Modeling PIRAQ IF signal processing for wind profilers and cloud radars.

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Based on discussions with Paul Johnston, Jim Jordan and Eric Loew, I have modeled the PIRAQ-based 60-MHz digital receiver using MATLAB/Simulink. This model is useful for designing optimal matched filters and investigating such phenomena as the image levels observed recently on the Darwin cloud radar.

Review

Let us first review the desired operation of the digital receiver for a 667-ns transmit pulse, corresponding to a 100-m range resolution. Fig. 1 shows the radar spectra at various stages of the signal processing. In Fig.1a, the real 60-MHz IF signal has a conjugate-symmetric spectrum. The negative frequency component is colored pink to distinguish it from the positive component shown in blue. We show a *sin(x)/x* spectrum typical of the rectangular transmit pulse and note that the nulls occur at integer multiples of 1.5 MHz. We should also remember that the actual transmit spectrum is a comb of frequencies spaced at the pulse repetition frequency that are enveloped by the *sin(x)/x* function. Before it is sampled, the IF signal filtered by the analog bandpass filter shown in green. This is an *anti-aliasing* filter required to attenuate signals that could alias into the passband. Fig. 1a also shows a DC component in black. Even if the IF signal has no DC component, a DC offset in the analog-to-digital converter circuitry may produce a DC term in the digitized signal.

The spectrum of the 48-Msps sampler is shown in Fig. 1b in red. The sampling replicates the IF spectrum at 48-MHz intervals. As seen in Fig. 1c, this has the effect of mixing the original 60-MHz IF spectrum, including the DC term, to 12 MHz. The figure also shows the Nyquist frequency at ± 24 MHz. The next step in the processing is to shift the positive part of the spectrum down to 0 Hz. This is accomplished by multiplying the signal by a complex exponential, the spectrum of which is the single delta function at -12 MHz, shown in Fig. 1d. The result of the downconversion to baseband is shown in Fig. 1e. (Note the change in frequency scale.)

At this point we can pass the signal through the *matched filter*, a non-optimal example of which is shown in green in Fig. 1f. The matched filter should be the complex conjugate of the transmit spectrum while simultaneously filtering out the DC component at -12 MHz and the negative (*image*) frequency components around Nyquist. Since we digitized at 48 Msps, 32 samples span the 667-ns pulse and this limits the number of taps in our matched filter to 32 (unless we are willing to tolerate some range smearing). After filtering, the signal is typically decimated to reduce the data rate. Fig. 1f shows the spectrum of a 32X decimation (48 MHz/32 = 1.5 MHz) in red. The decimation aliases the spectra of Fig. 1e after it is filtered. The resulting spectrum is shown in Fig. 1g. Note the change in frequency scale and new Nyquist frequency. We see that the spectrum contains the desired component in blue within the Nyquist interval, as well as aliased parts of the positive frequencies (blue) any residual components from the negative spectrum (pink)

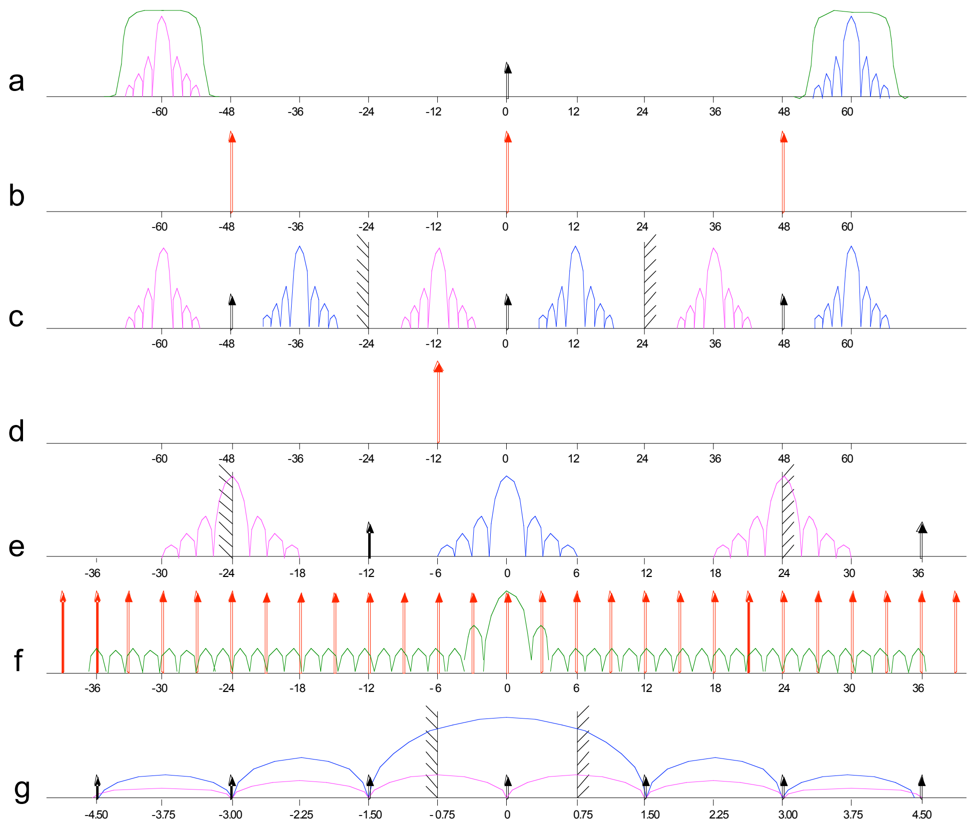

Figure 1. 60-MHz IF spectrum at various stages of processing.

and DC that have passed through the matched filter. This is a critical step since these aliased artifacts have now contaminated the spectrum. The decimation must be taken into account when designing the matched filter. The filter should approximate the conjugate of the transmit spectrum over the final Nyquist interval determined by the decimation. It should also provide maximum attenuation outside this band particularly at -12 MHz for the DC component and near 24 MHz to filter out the image.

The relative level of these undesirable signals is typically determined by the number of taps in the matched filter which, in turn, is determined by the original sample rate and the desired range resolution. For example, in the above case, if the original sample rate was 96 Msps, we could use a 64-tap filter which would allow for better matched filtering.

PIRAQ processing

The PIRAQ exploits the fact that the 48-MHz sampling frequency is four times the required 12 MHz downconversion frequency (see Fig. 1c). Therefore, the 12 MHz complex exponential, $e^{-j(2\pi n12/48)} = cos(n\pi/2) - j\ sin(np/2)$ of Fig. 1d reduces to a sequence of ones and zeros, {1 0 -1 0 …} for the real (in-phase, I) channel, and {0 -1 0 1 …} for the imaginary (quadrature, Q) channel. This is a well-known [1, 2] and clever technique for downconversion without the need for multipliers.

Suppose we wish to downconvert the 12-MHz signal and filter it with a 32-tap FIR matched filter as described above for 100-m range resolution. Fig. 2 shows how this might be done. The input sequence is multiplied by the cosine and –sine sequences and then fed into a 32-tap transversal FIR filter. The figure shows a snapshot of the machine state after 32 clock cycles. The first input sample, x_0 has worked its way to the end of the filter and is multiplied by filter coefficient h_{31}. At this point, the first fully-filtered sample is available at the output. The reader has undoubtedly discovered the inefficiencies of this configuration. Expensive hardware multipliers have been used to multiply the input sequence by ± 1 and 0. More multipliers are wasted multiplying terms in the product sequences that are zero. However, this model is useful as the baseline configuration.

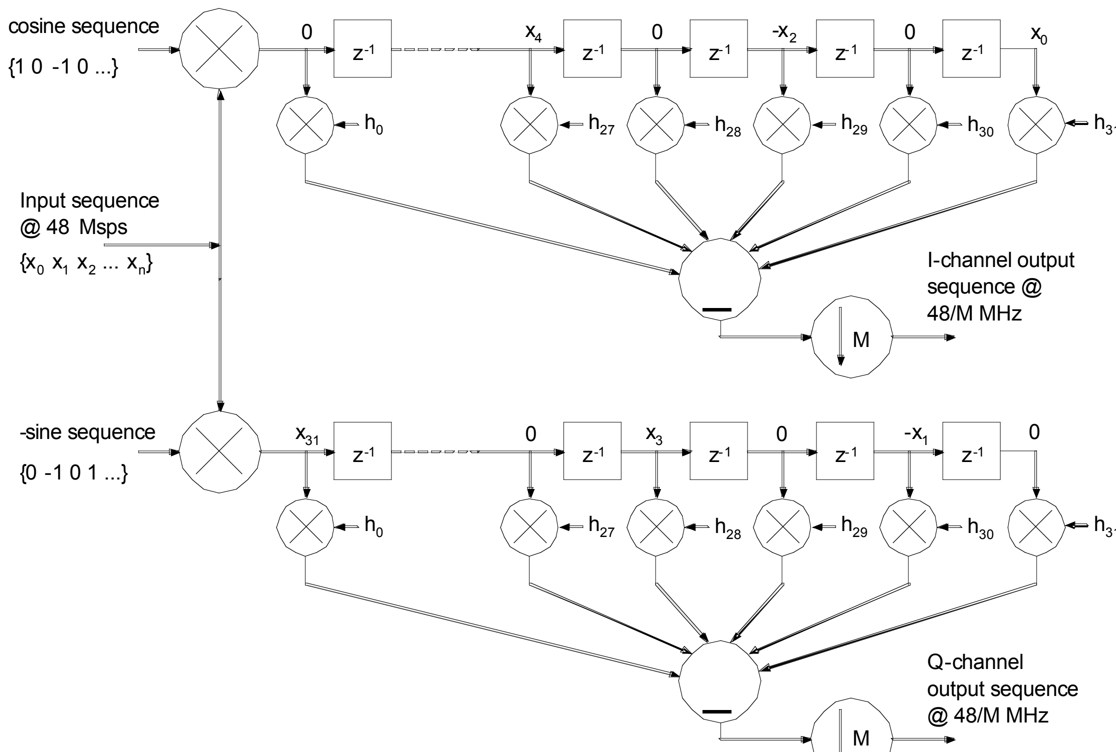

Figure 2. 12-MHz downconverter with 32-tap FIR filter and decimation by M.

Figure 3 shows an equivalent circuit that reduces the multipliers by more than half. The input sequence is sorted into the I- and Q-channels, with the first sample going to the I-channel. In each channel the sample is either passed through or negated (symbolized by the inverters, although two's-compliment inversion is a little more complicated). The resulting sequence is put into 16-tap FIR filters with the I-channel using the odd coefficients, and the Q-channel using the even coefficients of our original 32-tap filter of Fig. 2. Notice that these filters are running at a clock rate of 24 MHz and are out of phase with respect to each other. The outputs of the filters are decimated by M/2. The resulting complex output sequence is identical to the sequence generated in Fig. 2, except that the samples are coming out of the machine staggered.

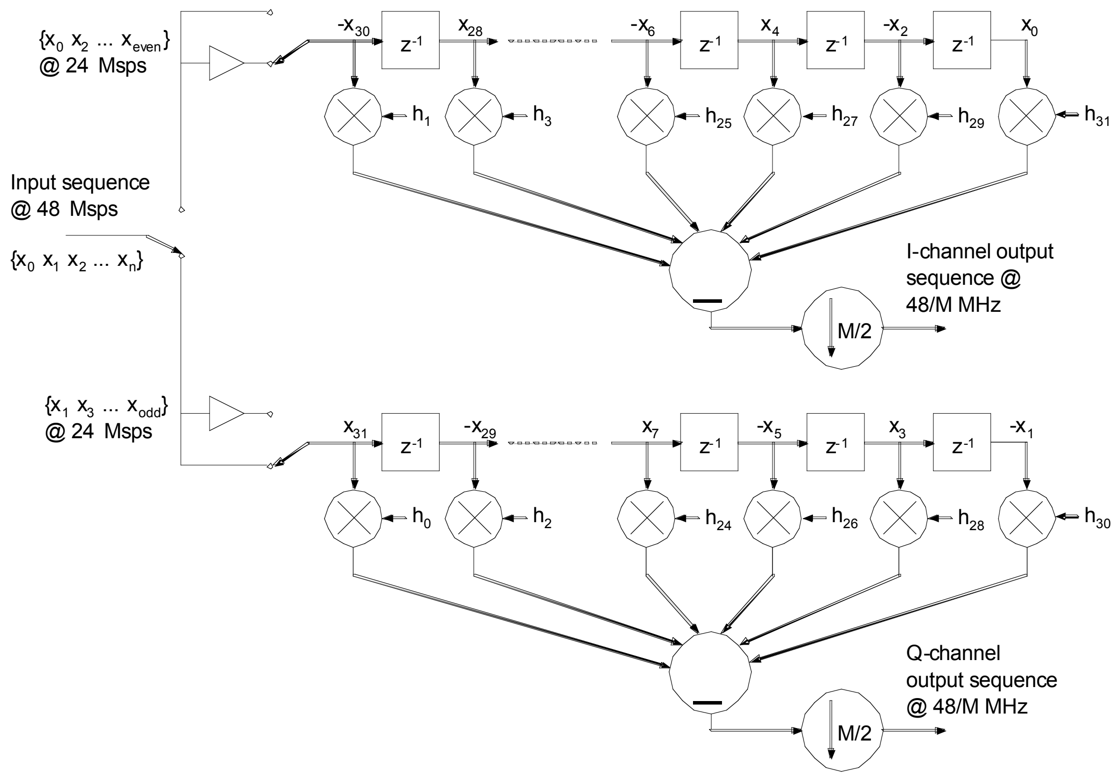

Figure 3. 12-MHz downconverter with 16-tap filters and M/2 decimation. Output is identical to Fig. 2.

There are two other techniques commonly used to reduce the number of multipliers in an FIR filter [2]. If the filter coefficients are symmetric then number of multipliers required to implement the filter may be reduced by a factor of two. In the case of a decimating FIR filter like those of Figures 2 and 3, we are not interested in the intermediate sum of products between the decimated output values. A decimating FIR may be used to reduce the number of multipliers by a factor proportional to the decimation factor.

The PIRAQ uses both of these circuit optimization techniques.

Figure 4 shows a simplified block diagram of one of two identical channels of the PIRAQ board. We show only those components germane to our present discussion. It contains a 14-bit ADC and two GC2011A digital filter chips from Graychip, Inc [3]. The chips multiply the incoming data word by ±1, then pass the data through a filter bank which may be configured in various ways that will be discussed.

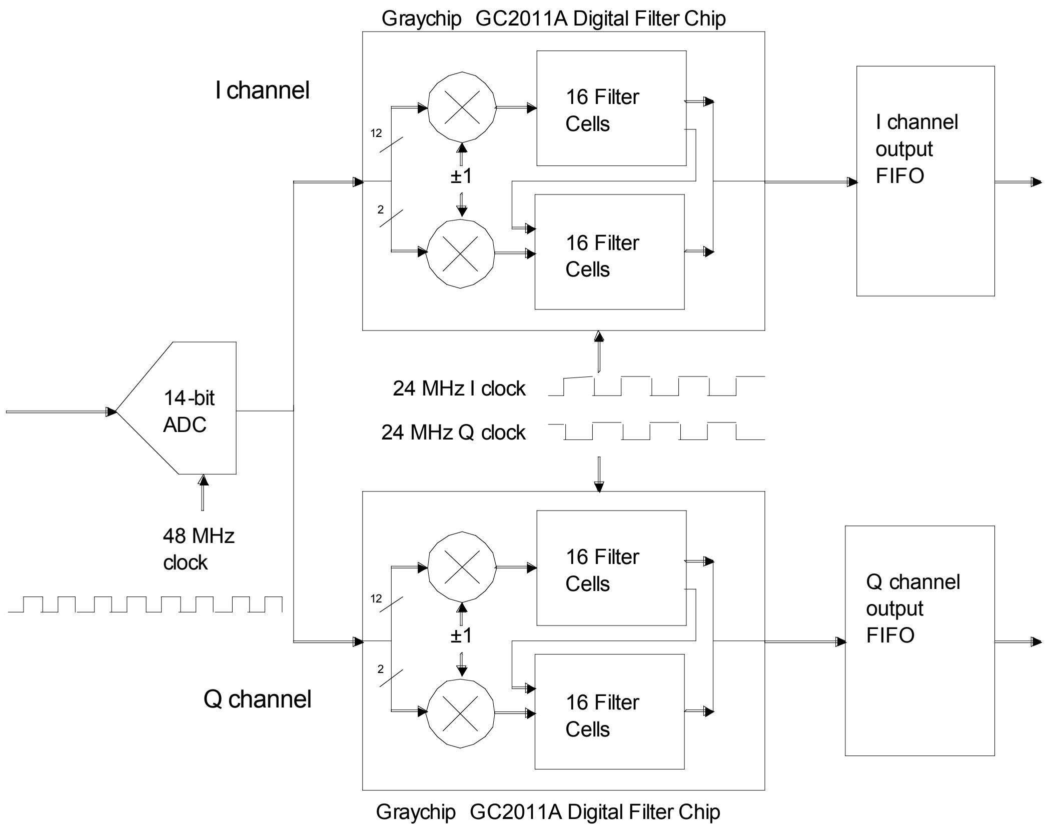

Figure 4. Simplified block diagram of one of two PIRAQ channels.

References

1. Marvin E. Frerking, “Digital Signal Processing in Communicsation Systems”
2. Richard G. Lyons, “Understanding Digital Signal Processing”
3. http://www.ti.com/graychip/GC2011/GC2011.html
4.