Efficient Simultaneous Detection and Isolation of Many Linearly Modulated Signals

Brian H. Hulette, Amir I. Zaghloul
{hulettbh, amirz}@vt.edu
Virginia Polytechnic Institute and State University
Dept. of Electrical Engineering

(ABSTRACT)

One common problem in Software-Defined Radio (SDR) systems is that of detecting and then isolating narrow signals of interest in wideband sampled data. This involves using some means to detect the frequency of one or more signals of interest and then tuning, filtering, and decimating them. This problem is particularly challenging due to the high sample rate of this wideband data, so efficient algorithms are very desirable. Solutions to this have applications in many areas, including both Cognitive Radio and Electronic Warfare.

A framework for simultaneously detecting and sub-band tuning multiple signals is presented, with a focus on finding an efficient means to combine the two algorithms. Detection is performed using the Spectral Correlation Density (SCD) function which exploits the cyclostationary property of digital signals [1]. These detections are then used to direct a configurable filter bank which will tune, filter, and decimate all of the detected signals. Two different configurable filter banks are examined: a polyphase analysis/synthesis filter bank [2] and an overlap-save filter bank [3].

MATLAB simulations of this framework using both filter banks have been created, and are used to evaluate their performance. Measures of average runtime for each filter bank are presented as a proxy for each algorithm's efficiency for various numbers of signals of interest.

Keywords: Software-Defined Radio, SDR, Cyclostationarity, Spectral Correlation Density, SCD, Detection, Filter Banks

Contents

List of Figures

1 Introduction

A common problem when designing a Software-Defined Radio (SDR) system is the problem of detecting and tuning multiple signals of interest. A typical SDR will sample at a high rate to acquire a large section of the frequency spectrum (referred to as a wideband acquisition throughout this paper). Then it will attempt to detect the frequencies within the acquired range that contain signals of interest, and then tune, filter, decimate, and demodulate those signals.

Many popular SDR applications perform this detection step with a “man in the loop”. A falling raster of the wideband data is presented and the man in the loop will select the frequency that he would like to process ([4] is a web-based implementation of this concept). A common example of this is to acquire a wideband slice of the HAM radio spectrum and then select different frequencies to demodulate new FM push-to-talk signals and listen to them.

This paper investigates a structure that automates this process - but rather than processing a single analog signal, this framework is intended to process an arbitrary number of digital signals at once. Detection is performed by a cyclostationary detector rather than by a man in the loop. This technique exploits the cyclostationary properties exhibited by most digital signals with a fixed symbol rate. Tuning, filtering, and decimating are performed by a configurable filter bank, so that many signals can be isolated simultaneously.

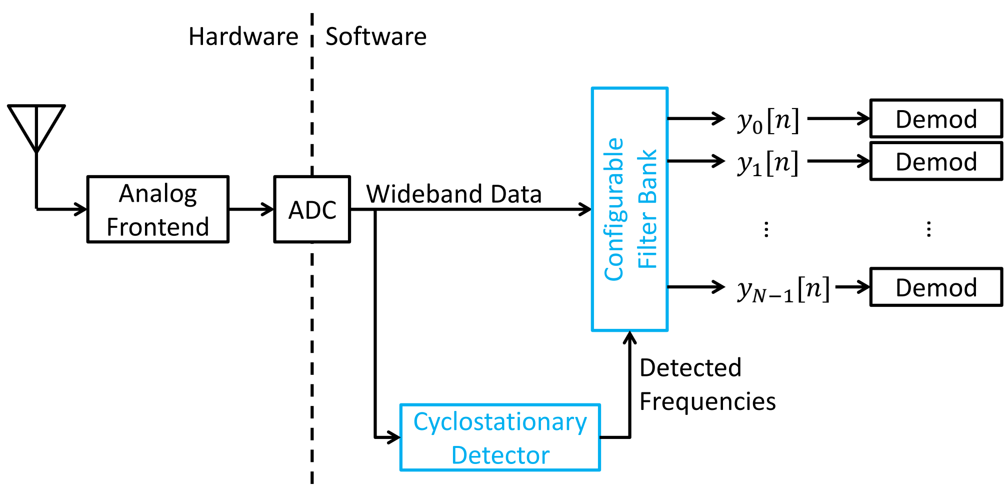


Figure 1: High-level block diagram of our system. An analog front-end and ADC are used to bring wideband data into our Software-Defined Radio. There, a detector directs a configurable filter bank to isolate N signals of interest, which are then demodulated. The highlighted detector and filter bank are the components examined in this paper.

The block diagram shown in Figure 9 illustrates this system. A front-end consisting of analog/RF components and an analog-to-digital converter (ADC) isolate and sample a large section of the frequency spectrum. This wideband sampled data is then brought into our SDR, where it is processed by a cyclostationary detector and a configurable filter bank. The frequencies identified by the detector are fed to the filter bank which will configure itself to isolate those signals. As the environment changes, with new signals of interest coming up and current signals going down, the detector will update the filter bank so it can re-configure itself as needed.

Each of the outputs of the filter bank are sent to a demodulator and/or various other follow-on processors to extract the signal internals in real-time. These demodulators must be allocated dynamically as the signal environment changes.

In this paper, we are primarily concerned with two components: the cyclostationary detector, and the configurable filter bank. Two different filter bank structures are examined: an overlap-save filter bank, and a combined polyphase analysis/synthesis filter bank, composed of a polyphase analysis filter bank followed by several polyphase synthesis filter banks. These structures are evaluated on two important criteria: 1) Their ability to accurately reproduce every detected signal, and 2) Their computational efficiency when combined with a cyclostationary detector.

Section 2 provides background information on cyclostationary properties and the SCD function upon which the detector relies, Sections 3 and 4 describe the overlap-save filter bank and the polyphase filter bank, respectively. Finally, Section 5 describes the simulation and its results.

2 Cyclostationary Detection

2.1 Cyclostationarity

Often signals are modeled as stochastic random processes with properties such as mean and variance, but many man-made signals, particularly digital communications, can be modeled with another statistical property called *cyclostationarity*, which implies the signal has some parameter which varies periodically with time. The frequency of this variation is called the cyclic frequency, α. [1].

Of particular interest for this project is second-order cyclostationarity, where a signal has a periodic auto-correlation, $R_{xx}(t, t+\tau)$:

$$R_{xx}(t, t+\tau) = E\{x(t)x^*(t+\tau)\} = \sum_{\alpha} R_{xx}^{\alpha}(\tau)e^{j2\pi\alpha t} \tag{1}$$

The function $R_{xx}^{\alpha}(\tau)$ is the cyclic auto-correlation function (CAF), given by:

$$R_{xx}^{\alpha}(\tau) = \lim_{T\to\infty} \frac{1}{T} \int_{-T/2}^{T/2} R_{xx}(t, t+\tau) e^{-j2\pi\alpha t} dt \tag{2}$$

The CAF is a function of two parameters, the cyclic frequency, α and the delay, τ. This function can be used for signal detection in the time domain. An estimate of the two-dimensional function versus α and τ is computed, then features within this plane are used for detection (e.g. [5, 6]). However, what is of interest for this project is detection in the frequency domain. This can be performed using the Fourier transform of the CAF, called the Spectral Correlation Density (SCD), given by:

$$S_{xx}(\alpha, f) = \int_{-\infty}^{\infty} R_{xx}^{\alpha}(\tau) e^{-j2\pi f\tau} d\tau \tag{3}$$

The SCD is a generalization of the conventional Power-Spectral Density (PSD), which it reduces to at $\alpha = 0$ [6]. Like the CAF, the SCD also contains unique features based on the modulation type, symbol rate, and carrier frequency of the transmitted signal(s), which we can exploit for detection. In [7], Gardner defines this SCD for various basic modulation types, including QPSK, which is used in the presented simulation.

2.2 Estimating the SCD

We can estimate a slice of the SCD at cyclic frequency α as the cross spectral density of the two frequency-shifted time sequences $x_L(t) = x(t)e^{j\pi\alpha t}$ and $x_U(t) = x(t)e^{-j\pi\alpha t}$ [1].

$$S_{xx}(\alpha, f) = X_L(f) X_U^*(f) \tag{4}$$

Note that $x_L(t)$ has been shifted down in frequency by $\alpha/2$, and $x_U(t)$ has been shifted up by the same amount. In order to estimate the SCD in this way we must perform two FFTs: one for both $X_L(f)$ and $X_U(f)$. However, it is possible to estimate both of these spectra using a single FFT, $X(f)$, in some cases:

$$\begin{aligned} S_{xx}(\alpha, f) &= X_L(f) X_U^*(f) && (5) \\ &= X(f-\alpha/2) X^*(f+\alpha/2) && (6) \end{aligned}$$

Using the second relation we can compute $S_{xx}(\alpha, f)$ using $X(f)$, by shifting that FFT in either direction by $\alpha/2$, and then conjugate multiplying the two shifted spectra together.

This is significant, since a simple forward FFT of the acquired data $X(f)$ is useful for other operations, as we will see in Section 3.2.

In order to perform this shift accurately using a single FFT we need to circular shift the FFT by an integer number of bins. This means we can only accurately estimate the SCD with this approach when $\alpha/2$ is a multiple of the FFT resolution. Thus α must satisfy the following relationship:

$$\alpha = \frac{2kf_s}{N_{fft}},\ k \in \mathbb{Z} \tag{7}$$

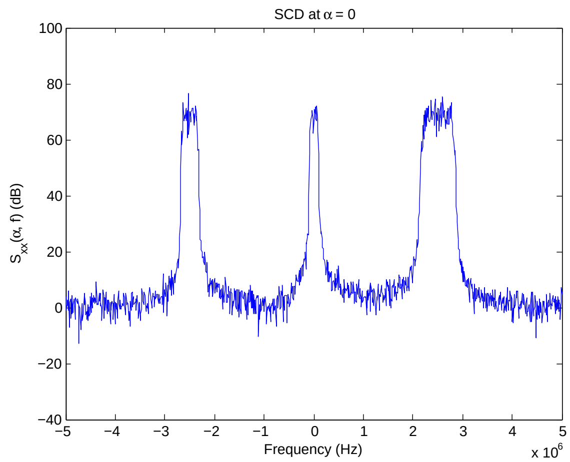
If an estimate of the SCD at a cyclic frequency that does not satisfy this relation is required, then we must take the original approach (perform the frequency shift in the time domain) to be perfectly accurate. Alternatively, we can still use a single forward FFT and approximate the shift by interpolating between bins, but this is only an approximation.

2.3 Using the SCD for Detection

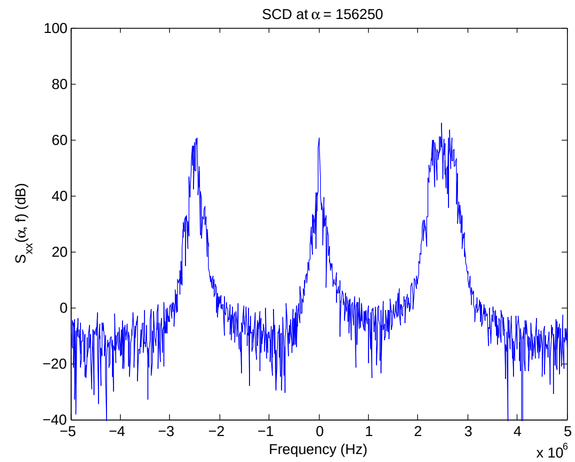
According to [7], QPSK signals generate a large peak in the SCD at the signal's center frequency when α is equal to the signal baud rate. So for our application SCD estimates are computed at cyclic frequencies corresponding to baud rates of interest. Figure 2 shows examples of these estimates for our test signal. These plots were generated using our MATLAB simulation, described in Section 5.

In each of these estimates it easy to detect the large, narrow peak at the center frequency of the signal with the corresponding baud rate. However, there are also other features, which agree with the theoretical SCD from [7]. The higher baud rate signals generate wide peaks in the lower cyclic frequency estimates, like the peak at 2.5 MHz in Figure 2b. These are easy to distinguish from the main peak with the human eye since they are much wider, but distinguishing them computationally is more challenging. The lower baud rate signals also generate features in the higher cyclic frequency estimates, in the form of two, much smaller peaks, straddling the actual center frequency. Examples of this are visible around 0 MHz in Figure 2c and 2d. These features are easily filtered out by either a human or a computer with a simple threshold.

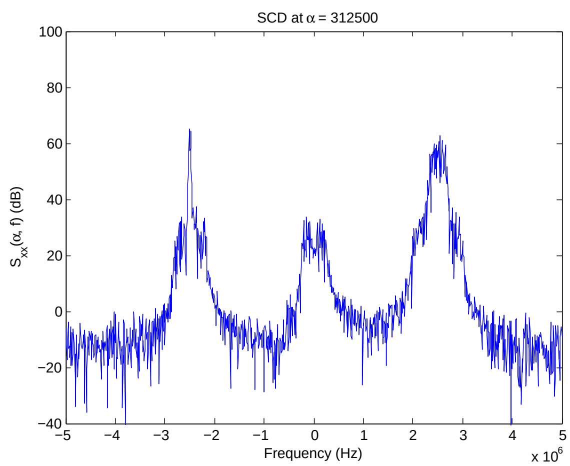
There are many potential approaches for identifying signals of interest in these SCD estimates (e.g. [8, 9]). However, in this paper we are primarily concerned with efficiently estimating the SCD, so we choose to use a relatively naïve approach: a thresholded peak detection that favors the higher cyclic frequency SCDs. In the example in Figure 2, favoring the higher cyclic frequencies allows us to correctly identify the 625 ksymbol/s signal at 2.5 MHz rather than mistakenly detecting the 156.25 or 312.5 ksymbol/s peaks.



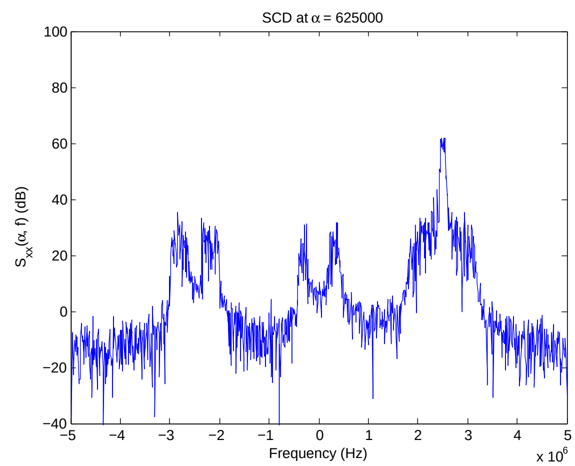
(a) SCD at $\alpha = 0$, Equivalent to a PSD



(b) SCD at $\alpha = 156.25$ kHz, baud rate of signal at 0 MHz



(c) SCD at $\alpha = 312.5$ kHz, baud rate of signal at -2.5 MHz



(d) SCD at $\alpha = 625$ kHz, baud rate of signal at 2.5 MHz

Figure 2: Estimates of the SCD at various cyclic frequencies for a simulated acquisition with signals at -2.5, 0, and 2.5 MHz.

3 Overlap-Save Filter Bank

The overlap-save filter bank structure used in this simulation is based on a description by Mark Borgerding from March 2006 [3]. Borgerding's concept is based on the well known Overlap-Save (OS) fast convolution technique.

OS fast convolution can be used to speed up convolution with a filter that has a long impulse response. The concept is that rather than convolving in the time-domain, $O(N^2)$, it is faster to first perform an FFT of both the signal and the filter and conjugate multiply in the frequency domain, then IFFT to go back to the time domain, $O(N \log_2 N)$. There is nothing new about this idea, but Borgerding's innovation is that he shows how to extend this concept to tune, filter and decimate any number of channels with arbitrary center frequencies and bandwidths.

We will now see how the OS filter bank performs tuning, filtering, and decimation. The same terms that Borgerding defines will be used in this description:

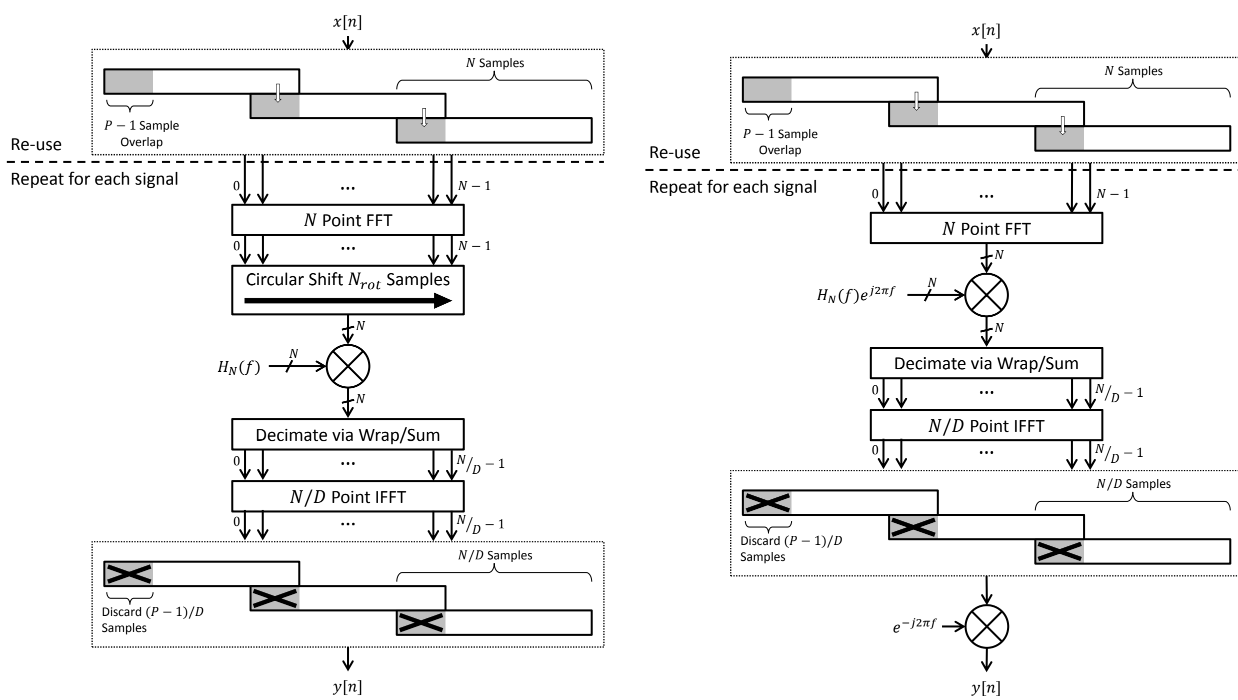
$x(n)$	Input data
$h(n)$	Baseband filter response
$y(n)$	Tuned, filtered and decimated output data
P	Length of $h(n)$
N	FFT size
D	Decimation factor
$V = N/(P-1)$	Ratio of FFT size to filter order

Tuning: If the frequency of the signal of interest corresponds exactly to one of the FFT bins then we can simply circular shift the FFT output to place that bin at the center. Then filtering can be performed at baseband. If all channels satisfy this criterion then we can re-use a single forward FFT for every channel, simply by circular shifting it to the appropriate bin in every case. This approach is shown in Figure 3a.

However, if this condition is not satisfied then we will need to perform the frequency shift in the time domain by multiplying by a complex rotator. Fortunately, there is a still a way to re-use the forward FFT in this case. After taking an FFT of the non frequency-shifted signal we can shift the baseband filter response to the desired center frequency, then after the filtered signal is returned to the time domain with an IFFT, we perform the frequency shift with a complex rotator. In addition to allowing us to re-use the forward FFT for each channel, this is more efficient than frequency shifting before the forward FFT since the multiplication is performed after decimation. This approach is shown in Figure 3b.

Filtering: As discussed earlier, this structure relies on OS fast convolution for filtering. The low-pass filter impulse response for each channel is Fourier transformed and then conjugate multiplied with the FFT of the input data after the signal of interest has been tuned to baseband. Alternatively, if tuning is being performed in the time domain, the filter impulse response must be tuned up to the signal frequency before being Fourier transformed.

Decimation: Following the frequency domain filtering we have two different options for decimation. The first option is to perform a full size inverse FFT of the output, and then decimate in the time domain. The issue with this approach is that we are computing a larger inverse FFT than we really need to. Borgerding suggests decimating in the frequency domain as a solution to this. The approach (called "Wrap/Sum" in Figure 3) is simple: coherently add together the components of the frequency spectrum that would be aliased upon time domain decimation. For example, if the FFT size is 1024 and we need to decimate by a factor of 4, then coherently sum FFT samples 0 to 255, 256 to 511, 512 to 767 and 768 to 1024. The result is a 256 sample frequency spectrum which we can now inverse FFT to produce the decimated output. The only trick with this approach is that we must discard just $(P-1)/D$ samples rather than $P-1$ to account for the overlap. This reveals one restriction of this structure: the filter order $P-1$ must be an integer multiple of the decimation, D.



(a) Performing frequency shift by circular shifting the FFT output

(b) Performing precise frequency shift in the time-domain

Figure 3: Overlap-save filter bank structures

3.1 Limitations

There are a few limitations of this structure that are worth mentioning. First, as mentioned earlier, the filter order $P-1$ must be an integer multiple of the decimation factor D. This problem is pretty easy to solve though - simply zero-pad the filter to achieve an appropriate length. Another limitation that is potentially challenging is that the FFT size, N, must be an integer multiple of the decimation rate, D. So decimation rates whose prime factors are larger than 2 or 5 (or others, depending on the FFT implementation) could lead to FFT inefficiency for this structure.

One more limitation arises when attempting to tune by rotating the FFT, as in Figure 3a. As previously mentioned, the precision of the frequency shift is limited by the resolution of the FFT. However, the precision is also limited further by V, the ratio of FFT size to filter order. This is because we must restrict mixing to the frequencies whose period completes in $L = N - (P-1)$ samples. Borgerding provides the following equation for computing the number of FFT bins to rotate to shift to frequency f ([3] Equation (1)):

$$N_{rot} = \text{round}\left(\frac{Nf}{Vf_s}\right)V \tag{8}$$

Where f_s is the wideband sample rate. This equation simply adjusts N_{rot} to the nearest multiple of V bins. It is worth noting that the solution presented for making $P-1$ an integer multiple of D is only making this problem worse by making V larger - but there's nothing to be done about that other than opting for a shorter filter.

The easiest solution to this issue is to simply use the time-domain tuning approach illustrated in Figure 3b. It is slightly more complicated since a new band-pass filter must be generated from a low-pass prototype for each signal, and a complex rotator is required at the output, but it can easily select arbitrary frequencies.

3.2 Advantages

The greatest benefit of the overlap-save filter bank for our application is the ease with which it can be combined with SCD Estimation for cyclostationary detection. Since the first step for both algorithms is a forward FFT of the wideband input, when designing a joint cyclostationary detector/overlap-save filter bank we can re-use the same forward FFT for both algorithms.

The only design challenge is finding a combination of FFT size, sample rate, and decimation factor that will work for both algorithms.

Additionally, averaging the overlapped FFT frames together may have an effect on the accuracy of the SCD estimation, but analyzing this effect is beyond the scope of this paper.

4 Polyphase Analysis/Synthesis Filter Bank

The second filter bank structure we will examine relies on the analysis and synthesis polyphase filter banks described by fred harris[10]. A polyphase analysis filter bank (AFB) can be used to break up a wideband signal into M distinct channels, each decimated by a factor of D from the wideband sample rate. While a polyphase synthesis filter bank (SFB) performs the reverse operation, combining M channels into one wideband signal. The channel frequencies used in both structures are given by the expression:

$$f_k = k\frac{f_s}{M} \tag{9}$$

where f_k is the center frequency of channel k and f_s is the input sample rate.

These two structures can be combined to create a very flexible filter bank that can isolate signals with arbitrary center frequencies and various bandwidths [2].

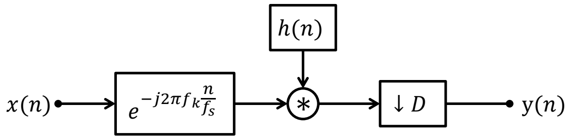
To introduce the concept of analysis and synthesis filter banks we will start with the maximally-decimated case, where the number of channels, M, is equal to the decimation factor D. In [10], harris shows how the polyphase AFB works by starting with a basic filter/decimator and making incremental modifications until it becomes an analysis filter bank as shown in Figure 5a. A brief summary of this explanation is provided here, which follows Figure 4.

We start with a single channel of a basic filter bank in Figure 4a. This channel consists of a multiplication with a complex rotator to tune the desired center frequency down to baseband, followed by a low-pass filter, represented by $h(n)$, and finally a decimation by D. The first modification we can make takes advantage of the Equivalency Theorem, which says that we can switch the order of the tuner and the filter, **if** the filter is changed to a bandpass filter centered at f_k. This modification is shown in Figure 4b. In this figure the tuner has also been moved after the decimator. Note that if the channel frequency, f_k, is an integer multiple of the output sample rate, $\frac{f_s}{D}$, then the tuner simplifies to $e^{j2\pi n} = 1$, so it can be dropped completely. Thus we will restrict the polyphase analysis filter bank to center frequencies which are integer multiples of the output sample rate.

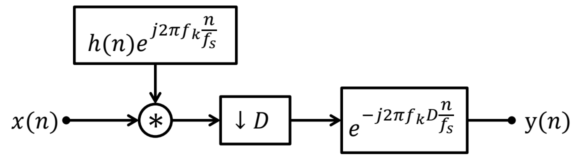
Next we can use the Noble Identity to swap the decimator and the filter, as shown in Figure 4c. In order to do so we need to define a set of partitions of the low-pass filter, $H_r(z)$, such that:

$$H(z) = H_0(z^D) + z^{-1}H_1(z^D) + \ldots + z^{-(D-1)}H_{D-1}(z^D) \tag{10}$$

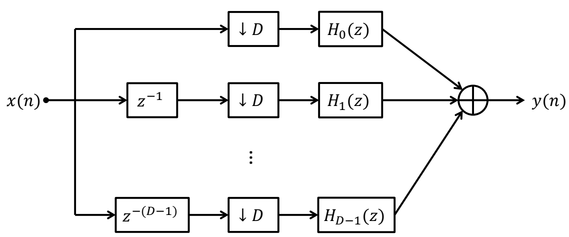
This means that each filter, $H_r(z)$, has has an impulse response which is $h(n)$ shifted by r samples and decimated by D. Using the Noble Identity in this way saves us from computing



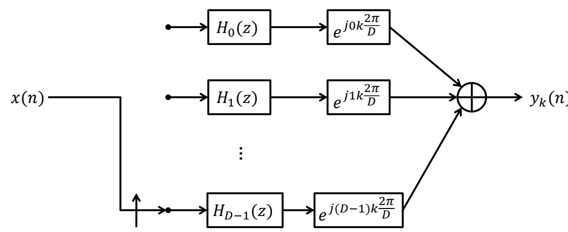
(a) Single channel of a simple tuning, filtering, and decimating filter bank. $h(n)$ is a baseband low-pass filter.



(b) Equivalency theorem allows the tuning step to be moved after the decimation. The baseband filter must be shifted up to f_k. If f_k is a multiple of the output sample rate, $\frac{f_s}{D}$, the tuning step can be dropped.



(c) The Noble Identity allows the filtering and decimation to be combined to remove excess computations.



(d) The series of delays and decimation steps can be replaced with an input commutator and complex rotators can be added after each filter to select channel k.

Figure 4: Creating a single channel of a polyphase analysis filter bank

samples that will just be dropped by the decimator. Note that in Figure 4c the tuner has been dropped as well, since we are restricting the filter bank to frequencies which are multiples of the output sample rate.

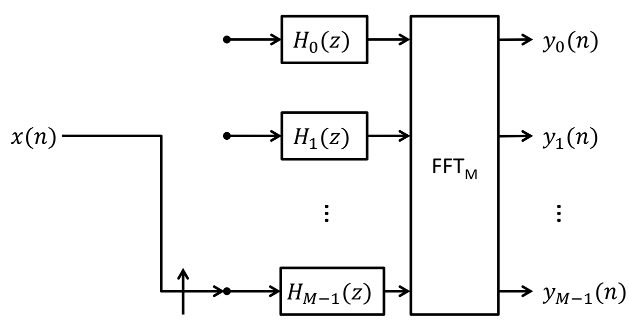
Finally, we complete the structure for a single channel of a polyphase analysis filter bank in Figure 4d. The combination of delays and decimators at the front of the previous structure is actually just a commutator. In [10] Harris explains this by thinking about the decimators as a switch that closes every D samples. So in Figure 4c, when all of the switches close the filter at the bottom gets the oldest sample, and each filter as you go up gets one newer sample until you get to the most recent sample on the top. The next sample is not processed until the switches close again, and it is passed to the filter at the bottom. With this explanation it is easy to see the whole structure can be replaced with a commutator.

The final structure has one more modification. The outputs are multiplied by complex rotators which select the individual channel, k, centered at f_k. A detailed explanation of this can be found in [10]. When all of the channels are combined to form the full filter bank these complex rotators all represent a DFT. The set of rotators that are multiplied and then summed together for channel k correspond to the kth output of a DFT:

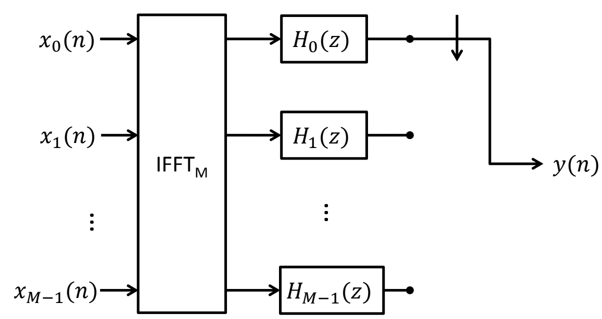
$$y_k(n) = \sum_{r=0}^{D-1} y_r(n) e^{j(2\pi/D)rk} \tag{11}$$

Where $y_r(n)$ represents the output of filter $H_r(z)$, and $y_k(n)$ is the output of channel k. Thus we can compute all D of the complex rotators with a D point FFT, as shown in Figure 5a.

Figure 5b shows the complementary structure: the polyphase synthesis filter bank. It is essentially a mirror image of the analysis case - an IFFT followed by filters and a commutator. It combines M channels at an equal sample rate, f_s, into a single wideband signal at sample rate Mf_s. The following section shows how these two structures can be used together to produce a very flexible filter bank.



(a) Analysis



(b) Synthesis

Figure 5: Full polyphase analysis/synthesis filter bank structures

4.1 Combining Analysis and Synthesis Filter Banks

The primary limitations of analysis filter banks alone are that, 1) they are limited to a single output sample rate, and 2) they are limited to specific channel center frequencies. These limitations are perfectly acceptable when every signal being processed is at the same bandwidth and symbol rate, and they are channelized with a known spacing, but in many SDR applications this is not the case.

However, there is a clever way to continue to take advantage of these efficient structures, while getting around the limitations: use analysis and synthesis filter banks together. The basic idea is to completely “deconstruct” the spectrum with a large analysis filter bank and then allocate synthesis filter banks for each target signal to recombine the sections of the spectrum which contain thos signals. This approach, described by fred harris in [2], can isolate multple signal bandwidths at arbitrary center frequencies.

Figure 6 illustrates the concept. The first step of this process uses an analysis filter bank to break up the wideband into small equal parts (1). Then, synthesis filter banks are used to re-combine the parts of the signal that correspond to signals of interest (2). Finally, complex rotators are used to center the signal (3).

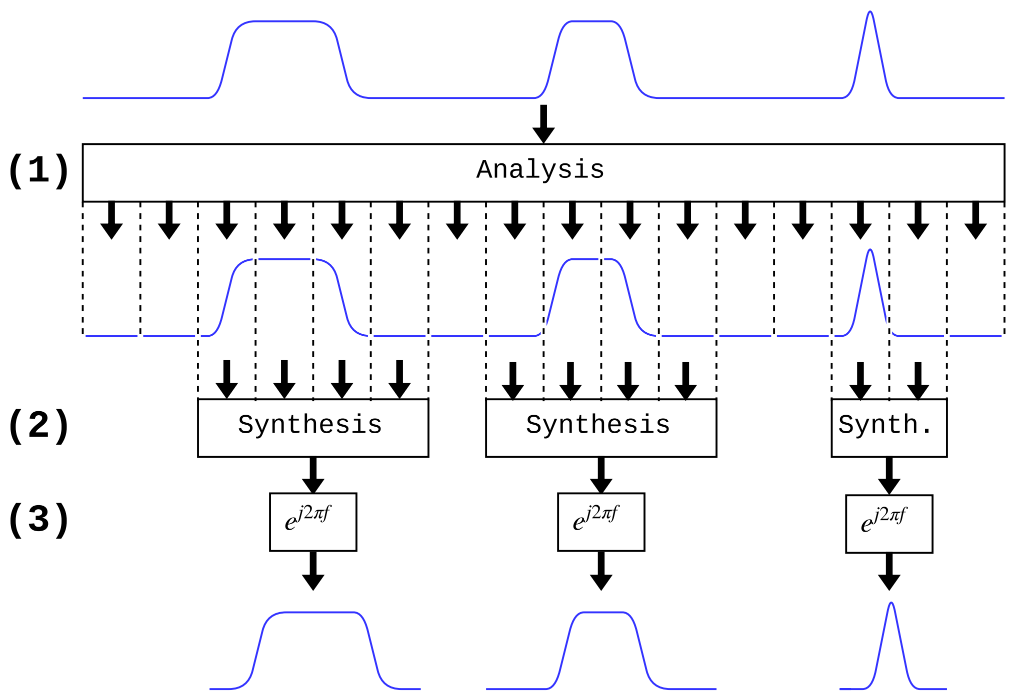


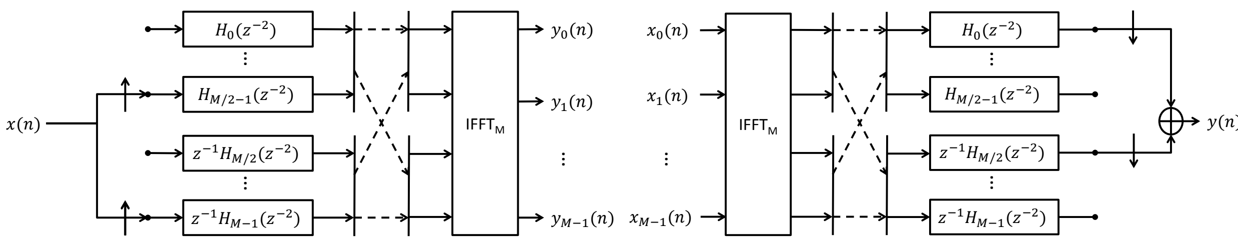
Figure 6: Using polyphase analysis/synthesis filter bank structures in concert. Step (1) is an analysis filter bank that breaks the signal into 16 channels, Step (2) is a set of synthesis filter banks for each signal of interest, and Step (3) is a complex rotators that corrects the frequency offset. Note that this figure depicts the signal in the frequency domain, simply for ease of illustration - all inputs and outputs are time-domain signals.

The analysis filter bank can be running all the time, while the synthesis filter banks and complex rotators in steps (2) and (3) can be dynamically allocated as signals of interest are detected (by some external detector).

Unfortunately, there is an issue with this approach, as we have described it so far. Since the analysis filter bank is maximally decimated, meaning the filter cutoff is equal to the nyquist frequency of the output signals, the filter rolloff aliases into the output. Then when these outputs are re-combined by synthesis filter banks to reconstruct a signal of interest, that aliasing is left in the middle of the signal bandwidth. For this reason, [2] utilizes a modified version of these structures: non-maximally decimated filter banks.

4.2 Non-Maximally Decimated Analysis/Synthesis Filter Banks

Non-maximally decimated analysis and synthesis filter banks are much like the analysis and synthesis filter banks we have already discussed, but modified to use a lower decimation factor. An excellent overview of these structures can be found in [11]. It starts with a general structure that can be used for arbitrary decimation rates (where D is a factor of M), and a realization of this general structure when $D = M/2$. This realization, which is used in our simulation, can be seen in Figure 7.



(a) Analysis (b) Synthesis

Figure 7: Non-maximally decimated polyphase analysis/synthesis filter bank structures

Clearly, these structures are very similar to the standard analysis and synthesis filter banks. Let's examine the differences in the analysis filter bank (Figure 7a), from left to right. The first difference is that the input commutator has been replaced with a double input commutator. This makes intuitive sense - with a double input commutator the structure will accept half as many input samples for each output sample compared to the standard analysis filter bank, leading to a decimation factor of $M/2$ rather than M. The second difference is the filter partitions: each partition has been downsampled by a factor of 2, and the filters on the bottom half have been delayed by one sample. Again, the downsampling makes intuitive sense given the change to $D = M/2$. The next difference is the conditional circular buffer after the filter partitions. This circular buffer simply swaps the upper and lower halves - but it only does it for every *odd* set of samples - the even sets are simply passed through. This swapping is required to absorb a phase offset. The final difference is the use of an M-point IFFT rather than an FFT [11].

Using these non-maximally decimated filter banks (NMDFBs) in the combined analysis/synthesis filter bank allows the synthesis filter banks to perfectly reconstruct the received signal.

4.3 Limitations

We have already examined the limitations of a polyphase analysis filter bank used alone: every channel must be at the same output sample rate, f_s/D, and they must have center frequencies which are multiples of that sample rate.

If the combination analysis/synthesis structure is used then this limitation can be avoided, however there are still other issues. First, the output sample rate must still be a decimation of the input sample rate - arbitrary rates are not allowed. Also, like the overlap-save filter bank, this structure relies on the efficiency of the FFT to save computation, and the size of the IFFT in the analysis filter bank is directly related to the decimation factor used. If a decimation factor with large prime factors is required then the analysis filter bank's IFFT will be inefficient.

Finally, there is no obvious way that this structure can be efficiently combined with SCD estimation for detection. Nothing about an polyphase filter bank can be re-used for the detection approach discussed in this paper.

4.4 Advantages

The primary advantage of a lone polyphase analysis filter bank is its simplicity and efficiency, but its limitations make it an untenable solution for this project.

The combination polyphase analysis/synthesis filter bank adds some complexity, but it is still quite efficient for the amount of flexibility that it provides. Another major benefit of both analysis and synthesis filter banks is that they can be implemented in hardware relatively easily. One can imagine a hardware implementation of this structure where an analysis filter bank is always running, and a bank of synthesis filter banks are dynamically connected and disconnected as signals are detected.

5 Simulation Results

A MATLAB simulation of the structures described in this paper has been created[1]. This simulation uses simple QPSK carriers as test signals, but the cyclostationary detection and configurable filter bank software are applicable to many types of signals.

5.1 Runtime Comparison

The simulation was used to compare the efficiency of the two filter bank structures. It is important to note that, while the runtime for both structures can clearly be improved with parallelization, this simulation is completely single-threaded. This approach allows us to use overall runtime measurements as an analog for the total CPU time required for each approach. Of course, a practical implementation of either structure should run as much in parallel as possible.

[1]The entire source code for this simulation is available on Github: https://github.com/TheNeuralBit/cyclo_channelizer/

A comparison of runtimes for the overlap-save structure and the polyphase structure can be seen in Figure 9. The plot shows the runtime when both structures are used to simultaneously detect and tune various numbers of signals. For each number of signals, a 10 MHz wide simulated wideband acquisiton is created. This acquisition contains 78.125 ksymbol/s QPSK signals distributed evenly throughout the spectrum. Each filter bank is configured to decimate every signal down to two samples per symbol.

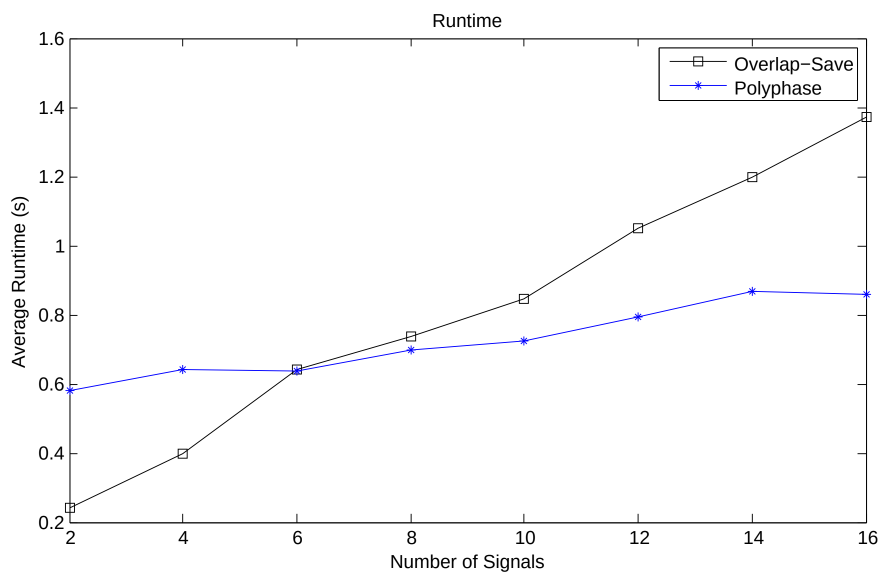


Figure 8: Runtime comparison of detector combined with the overlap-save filter bank and with the polyphase analysis/synthesis filter bank

From this runtime comparison, we see a clear trend: for a very small number of signals the polyphase structure is less efficient. However, each addiditional signal costs more time for the overlap-save filter bank to process than the polyphase filter bank. When isolating more than six signals, the polyphase structure is more efficient. There are two clear takeaways from this result: 1) there is a larger cost up-front when using a polyphase filter bank, but 2) the marginal cost for each additional signal being isolated is lower for the polyphase filter bank.

These takeaways are actually easy to predict just by examining the structures. The polyphase filter bank has a large up-front cost because it must run a large analysis channelizer no matter the number of signals being processed. In addition, the overlap-save filter bank saves some time up-front by using a single FFT for detection and filtering. On the other hand, for each additional signal, the polyphase fiter bank just needs to perform one additional small IFFT and a couple more filters at a low data rate, while the overlap-save filter bank must perform an additional multiplication in the frequency domain at the wideband data rate.

The situation simulated in this scenario represents a best-case scenario for both structures.

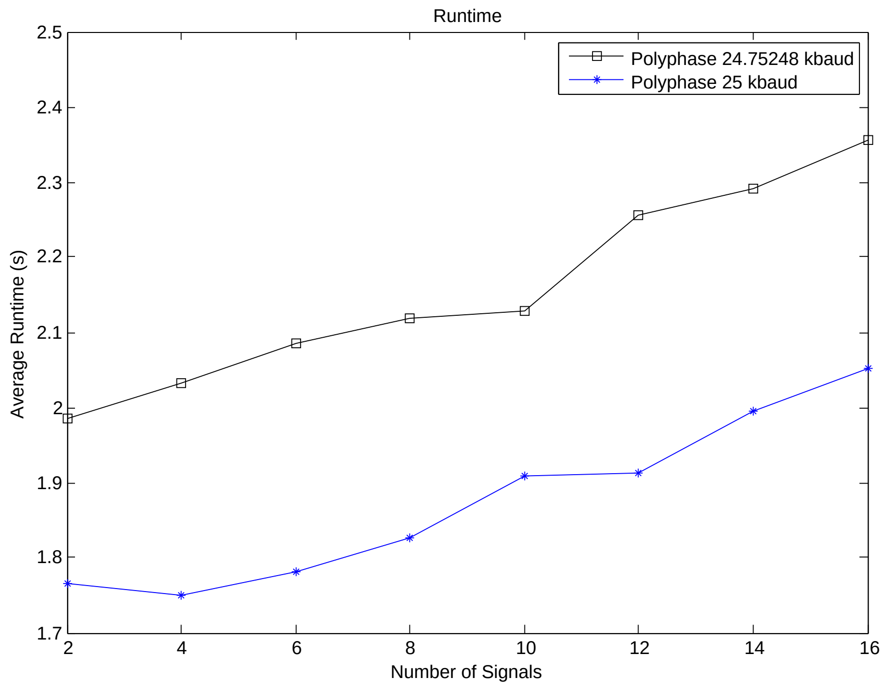


Figure 9: Runtime comparison of detector combined with the polyphase filter bank. The 24.75248 kbaud signal requires a 202 point IFFT, while the 25 kbaud signal can use a 200 point IFFT. The former has a large prime factor, 101, leading to a much longer runtime. A similar effect can be seen in the overlap-save structure.

The baud rate of the test signals was chosen so that the decimation factor would be a power of two. This means that polyphase analysis filter bank and the large FFT in the overlap-save filter bank can use a very efficient power of two FFT size. Figure 9 illustrates the effect of a poor decimation factor on the polyphase filter bank - just changing the decimation factor from 200 to 202 severely harms performance. It should be noted that, in general, it is easy to avoid this pitfall for both structures by decimating with a good FFT size. If a different narrowband sample rate is required, the outputs can be resampled after the filter bank.

In both of these test cases, since all of the signals are the same bandwidth, the polyphase filter bank can be designed so that only size $M = 2$ synthesis filter banks are required. One can imagine a worst-case scenario for the polyphase filter bank where there is one low bandwidth signal and many high bandwidth signals. In this scenario, a large analysis filter bank is required to isolate the low bandwidth signal, but then very large synthesis filter banks must be used to reconstruct each of the high bandwidth signals.

Clearly, it is difficult to say definitively that either structure is much more efficient than the other. For any given application, each structure should be evaluated to determine which is better suited for it.

6 Conclusion

Simulations of a cyclostationary detector and two separate configurable filter bank structures, an overlap-save filter bank and a combination polyphase analysis/synthesis filter bank, have been presented. Both filter bank structures were combined with the cyclostationary detector in order to evaluate their performance when detecting and isolating QPSK signals at various symbol rates in various wideband test signals. Both structures proved effective at accurately detecting and isolating these signals, but the runtime performance of each varied greatly depending on the number of signals being isolated. In general, the overlap-save filter bank is a better choice for a smaller number of signals, and the polyphase filter bank is better for a large number of signals.

Currently these simulations only uses the beginning of the test signal for detection, but in an actual implementation the detector would be running continuously and dynamically directing the filter banks to begin processing new signals. It should be noted that in this described system, both filter banks would need to have knowledge of the required signal bandwidths and output sample rates before processing begins, so that appropriate filters can be designed, and in the case of the polyphase analysis/synthesis filter bank, so that the size of the analysis filter bank can be chosen.

6.1 Future Work

This work could be extended in many different ways. In this project, the limitations of the various structures were taken as a given, however many of these limitations can actually be avoided. For example, Harris shows that it is in fact possible to adjust the output sample rate of the analysis filter bank by adjusting the input commutator [10]. That topic was left out of this simulation since it is difficult to devise a flexible version of it that will work for any sample rate. Future work could focus on what output sample rates are possible and devise an approach for a general purpose solution.

Another limitation taken as a given is that SCD estimation with a single FFT is impossible for cyclic frequencies that do not satisfy Equation 2.2 - but this could potentially be overcome as well. For cyclic frequencies, where $\alpha/2$ does not correspond to a specific FFT bin the FFT could be interpolated before shifting. This would not be perfectly accurate, but it would be worth evaluating how close of an approximation this yields. Additionally, one could evaluate how complex of an interpolation could be used while still maintaining the computational advantage of using a single FFT.

Finally, it would be useful to complement the simulated runtime analysis with an in-depth theoretical analysis of the computational complexity of both combined detector/filter banks.

References

[1] W. A. Gardner, "Exploitation of Spectral Redundancy in Cyclostationary Signals," *IEEE Signal Processing Magazine*, vol. 8, no. 2, pp. 14–36, Apr. 1991.

[2] F. J. Harris, E. Venosa, X. Chen, and B. D. Rao, "Polyphase analysis filter bank down-converts unequal channel bandwidths and arbitrary center frequencies," *Analog Integrated Circuits and Signal Processing*, vol. 71, no. 3, pp. 481–494, 2012.

[3] M. Borgerding, "Turning Overlap-Save into a Multiband Mixing, Downsampling Filter Bank," *IEEE Signal Processing Magazine*, vol. 23, no. 2, pp. 158–161, Mar. 2006.

[4] WebSDR. [Online]. Available: http://websdr.org

[5] J. Wang and T. Chen, "Cyclo-Period Estimation for Discrete-Time Cyclo-Stationary Signals," *IEEE Transactions on Signal Processing*, vol. 54, no. 1, pp. 83–94, Jan. 2006.

[6] M. Öner and F. Jondral, "Cyclostationarity Based Air Interface Recognition for Software Radio Systems," in *IEEE Radio and Wireless Conference*, Sep. 2004, pp. 263–266.

[7] W. A. Gardner, Brown, III, William, A., and C.-K. Chen, "Spectral Correlation of Modulated Signals: Part II - Digital Modulation," *IEEE Transactions on Communications*, vol. COM-35, no. 6, pp. 595–601, Jun. 1987.

[8] Q. Thai, S. Reisenfeld, S. Kandeepan, and G. M. Maggio, "Energy-Efficient Spectrum Sensing Using Cyclostationarity," in *Vehicular Technology Conference*, May 2011, pp. 1–5.

[9] D.-S. Yoo, J. Lim, and M.-H. Kang, "ATSC Digital Television Signal Detection with Spectral Correlation Density," *Journal of Communications and Networks*, vol. 16, no. 6, pp. 600–612, Dec. 2014.

[10] F. J. Harris, C. Dick, and M. Rice, "Digital Receivers and Transmitters Using Polyphase Filter Banks for Wireless Communications," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, no. 4, pp. 1395–1412, Apr. 2003.

[11] X. Chen, F. J. Harris, E. Venosa, and B. D. Rao, "Non-Maximally Decimated Analysis/Synthesis Filter Banks: applications in Wideband Digital Filtering," *IEEE Transactions on Signal Processing*, vol. 62, no. 4, pp. 852–867, Feb. 2014.