Performance Options of a High Performance   
Receiver Filter Bank Channelizer

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*Abstract*-The M-path polyphase analysis filter bank channelizer is quite a remarkable digital signal processing algorithm. In its simplest realization, the maximally decimated filter bank, the bank outputs M baseband time series from translated spectral spans with bandwidth and sample rate fs/M from M spectral bands centered at integer multiples of fs/M. Modifications to the channelizer are many and include offsets of channelizer center frequencies, non-maximal decimation from M-to-1 to M/2-to-1 or to 3M/4-to-1 along with various post channelization signal conditioning options. We present a set of severe channelizer specifications and illustrate various channelizer modifications to reduce processing cost while meeting original design specifications.

***Keywords –polyphase filter bank; maximally decimated; non-maximally decimated; offset center frequencies; post channelizer signal conditioner; post channelizer interpolation;***

1. Introduction

In its most common incarnation, a polyphase down sampling channelizer simultaneously down converts and down samples M equally spaced, fixed bandwidth signals. Figure 1 shows the channelizer structure formed by an M-port commutator, an M-path partitioned low-pass prototype filter and an M-point inverse discrete Fourier transform (IDFT). For computational efficiency the IDFT is implemented with the IFFT algorithm.

In this configuration, the commutator delivers M consecutive samples to the M input ports of the M-path filter. Each port receives a data sequence sampled at fs/M with successive one-sample time delay offsets in successive paths. The sample rate reduction causes M-fold spectral aliases of the input spectrum, an effect easily observed in the frequency domain. The time series of each aliased band have an output sample rate of fS/M. In each arm, every spectral band centered at the M multiples of the output sample rate alias to the base band span centered at DC. The alias terms in each arm have distinct phase profiles due to their distinct center frequencies and the different delays of the sampled time series delivered to each commutator port. In particular, each of the aliased terms exhibits a phase shift equal to the product of its center frequency k with its path time delay rTS. These phase

Figure 1. Standard M-Path Polyphase Channelizer: M-Port Commutator, M-Path Polyphase Filter and M-Point IFFT.

 (1)

shifts are shown in Eq. (1) where fS is the sample rate at the input to the polyphase filter and TS, its reciprocal, is the time interval between input samples. The time delay response of each path filter aligns the time origin of their sampled data sequences formed at their outputs to a single common output time origin. This task is accomplished by the all-pass characteristics of the M-path partitioned filter that apply the required differential time delay to the individual input time series. Finally the IFFT block performs the equivalent of a beam-forming operation; the coherent summation of the time aligned signals at each output port with selected phase profile. Note that the channel spacing, the channel bandwidth, and the sample rate are all fs/M. This form of the channelizer is called a maximally decimated filter bank.

As a multi-channel channelizer in which we extract and separate adjacent channels the signal bandwidth must be less than the channel spacing. Under this this condition, there is spectral gap between the input channel bands. The gap is required for the channel filters to have a non-zero transition bandwidth between the channel bands. We will specify the signal bandwidths and channel spacing in the next section where we define the required filter characteristics. In a later section we will examine the option of increasing the transition bandwidth of the channelizer and follow the channelizer



Figure 2. Spectral Description of Multi-Channel Signal Presented to 30 Path Maximally Decimated Polyphase Filter Bank

with filters that form the desired narrow transition bandwidth. Because these filters operate at the reduced output sample rate, their reduced length and clock speed offer significant implementation advantages.

II Channelizer Specifications

Figure 2 presents an illustration of the multi-channel input spectra to be processed by the filter bank we discuss in this paper. A quick description of this signal set is that there are 24 bands spanning 576 MHz, sampled at 720 MHz. The band centers are symmetric about DC, with bandwidths slightly narrower than 24 MHz, and separated by 24 MHz centers. The required performance specification of the channel filter is that the 0.1 dB ripple bandwidth is 23.0 MHz and the -50 dB stopband bandwidth is 24 MHz. The number of paths in the polyphase filter and the IFFT size is determined by the ratio of input sample rate to output sample rate, a relationship shown in (1).

 (1)

The expression to determine the number of taps in a FIR filter is shown in (2) where fS is sample rate , Δf is transition bandwidth and K(A) is a parameter proportional to A, the out of band attenuation level. The estimate from (2) sets the filter length to 3273 taps. When designed by the FIRPM algorithm the estimate proved to be very good. We adjusted the filter length to be 1 less than the closest multiple of 30, a filter of length 3269 taps which met the design specifications. When partitioned into the 30 arms of the 30 path polyphase filter we find each arm contains 109 taps.

 (2)

We can implement the 30 path filters of the channelizer with each path operating at 1/30-th of the input rate 720 MHz/30 or 24 MHz, a comfortable speed for an FPGA implementation. We simulated the design in MATLAB and it worked well. See spectral responses of prototype filter in Figure 3. A minor problem is the filter centers of the channelizer are offset 12 MHz from the channel centers of the input signal. One solution to this problem is to use a complex heterodyne between the input signal and the channelizer to shift the input spectrum the 12 MHz offset between the input and output centers. This heterodyne, operating at the high input sample rate is shown in Figure 4.

Figure 3. Spectrum of 30-Channel Channelizer Filter with Zoom to Passband Ripple and Transition Bandwidth

This is where the lessons of this paper start to make contributions to workload reductions. Rather than apply the frequency shift phase rotations in the time domain at the high input rate we can slide them into the polyphase filter and apply them at the IFFT rate which is 1/30th of the input rate.



Figure 4. Complex Heterodyne Aligns Spectral Centers of input Signals with spectral centers of Channelizer Channels.

The rotator sequence is periodic in twice the length of the output vectors formed by the polyphase filter. Noting the sign change at the midpoint of the rotator vector we apply the rotators to the filter output in the same way the lower half of the butterfly of a radix-2 FFT forms its sum. We form the weighted sum of the even indexed data vectors and the weighted sum of the odd indexed data vectors, and apply the complex rotator weights to their difference. Alternate vector outputs have to be sign changed to keep the channel spectral bin center at DC rather than at the half sample rate. The modified form of the channelizer incorporating the half band width frequency offset is shown in Figure 5. Here we see the phase rotation corrections inserted between the polyphase filter output and the IFFT input.



Figure 5. Modified M-Path Polyphase Channelizer: M-Port Commutator, M-Path Polyphase Filter, Frequency Offset Rotators, and M-Point IFFT

III Non maximally Decimated Channelizer

We now consider a modification to the channelizer that holds promise of reduced workload but still meets the design requirements. We recognize that the workload in the channelizer is dominated by the large number of coefficients in the polyphase filter partition. As commented on earlier, in (2), this number is large because the ratio of sample rate to transition bandwidth is large. We can reduce the channelizer computational workload if we increase the transition bandwidth. This would reduce the filter length but would result in a filter that doesn’t meet the design requirements. Our response to this problem is we use a second filter applied to the channelizer output to form the reduced transition bandwidth at reduced cost because of its reduced sample rate. We have been presenting this approach in a number of recent papers from the perspective of *Green* technology.

If we increase the transition bandwidth of the channelizer filter we will also have to increase the output sample rate of the channelizer. We present the harris version of the Nyquist theorem. The Nyquist criterion tells us that the sample rate should exceed the two side bandwidth. The engineer’s question is, “By how much?” The harris version tells us by how much! As shown in (3) we should exceed the signal’s two sided BW by the transition BW of the anti-alias filters.

 (3)

The excess bandwidth typically increases the sample rate by 10% to 20%. In the modern era, we raise the sample rate to accommodate a significant increase in transition BW and then use a follow-up DSP filters to reduce the bandwidth and sample rate to the desired lower values.

When we run the M path polyphase filter bank at rates above fs/M the architecture changes and the channelizer is known as non-maximally decimated filter bank. We have a few options for the amount to raise the output sample rate. One common and easy to implement option is to double the sample rate from fs/M to 2 fs/M. For our particular example we raise the output sample rate from fs/30 or 720/30 or 24 MHz to 720/15 or 48 MHz by delivering 15 samples to the channelizer and form 30 output samples for every 15 input samples. We could have selected some other ratio which increased the sample rate by a smaller amount such as 720/20 or 36 MHz by delivering 20 samples to the channelizer and form 30 output samples for every 20 input samples. In the first case we would increase the sample rate by 100% and in the second we would increase the sample rate by 50%. We will consider the first option here and, following the precedence of many text books, leave the other options as an exercise for the reader. The option space for this choice is quite wide. Whichever choice we would finally select, we would be sure to include the internal frequency shift option in the resampled channelizer.

The common configuration of the M/2-to-1 resampling channelizer is shown in Figure 6. The output sample rate here would be 48 MHz. The details of this architecture can be found in many of the reference papers cited in this paper. Figure 7 shows the spectral response of the widest possible, alias free, transition bandwidth we could use at the 48 MHz output sample rate. This increase in transition BW from 0.5 MHz to 12 MHz would reduce the channelizer filter length by a factor of 48-to-1. The length would go from 109 samples per path to 3 sample per path after rounding up to the nearest integer. The problem with this transition BW filter is we can’t easily demonstrate the out-of-band spectral rejection of the channelized baseband filter. Thus for demonstration purpose we design the channelizer filter to satisfy the spectral response shown in Figure 8, a filter with transition BW of 6 MHz The 30-path channelizer filter will have 6 samples per path, still a significant reduction from 109 samples per path.



Figure 6. M-path, M/2-to-1 Down Sample Polyphase Analysis Filter Architecture

Figure 7. Frequency Response of Widest Transition BW Filter for 48 MHz Output Sample Rate Channelizer

Figure 8. Frequency Response of Narrower Transition BW Filter for 48 MHz Output Sample Rate Channelizer

We deigned the 30 path channelizer to meet the specifications indicated in Figure 8. The spectral responses of that design are shown in Figure 9. The first significant difference we see here, besides the wider transition BW, is the order of magnitude reduction of the in-band ripple. We designed the filter for a reduced level of in-band ripple because the channelized baseband series will be passed through a second filter which will add its in-ripple levels to the channelizer ripple. The next matter for us to explore is that follow-up house cleaning filter.

Figure 9. Spectrum of 30-Channel Channelizer Filter with Zoom to Passband Ripple and Wider Transition Bandwidth

IV Cascade House Cleaning Filters

We now have the task of designing the cascade filter that will reduce the transition bandwidth of channelized time series to the desired 0.5 MHz. It would be nice if that filter be configured to reduce the sample rate from 48 MHz to 24 MHz. The first filter option that comes to mind is a true half-band FIR filter. Access to this option is why we selected 48 MHz for the channelizer output sample rate. The qualifier, *true* for the half band filter is that we want the design to have zero value on alternate output samples. We can achieve that goal with a windowed sinc series or with the *half-band trick* that uses the FIRPM algorithm to design the filter’s odd index non-zero weights and inserting the even index zeros and center tap. The former design is characterized by non-uniform pass band and stop band ripple levels while the later has the equal ripple response of the standard FIRPM design.

We selected the half-band trick and designed the half band filter to operate at the 48 MHz sample rate with the desired 0.5 MHz transition BW and the 50 dB stopband levels. The filter length require to meet these requirements is 233 taps and the spectral characteristics of the filter is shown in Figure 10. Note the Half band filter has the same value in-band and out-of-band ripple levels. Consequently the in-band ripple is approximately 0.03 dB, the same deviation from unity gain that -50 dB is a deviation from 0. The combined ripple of the channelizer and its cascade filter we easily meet the 0.1 dB requirement. Figure 11 shows the block diagram of a 2-path, 2-to-1 down sample half band filter. The upper path contains the even indices of the low pass filter. These are the inserted zeros of the design process and thus has only 1-non-zero, trivial valued coefficient offset to the filter center sample. The bottom path contains 116 even symmetric filter coefficients. When the filter are folded to share the weights, the lower path has 58 multiplies. These 58 multiplies are performed every time 2 samples are delivered to the filter, so the filter workload is 29 multiplies per input sample. This work is performed at the 48 MHz clock rate which if referred back to the 15 times higher, 720 MHz input clock would be equivalent to approximately 2 multiplies per input sample. Of course we are performing this task 24 times, once for each output channel.



Figure 10. Spectrum of True Half Band FIR Filter with Zoom to Passband Ripple and Desired Transition Bandwidth

A second option for the half-band filter is a linear phase all-pass IIR filter. If you are not familiar with this class of filters, see Chapter 10 in the harris textbook on Multirate Signal Processing. We designed and simulated the IIR version of the half band filter which has a 2-to- down sampling option implementation very similar to that of Figure 11. This filter using a cascade of 1-st and 2-nd order all-pass polynomial in Z2 required



Figure 11. Block Diagram of Two Path 2-to-1 Down-Sample Half Band Filter

1-first order filter with one coefficient and 23-second order filters with 2-coefficients, for a total of 47-coefficients to implement the IIR version. Figure 12 shows the impulse response and spectral characteristics of this option. Till now we did not display the filter impulse responses because they have been uninteresting windowed sinc sequences. The impulse response of the linear phase IIR filter with its shorter causality delay is definitely interesting. The other interesting characteristic of this filter is the extremely low level of in-band ripple, slightly below 50 μdB. Like the FIR half band filter, the IIR version performs its 47 multiplies for every 2-input samples which results in less than 24 multiplies per input sample which gives this filter a slight lead relative to the FIR filter option.



Figure 12. Impulse Response and Spectrum of True Half Band IIR All-Pass Filter with Zoom to Passband Ripple and Desired Transition Bandwidth

A third option for the half band filter is the cascade of a pair of analysis and synthesis channelizers. The analysis synthesizer partition the input spectrum into a set of reduced sample rate baseband channels. The prototype filter in the analysis channelizer is Nyquist filter designed so adjacent channels cross at their -6 dB levels. The synthesis channelizer performs the perfect reconstruction of these baseband channels as they are aliased up to their original center frequencies by the M/2 up-sampling process. The bandwidth filtering option is performed by the binary mask between the analysis and synthesis banks. The stop band corresponds to the channels not participating in the assembly of the super channel formed by the channels passed to the synthesis process from the output of the analysis process. This architecture is shown in Figure 13. We select the number of paths in the channel for which the channel transition matches the super channel’s desired transition BW. We selected a 40 path system because the channel widths and spacing operating at 48 MHz is 48/40 or 1.2 MHz and the transition BW of the channelizer starts at 1/3 of the width or 0.4 MHz. We can adjust the transition BW with filter length and window main lobe width.

We designed the 40 path filter with 6-taps per path and synthesized the half band filter with 20 selected channels in an offset bin channelizer. The spectral characteristics are shown in Figure 14. The computational workload for the synthesized filter is 12 multiplies per input sample, 12 multiplies per output sample for the path filters and 10 multiplies per input sample for the two 40 point IFFTs. When the output 40 path filter is replaced with a 20-path filter to reduce the output sample rate the workload for the output channelizes is cut in half for a combined workload of 25 multiplies per input sample. Again this is operating at the reduced channel sample rate and we have to implement 24 versions of it. There are many other options to consider to determine the minimum workload version of this channelizer. We hope this paper helps widen your search horizons.



Figure 13 Cascade of Analysis and Synthesis Channelizers with Super Channel Passband formed by the Binary Mask between the Pair



Figure 14. Spectral Characteristics of Half Band Super Channel Formed by Cascade Analysis and Synthesis Channelizers

# V. Conclusions

In this paper we presented the problem of designing a hypothetical multichannel analysis channelizer to meet a severe set of specifications. By severe we mean operation at high sample rate fS with specifications that lead to very long filter lengths. Fortunately, the M-path polyphase channelizer performs M-to-1 down sampling to each of the M-paths. This means that each path operates at the reduced sample rate fS/M. This is a good start but we still may have objectionably long path filters, greater than 100 taps per path in our example. The lesson we have learned in the *Green* implementation arena, is that we should avoid implementing filters with a large ratio of sample rate to transition bandwidth. This leads us to solve the problem with two filters, one operating at the high input sample rate and one operating at the lower output sample rate. In this process, the first filter, with a wider transition bandwidth, reduces the bandwidth and sample rate. We offered two, of many possible sample rate reduction filter options and selected one for our example. We then discussed three options for the second filter that was designed to reduce transition bandwidth and further reduce output sample rate. The hypothesized example was formulated and a few options were examined to show the wealth of design options we can use to reduce computational workload. We hope this perspective and the process was useful to the reader.

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