

# Evaluation of Polyphase Filter Architecture for pulse detection and measurement

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**Abstract:** This paper addresses the design of polyphase filter architecture for pulse detection and measurement as applied to the current EW (Electronic Warfare) systems. The filter is designed for a rectangular input of length 256. The comparison is made based on the simulation results.

## I. INTRODUCTION:

An EW system is used to protect military resources from enemy threats. The field of EW is recognized as having three components. They are:

1. Electronic Support Measure: collects the information on the electronic environment.
2. Electronic Countermeasures: jams or disturbs the enemy systems.
3. Electronic Counter countermeasures: protects the equipment against ECM.

Wideband and narrowband digital receivers are the two important types of systems used in EW and communications. However Wideband receivers have a very wide instantaneous input bandwidth of about 1GHz or larger. Another difference between the two is that wideband intercept receivers operate in an environment where the information of the input signal is unknown. Another major difference between the two is EW receivers output pulse descriptor words (PDWs), which describe the characteristics of the detected signal.

The basic block diagram of the digital wideband EW receiver is as shown figure:

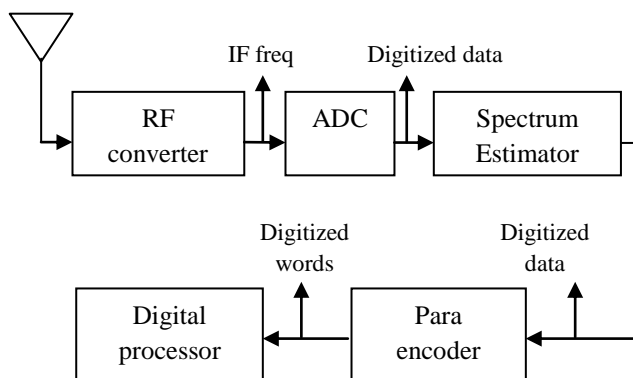


figure1 :digital wideband EW receiver

The antenna at the left captures pulsed RF signals that are generated by most radars. RF ranges from 2GHz-100GHz, but for radar, the most popular frequency range is from 2GHz-18GHz. Next an RF converter down converts the frequency signal into a lower intermediate frequency (IF) signal so that the EW receiver can more easily process the same bandwidth signal at a lower frequency. The signal then proliferates into the spectrum estimator. Spectrum estimation is commonly done using Fast Fourier Transform. The digital representation of the frequency spectrum is then output to the parameter encoder which generates PDWs.

The performance of an EW receiver mainly depends upon its update rate or its time resolution. Improved time resolution results in quicker and accurate TOA and pulse width measurements. For a digital EW receiver, the spectrum estimator is the limiting factor for the update rate.

The proposed method in this research for increasing time resolution is to use decimation in frequency domain. By decimating in the frequency domain, the computational load on the spectrum estimator is greatly reduced. This decimation in the frequency domain is implemented through channelized filtering structure.

Polyphase filtering is a multirate signal processing operation leads to an efficient filtering structure for hardware implementation. Polyphase filtering parallelizes the filtering operation there by reducing the computational load on the spectrum estimator.

The paper is organized as follows, first a brief introduction on the topic is given. Next the design procedure is discussed, followed by the algorithm and then based on the simulation results conclusions are drawn.

## II. DESIGN PROCEDURE:

The basic block diagram of the Polyphase FFT filter architecture is as shown in the figure:

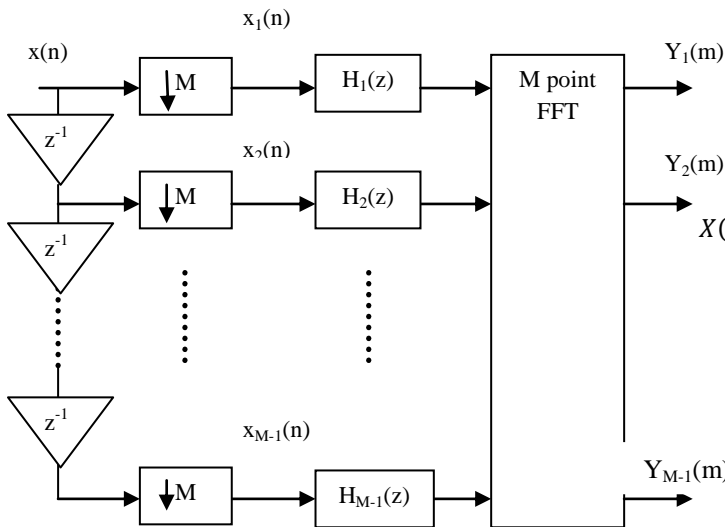


figure 2 : polyphase FFT architecture

The incoming N data samples are distributed into M branches and an M point FFT is performed. The number of points M is selected based on the time resolution required. The decimation results in gaps in the frequency domain. Hence each FFT filter must be widened to cover the gaps in the frequency domain. This is done by applying time domain window to the incoming data.

## III. Algorithm: Decimation in Frequency domain through polyphase filtering:

The frequency domain decimation operation is best viewed visually in the context of a filter bank. A filter bank is simply the filter response of each sub band in the spectrum overlapped on each other. Let us assume that one frequency component from the FFT output is equivalent to one filter output at a specific point in time from a filter bank. Let us assume that the 256 –point FFT is decimated by a factor of 8.

A 256 point FFT can be written as:

$$X(k) = \sum_{n=0}^{N-1} x(n) \cdot e^{-j2\pi nk/N} \quad (1)$$

where N=256. There are 256 outputs in the frequency domain. If every 8th output is kept and the other outputs are discarded, the resulting outputs are k=0,8,16,.....,248. There are a total of 32(256/8) outputs. These outputs can be written as:

$$X(0) = \sum_{n=0}^{255} x(n)$$

$$X(8) = \sum_{n=0}^{255} x(n) \cdot e^{-j2\pi 8n/256}$$

$$X(16) = \sum_{n=0}^{255} x(n) \cdot e^{-j2\pi 16n/256}$$

.....

$$X(248) = \sum_{n=0}^{255} x(n) \cdot e^{-j2\pi 248n/256} \quad (2)$$

Let us arbitrarily choose a frequency component k=16 and rewrite it by decomposing it into its polyphase components. The result is:

$$\begin{aligned} X(16) &= \sum_{n=0}^{255} x(n) \cdot e^{-j2\pi 16n/256} = \sum_{n=0}^{255} x(n) \cdot e^{-j2\pi 2n/32} \\ &= [x(0) + x(32) + x(64) + \dots + x(224)] \end{aligned}$$

$$+ [x(1) + x(33) + x(65) + \dots + x(225)] \cdot e^{-j2\pi 2/32} +$$

$$[x(2) + x(34) + x(66) + \dots + x(226)] \cdot e^{-j2\pi 2 \cdot 2/32} +$$

.....

$$+ [x(31) + x(63) + x(95) + \dots + x(225)] \cdot e^{-j2\pi 2 \cdot 31/32} \quad (3)$$

In the above equation the relation  $e^{-j2\pi mn} = 1$  where n is an integer. Now let us define a new quantity y(n) as:

$$\begin{aligned} y(n) &= x(n) + x(n+32) + x(n+64) + \dots + x(n+224) \\ &= \sum_{m=0}^7 x(n + 32m). \end{aligned} \quad (4)$$

where n=0 to 31. This y(n) represents the values in the bracket of the above equation. Each y(n) contains a total of 8 data points.

Using these y(n) values, the FFT results from (3) can be rewritten as:

$$X(0) = \sum_{n=0}^{31} y(n)$$

$$X(8) = \sum_{n=0}^{31} y(n) \cdot e^{-j2\pi n/32}$$

$$X(16) = \sum_{n=0}^{32} y(n). e^{-\frac{j2\pi 2n}{32}}$$

.....

$$X(248) = \sum_{n=0}^{31} y(n). e^{-\frac{j2\pi 31n}{32}} \quad (5)$$

All these equations can be written into one equation as:

$$X(8k) = \sum_{n=0}^{31} y(n). e^{-\frac{j2\pi kn}{32}} \quad (6)$$

where  $k=0,1,2,\dots,31$  and  $n=0,1,2,\dots,31$ .

The output  $X(8k)$  can be relabeled as  $Y(k)$ , thus the above equation can be written as:

$$Y(k) = \sum_{n=0}^{31} y(n). e^{-\frac{j2\pi kn}{32}} \quad (7)$$

This equation represents a 32 point FFT. In order to obtain the outputs of a 256 – point FFT decimated by 8, a 32 point FFT can achieve the goal. Thus the design of the FFT can be simplified. The input must be manipulated, however in order to obtain the desired result.

A general statement without further proof will be presented here. If one wants to perform an  $N$  point FFT and the outputs in the frequency domain are decimated by  $M$ , one can achieve the goal by performing an  $N/M$  – point FFT. A new input format  $y(n)$  must be built first. The generalization of the  $y(n)$  can be written as:

$$y(n) = \sum_{m=0}^{M-1} x(n + mN/M). \quad (8)$$

where  $n=0,1,2,\dots,(N/M)-1$ . The outputs in the frequency domain can be obtained as:

$$X(n) = \sum_{n=0}^{\left(\frac{N}{M}\right)-1} y(n). e^{-\frac{j2\pi kn}{N/M}} \quad (9)$$

Hence the design of the FFT can be simplified tremendously.

If the filter response is plotted for the (7), the resulting filter response will have very low side lobe level. The resulting filter bank has many gaps in the frequency domain. As a result each filter response must be widened. This is accomplished by applying a window function to the incoming data.

The window function that is used here is a Parks McClellan window function.

The input data  $x(n)$  will be modified by the window function  $h(n)$ . Here,  $h(n)$  instead of  $w(n)$  is used for the window function because  $h(n)$  will be used to represent the impulse function of the filter. The resulting data  $x_m(n)$  used as the input of the FFT can be written as:

$$x_m(n) = x(n).h(n) \quad (10)$$

where  $n=0,1,2,\dots,255$ . As stated previously the outputs are decimated by 8. Under this condition the modified data can be used in (4) to find the  $y(n)$  as:

$$y(n) = \sum_{m=0}^7 x_m(n + 32m) = \sum_{m=0}^7 x(n + 32m). h(n + 32m) \quad (11)$$

where  $n=0,1,2,\dots,31$ . A 32 point FFT is performed on these  $y(n)$  values, and the individual filter response as well as the filter bank is plotted to observe the improved filter shape.

#### IV. SIMULATION RESULTS:

The simulations are done by means of Matlab. A rectangular window was used as the input. As stated earlier the sample size was considered to be  $N=256$  and the decimation factor as  $M=8$ . The sampling frequency was assumed to be 1350MHz. The filter designed to widen each filter output was generated using FDATool Matlab for the following parameters:

$$F_s = 1350 \text{ MHz};$$

$$F_p = F_s * (M/N) = 42.1875 \text{ MHz};$$

$$F_s = 2 * F_p = 84.375 \text{ MHz};$$

$$w_p = 1/(N/M/2) = 1/64;$$

$$w_s = 2 * w_p = 1/32;$$

$$\text{density factor} = 20;$$

$$N = 256;$$

1. For the FFT of a rectangular window described by the equation 1, the individual filter response and the filter bank with overlapping and non-overlapping are as shown below respectively:

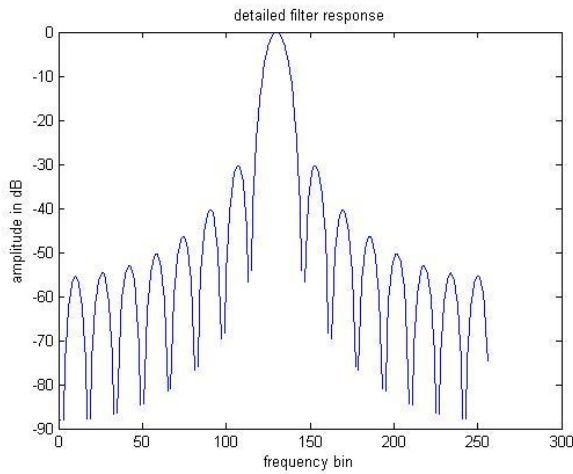


figure 3: 256 FFT individual filter response

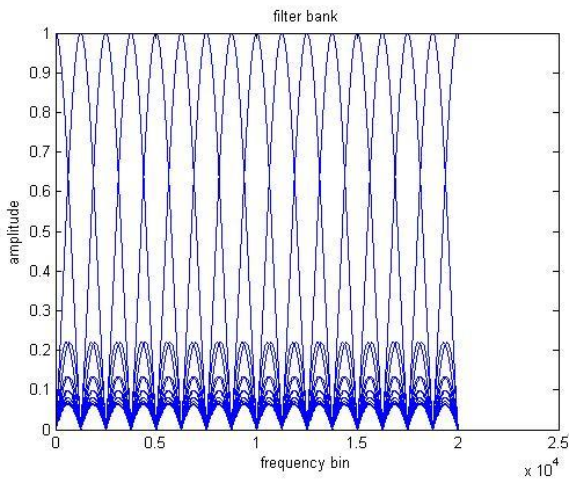


figure 4: 256 pt FFT overlapped filter bank

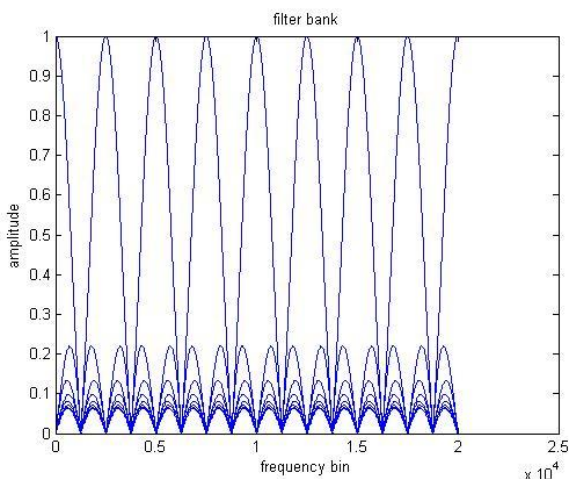


figure 5 : 256 pt, FFT non-overlapping filter bank.

It can be seen that the sidelobe levels are very high also the transition band is not sharp. As a result this type of a response is undesirable.

2. After applying the decimation in frequency domain algorithm to the same input and performing a 32 point FFT on the output values and plotting the individual filter response and its filter bank respectively:

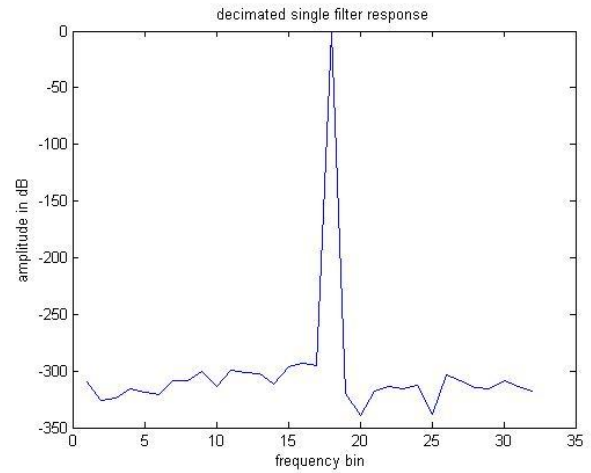


figure 6: 32 pt decimated FFT individual filter response.

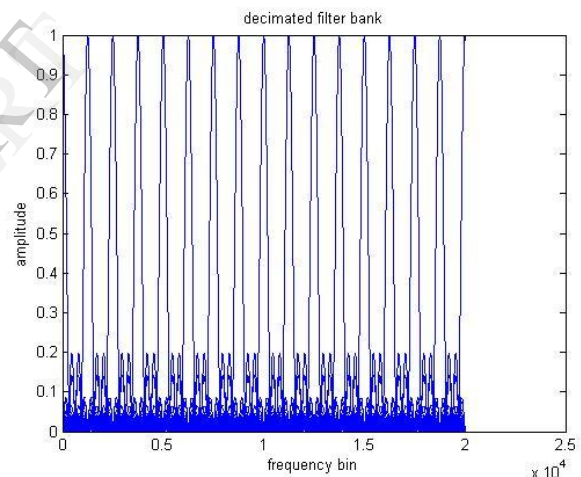


figure 7: 32 pt. decimated FFT filter bank

The filter bank as seen in the figure has many gaps resulting from the decimation operation. If an input signal frequency falls within one of the gaps, the receiver will not detect the signal. As a result this filter shape is not acceptable, each filter must be widened. This is accomplished by applying a window function to the incoming signal.

3. In order to widen each filter while suppressing the side lobes , a windowing function is applied to the incoming data. The magnitude and the impulse response of the filter that is used is as shown respectively below:

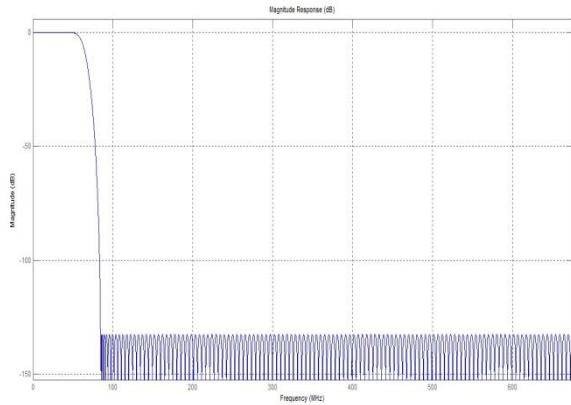


figure 7: magnitude response of the lowpass filter

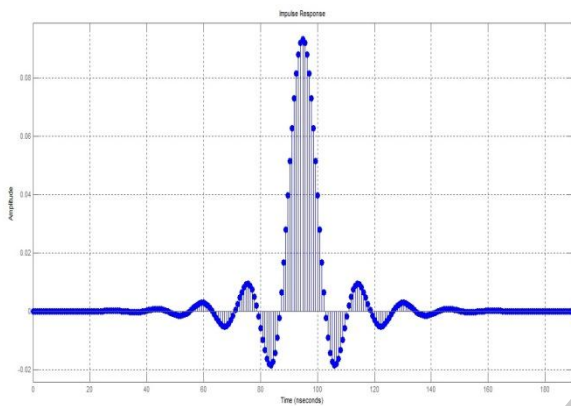


figure 8 : impulse response of the lowpass filter

If the 32 pt. FFT is performed on the resultant  $y(n)$  values, the FFT will modulate the prototype lowpass filter co-efficients and the input data to generate a filter bank of 16 unique filters.. The resulting individual filter response and the filter bank is as shown respectively:

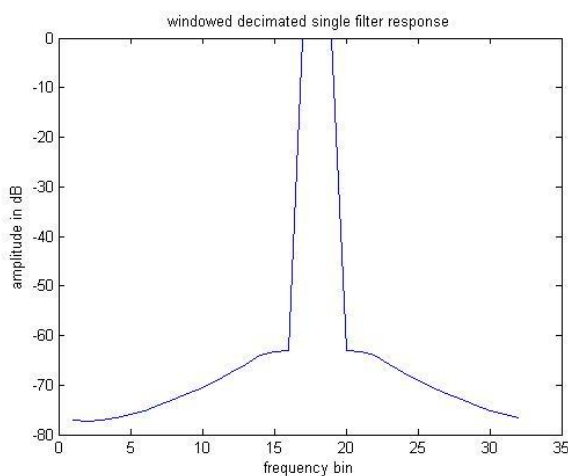


figure 9 : 32 pt. windowed decimated individual FFT filter response.

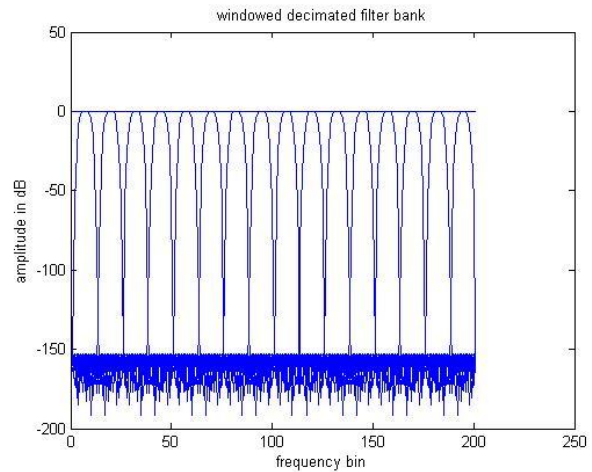


figure 10 : 32 pt. windowed decimated FFT filter bank.

## V. Conclusion:

A significant advantage of the decimation in frequency domain is the improvement in the time resolution, which improves the TOA and pulse width measurements.

For the original 256 point FFT with the sampling frequency of 1350MHz the time resolution is 189ns. For the decimated case with 32 point FFT the time resolution is 23ns. There is an eight fold improvement in the time resolution.

For the windowed filter bank, there are no missed frequencies, but its ability to resolve two frequencies that are close together is limited. For this reason the decimation in frequency domain has a direct effect on frequency resolution.

## VI. REFERENCES:

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