as the number of channel decreases. The total number of real multiplications required per second per channel is found to be [4]:

$$M_{\text{MS}} = \left[\left(\frac{N_F + 1}{2} + 1 \right) \left(\log_2 N_c - \frac{1}{2} \right) + N_G \right] 2w \quad (13)$$

where N_F denotes the number of coefficients of the half band filters, N_G is the number of coefficients of the last filter of the tree, and w and N_c are defined previously.

Hybrid Techniques

In a wideband system the channelization may be performed in a number of hierarchical stages and in principle different stages may apply different channelization techniques including both digital and analog to obtain the most flexible and optimum overall architecture. For example, if the total band is 500 MHz, it may be divided first in to six channels each with an 80-MHz bandwidth. This stage may be implemented using surface wave acoustic (SAW) filters or one of the digital

Table 1. Comparisons Between Various Channelizers

Architecture	Power, watts (Current/Future)	Weight, lb	Remarks
Per-channel/analytical sig.	12,4998/511	2,986	<ul style="list-style-type: none"> 80-MHz band is digitized. Channelizer employs analytic sig. filter.
Polyphase/DFT	2,358/308	707	<ul style="list-style-type: none"> 80-MHz band is digitized. Channelizer employs polyphase/DFT filter.
Hybrid Tree/FFT Option 1	1,438/362	498	<ul style="list-style-type: none"> 80-MHz band is digitized. 80- to 20-MHz band is digitized. Employs FIR, and 20 MHz to 64 kHz. Channelizer employs pipeline FFT.
Hybrid Tree/FFT Option 2	2,012/197	482	<ul style="list-style-type: none"> 500 MHz is digitized. 500- to 20-MHz channelizer employs FIR, and 20-MHz to 64-kHz channelizer employs pipeline FFT.
Hybrid Tree/FFT Option 3	4,176/197	969	<ul style="list-style-type: none"> 500-MHz band is digitized. 500- to 80-MHz channelizer employs FIR, and 80-MHz to 64-kHz channelizer employs pipeline FFT.
Hybrid Tree/FFT Option 4	720/282	344	<ul style="list-style-type: none"> 80-MHz band is digitized. 80- to 20-MHz channelizer employs ACT/FIR, and 20-MHz to 64-kHz channelizer employs pipeline FFT.
Multistage Tree (ACT)	2,138/1059	655	<ul style="list-style-type: none"> 80-MHz to 64-kHz channelization is done in analog domain using ACT technology. A/D conversion is performed at 64 kHz.

techniques. The second stage channelizer may divide the 80 MHz band into four bands of 20 MHz each. This stage may be implemented by the polyphase/FFT approach or Analytical approach. The third stage is used to separate signals with bandwidth ranging from 64 KHz to 10 Mhz. A frequency domain filtering approach using pipeline FFT architecture or a multistage tree approach may be used for this stage.

Obviously, there are many such variations possible for such a multistage hybrid channelizer. Table 1 summarizes the power and weight requirements for some of these schemes. For more details of such evaluation one may refer to [11]. In the table the first stage uses SAW filter except where digitization is indicated.

Conclusions

The paper has presented several channelization methods, which are applicable to a wideband satellite flexible architecture. Many multistage hybrid schemes are possible for implementation from which one needs to select an overall optimum and most flexible scheme. Weight and power requirements for some of these schemes have been presented for comparative evaluation.

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