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A generic analytical formula for range side-lobes cancellation filters in pulse compression phase coded waveforms

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Abstract. Pulse Compression (PC) is applied in radar systems to enhance range resolution and simultaneously preserve a high duty cycle. PC phase coded radars suffer from unwanted side-lobes at the Matched Filter (MF) output which may mask weak nearby signals. Different methods have been applied to deal with this problem. Each method is designed and applied particularly for certain phase coded signal. In this paper, a generic mathematical formula for a Side-Lobe Cancelling (SLC) Optimum Filter (OP-F) in phase coded pulse compression waveforms is presented. The derived filter followed the MF and can be applied to any type of phase coded signals with any length. The validity of the derived formula is verified for Barker, polyBarker and polyphase coded signals. The enhanced detection performance of the OP-F over MF alone is verified through receiver detection curves. Also, Implementation issues of the OP-F in either time and frequency domain are discussed.

1. Introduction

PC is an imperative module in modern radar systems. It is used to solve the main problem of a radar system that requires a long pulse to achieve large radiated energy but concurrently a short pulse for good range resolution [1]. The output of the receiver Matched Filter (MF) is the autocorrelation of the transmitted signal. Low autocorrelation side-lobes are required to prevent masking of weak targets that are located in the range side-lobe of strong target [1]. Different side-lobe reduction techniques may be used according to the transmitted waveform type. Mainly, there are two configurations of side-lobe reduction techniques: the first one is to design a mismatched filter directly realizing pulse compression and side-lobes suppression simultaneously; the other one is to add a side-lobes reduction filter follows the matched filter. In this paper, a generic mathematical formula of an Optimum Filter (OP-F) transfer function, which is applicable to any type of phase coded pulse compression waveforms, for side-lobe cancellation follows the matched filter is considered. Different types of waveforms including binary phase coded, Barker code, [2], polyphase P4 code [3] and poly Barker code [4] are used to verify the universality of the filter transfer function.

The remainder of this paper is organized as follows: after introduction, an overview of different phase coded waveforms and its peak side-lobe level (PSL) of Auto Correlation Function (ACF) is presented in Section II. Section III demonstrates traditional side-lobe reduction techniques for different phase coded waveforms. Details and mathematical derivation of a generic formula for the proposed filter are presented in section IV. Simulation results are presented in Section V. Finally, in Section VI, conclusions and future works are presented.

2. Phase coded pulse compression waveforms

Phase coded waveform is used in pulse compression. It divides a long pulse of duration (T) into (N) sub-pulses of identical duration ($\tau = T/N$) and each one is coded with a different phase value ranges



from $(0\text{-to-}2\pi)$ [4]. A good pulse compression code should have an ACF with high Peak-to- Side-Lobe Ratio (PSLR) and a narrow main lobe.

2.1. Barker codes

Barker codes are the most famous binary phase-coded sequence which possesses equal side-lobes after passes through the matched filter. It is designed as the sets of N binary phases achieving PSLR of N . However, the longest code length is limited to $N=13$ [2]. For Barker code of length equal 13, the maximum achievable PSLR at MF output is equal to 22.28 (dB). For binary phase coded signals with code length longer than 13, an exhaustive computer search for minimum peak side-lobe (MPS) codes are presented in [5, 6] up to length ($M=105$).

2.2. Polyphase Barker code

Polyphase Barker codes or sometimes called generalized Barker sequence, are not binary codes. It has no restriction on the phase sequence values and also has minimal PSLR excluding the outermost side-lobe which is always equal to 1. Examples of such polyphase Barker sequences are presented in [7]. The PSLR of ACF for 13-element length poly Barker code is equal to 25.13 (dB).

2.3. Polyphase codes

Polyphase codes are derived from the phase history of Frequency Modulated (FM) pulses [4]. These codes are characterized by low PSL, good Doppler tolerance and compatibility with band pass limited receivers. Frank proposed a polyphase code with good non-periodic correlation properties and named the code as Frank code. Kretschmer and Lewis proposed different variants of Frank polyphase codes called p-codes which are more tolerant than Frank codes to receiver band limiting prior to pulse compression [8].

Frank, P1 and P2 codes have been derived from step approximation to linear frequency modulation (LFM) waveforms, while P3 and P4 have been derived from LFM waveforms. In this paper, P4 code is used for simulation purpose where the phase sequence of this code is given by [3].

$$\phi_i = \left(\frac{\pi}{N}\right)(i-1)(i-1-N) \quad (1)$$

Where, $i=1, 2, \dots, N$. and N is the code length.

For P4 code of length 13-element, The PSLR of its ACF is equal to 17.32 dB where this value is get enhanced as the code length increases.

3. Conventional side-lobe reduction techniques

Variant side-lobes reduction techniques are applicable according to the transmitted waveform type. In this section, traditional techniques for side-lobe reduction applied with Barker and polyphase coded waveforms are presented.

3.1. Side-lobe reduction techniques for Barker coded waveform

The first method to reduce the side-lobe level is to design a mismatched filter directly from the codes. Variant techniques are applied to design such mismatched filter. The least mean squares principle was used to design a side-lobe reduction mismatched filter by Ackroyd and Ghani [9]. Zoraster applied the linear programming (LP) algorithm to minimize the peak sidelobe level [10]. Then, the basic integrated side-lobe level (ISL) minimization technique was developed by Baden and Cohen to allow weighting the side-lobe energy for the minimization process [11].

Another way is employing an additional reduction filter after the matched filter. Rihaczek and Golden introduced a simple structure with appropriate performance called the (R-G) filter [12]. Hua and Oksman, based on linear programming (LP) algorithm, developed the (R-G) filter (R-G)_{LP} filter [13]. Jung, et al used Lagrange multiplier method to improve the (R-G) filter performance and it is called (R-G)_{LS} filter [14].

3.2. Side-lobe reduction techniques for polyphase coded waveform

Although, aperiodic polyphase codes possess good ACF, efforts have been exerted to further enhance their PSLR. Some of them are classical amplitude weighting, least square amplitude and phase weighting and sliding windows [9, 15].

Also, Woo and Griffiths developed a side-lobe canceller to reduce PSL and integrated side-lobe level (ISL) [16, 17]. The modified version of Woo filter reduces the PSL and also overcomes the main lobe splitting drawback presented in Woo filter [18].

4. The derived analytical formula for side-lobe cancellation filter

The conventional side-lobe reduction techniques had been applied according to the transmitted waveforms type and resulted in reducing the Side-Lobe Level (SLL), while the generic OP-F transfer function, for Side-Lobes Cancellation (SLC), is applicable to any phase coded waveform for completely cancelling the range time side-lobes output of MF. The OP-F follows the MF as shown in figure 1.

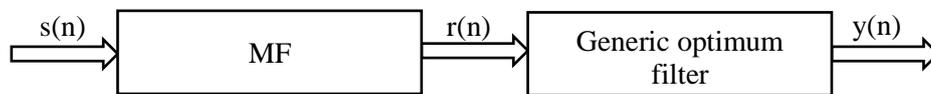


Figure1. The proposed structural block diagram for ACF SLC.

The OP-F is based on the concept of inverse filter which had been applied only to binary phase coded waveform in time domain [9, 19, and 20]. Recently this approach has been applied for Barker phase coded waveforms in frequency domain [21]. In this work, the same concept is generalized to be applicable for any phase coded signal through driving a generic formula for the OP-F transfer function. In the following subsections, a mathematical derivation of a generic OP-F transfer function, OP-F realization and performance analysis are discussed.

4.1. Mathematical derivation of a generic OP-F transfer function

The main idea behind the general formula is to investigate the phase elements that construct the phase coded waveform and then relate these elements with the construction of its ACF.

Consider a phase coded waveform with length N, let N = 4 for convenience, with sampling rate equal one sample per sub-pulse (Ns=1). The phase sequence (ϕ_i) of this signal is given by:

$$\phi_i = \{a_0, a_1, a_2, a_3\} \quad (2)$$

Where, (a_i) is the phase value and $i=0, \dots, N-1$ and $a \in [0, 2\pi[$.

The corresponding phase coded complex signal (S) can be written as:

$$S_k = e^{j\phi_k} = \{e^{ja_0}, e^{ja_1}, e^{ja_2}, e^{ja_3}\} \quad k=0, 1, \dots, N-1 \quad (3)$$

The output of the MF, which is the ACF of the signal S, is given by:

$$r(n) = \sum_{k=0}^{N-1-n} S_k S_{k+n}^* \quad n=0, 1, \dots, N-1 \quad (4)$$

So, the ACF sequence, $r(n)$, of the signal (S_k) is calculated as:

$$r(n) = \left\{ \begin{array}{l} e^{j(a_1-a_4)} + e^{j(a_1-a_3)} + e^{j(a_2-a_4)} + e^{j(a_1-a_2)} + e^{j(a_2-a_3)} + e^{j(a_3-a_4)}, 4, \\ e^{-j(a_1-a_2)} + e^{-j(a_2-a_3)} + e^{-j(a_3-a_4)}, e^{-j(a_1-a_3)} + e^{-j(a_2-a_4)}, e^{-j(a_1-a_4)} \end{array} \right\} \quad (5)$$

The ACF, $R(e^{j\omega})$, of length equal (L) in frequency domain is given by:

$$R(e^{j\omega}) = \sum_{n=0}^{L-1} r(n) * e^{-j\omega n} \quad L=0, 1, \dots, 2N-1 \quad (6)$$

Substitution of equation (5) into equation (6) and taking $N=4$, yields:

$$R(e^{j\omega}) = e^{-j3\omega} \{R_0 + R_1 + R_2 + R_3 + R_4 + R_5 + R_6\} \quad (7)$$

Where,

$$R_0 = 4 \quad (7.1)$$

$$R_1 = \{2 * \cos(a_1 - a_4) * \cos(3 * \omega) - 2 * \sin(a_1 - a_4) * \sin(3 * \omega)\} \quad (7.2)$$

$$R_2 = \{2 * \cos(a_1 - a_3) * \cos(2 * \omega) - 2 * \sin(a_1 - a_3) * \sin(2 * \omega)\} \quad (7.3)$$

$$R_3 = \{2 * \cos(a_1 - a_2) * \cos(\omega) - 2 * \sin(a_1 - a_2) * \sin(\omega)\} \quad (7.4)$$

$$R_4 = \{2 * \cos(a_2 - a_4) * \cos(2 * \omega) - 2 * \sin(a_2 - a_4) * \sin(2 * \omega)\} \quad (7.5)$$

$$R_5 = \{2 * \cos(a_2 - a_3) * \cos(\omega) - 2 * \sin(a_2 - a_3) * \sin(\omega)\} \quad (7.6)$$

$$R_6 = \{2 * \cos(a_3 - a_4) * \cos(\omega) - 2 * \sin(a_3 - a_4) * \sin(\omega)\} \quad (7.7)$$

Equation (7) represents the ACF of four phase coded signal in frequency domain which is the input to the generic OP-F. It consists of two parts; the main lobe $ML(e^{j\omega})$ and side-lobes $SL(e^{j\omega})$. This can be written as follow:

So, equation (7) can be written as.

$$R(e^{j\omega}) = ML(e^{j\omega}) + SL(e^{j\omega}) \quad (8)$$

Based on the inverse filter criteria, the transfer function of the OP-F that cancels the side lobes is given by:

$$H_{opt}(e^{j\omega}) = ML(e^{j\omega}) * [R(e^{j\omega})]^{-1} \quad (9)$$

Using equation (8) and equation (9), the OP-F output $Y(e^{j\omega})$ is given by:

$$Y(e^{j\omega}) = H_{opt}(e^{j\omega}) * R(e^{j\omega}) = ML(e^{j\omega}) \quad (10)$$

By the same manner, an equation for any phase coded system of any length could be derived. So, for N phase coded signal sequence $\phi_i = \{a_1, a_2, \dots, a_N\}$, a general form for the OP-F transfer function can be written as:

$$H_{opt}(e^{j\omega}) = \frac{N}{N + \beta_1 - \beta_2} \quad (11)$$

Where,

$$\beta_1 = 2 * \sum_{k=1}^{N-1} \sum_{i=1}^{N-k} \cos(a_k - a_{N+1-i}) * \cos(N+1-i-k) * \omega \quad (11.1)$$

$$\beta_2 = 2 * \sum_{k=1}^{N-1} \sum_{i=1}^{N-k} \sin(a_k - a_{N+1-i}) * \sin(N+1-i-k) * \omega \quad (11.2)$$

Thus the generic formula of the OP-F transfer function equation (11) enables us to calculate the OP-F coefficients that are result in cancelling the range time side-lobes of any phase coded waveform that has any length.

4.2. Realization of the generic OP-F

The OP-F realization can be executed in both frequency domain and time domain.

4.2.1. Time domain realization. For the considered phase coded signal of length N=4, the equivalent Z-transform of its ACF, R(Z), given by equation (4) is written as:

$$R(Z) = r_0 + r_1 z^{-1} + r_2 z^{-2} + r_3 z^{-3} + r_4 z^{-4} + r_5 z^{-5} + r_6 z^{-6} \quad (12)$$

Where, the term r_3 represents the main lobe amplitude and equal to (N=4) while the other terms represent the side-lobes. So, equation (12) can be divided into two terms main lobe and side-lobes and hence written as the OP-F transfer function in Z-Domain can be written as:

$$H_{opt}(Z) = ML(Z)(R(Z))^{-1} \quad (13)$$

So, a general form for the proposed OP-F in time domain considering any phase coded signal with any length N can be written as:

$$H_{optN}(Z) = \frac{NZ^{-(N-1)}}{\sum_{n=0}^{2(N-1)} r(n) Z^{-n}} \quad (14)$$

Which represents a recursive filter with an Infinite Impulse Response (IIR) given by:

$$h_{opN}(n) = \frac{1}{2\pi} \int_{-\pi}^{\pi} H_{opN}(e^{j\omega}) e^{jn\omega} d\omega \quad , n=0, \pm 1, \pm 2, \dots, \pm\infty \quad (15)$$

(This filter is unstable if one or more of its poles are located outside the unit circle in the Z-domain and also $h_{opN}(n)$ represents a non-causal IIR filter with infinite length). To circumvent this problem, the filter is approximated with a non-recursive Finite Impulse Response (FIR) filter. So, a truncation process accompanied with a shift in the impulse response with order equal to M should be applied to the IIR filter to be implemented as FIR filter with total length equal to (2M).

Numerical calculations of the PSLR and the hardware expenses of the FIR filter against the truncation order (M), for three different phase coded signals (Barker code, polyBarker and polyphase P4) all of equal length N=13, are reported in Table (1).

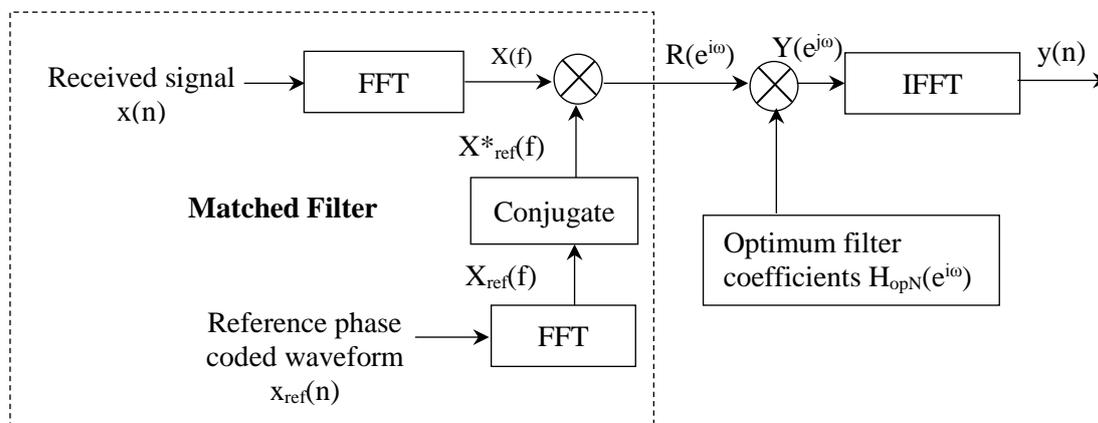
The hardware expenses and the side-lobe reduction level of the FIR OP-F are proportional to the truncation order (M).

Table 1. FIR OP-F truncation order versus PSLR and hardware expenses.

Truncation Order (M)	Filter Length	PSLR(dB)			Hardware Expense
		Phase coded signal			
		Barker Code	Poly Barker	Poly phase (P4)	# Adders & Multipliers
10	20	23.8	20	10.58	11
20	40	39.86	32.66	14.78	21
30	60	55.1	41.91	21.11	31
40	80	69.9	51.7	25.19	41
50	100	84.1	59.99	26.58	51
Matched Filter output		22.28	25.13	17.31	

4.2.2. Frequency domain realization. In this subsection, frequency domain realization of the OP-F is considered. Figure (2) shows a general block diagram for frequency domain realization of the MF accompanied with the generic OP-F. The figure is divided into two main parts; first one is the MF realization and the other is the proposed OP-F.

For MF, it is realized in frequency domain by applying the received signal, $x(n)$, to Fast Fourier Transform (FFT), with length equal to or more than the total number of range cells, to get its frequency response, $X(f)$. Then, $X(f)$ is multiplied by the complex conjugate of the frequency response of the reference signal, ($X_{\text{ref}}^*(f)$), which is already calculated and stored in the receiver.

**Figure 2.** Frequency domain realization of MF and optimum filter.

The result of the multiplication, $R(e^{j\omega})$, represents the correlation function of the two signals. $R(e^{j\omega})$ is directly multiplied by the stored OP-F transfer function coefficients, $H_{\text{opN}}(e^{j\omega})$, which has been calculated offline using the derived OP-F transfer function in equation (11) by Matlab program according to the reference phase coded elements. The output of the multiplication process, $Y(e^{j\omega})$, is applied to an Inverse Fast Fourier Transform (IFFT) to retrieve the output in time domain, $y(n)$. This approach is a direct representation of the derived generic formula without any truncation and its main advantage is achieving of complete elimination of the range time side-lobes.

5. Performance analysis of the generic OP-F

The performance analysis of the generic OP-F for different types of phase coded signals are examined, considering different conditions such as ideal (noise free) signal, noisy signal, and Doppler shifted signal. The three different phase coded signals, mentioned before, are generated to represent the

received echo from the target, and then they applied to the MF followed by the OP-F. Finally, the performance is tested through the receiver detection curve.

5.1. Case I: Ideal noise free signal

In this case, the generated signals represent a received echo from a fixed target (i.e. Doppler shift is zero, $f_d=0$). The PSLR at the output of the MF is found to be 22.27(dB) for Barker code, 25.13(dB) for polyBarker code and 17.31(dB) for polyphase P4 code.

When the MF output passes to the OP-F, the side-lobes are completely cancelled as shown in figure 3 and the PSLR tends to approximately infinity without any losses in the main lobe.

5.2. Case II: Noisy signals

Here, a normal Gaussian noise is added to the three mentioned signals with variable Signal to Noise Ratios (SNR). The added noise affects the performance of the OP-F leading to degrade the PSLRs at its output. Table (2) reports the PSLR values at the output of both MF and OP-F for the three selected signals at different values of SNR while figure 4 shows these signals at SNR equal to 10 dB.

Although the added noise degrades the PSLR at the OP-F output, but it is still greater than that at the MF output.

Table 2. PSLRs of MF and OP-F output at different SNRs.

Phase Coded Signals (N=13)	SNR			SNR		
	4dB	8dB	10dB	4dB	8dB	10dB
	MF output PSLR(dB)			Optimum Filter output PSLR(dB)		
Barker code	13.7	18.24	19.47	14.54	22.64	26.62
PolyBarker code	14.27	18.71	20.62	17.05	22.57	27.58
Polyphase P4 code	13.26	15.97	16.47	15.59	23.64	27.67

5.3. Case III: Doppler shift effect

For radar systems, the echo signal received from a moving target is modulated by Doppler frequency, f_d , that results in a phase shift, ϕ_d from one PRI, T_r , to another where:

$$\phi_d = 2\pi f_d T_r \quad (16)$$

The phase shifted received echo signal can be written as:

$$S_d(n) = X(n) e^{j2\pi f_d m T_r} \quad (17)$$

Where, m is the PRI index.

As the amplitude of MF output modulates according to f_d value, the side-lobes become asymmetric around the main lobe. When the MF output is applied to the OP-F no complete cancellation of the side-lobes is achieved and residues exist at the OP-F output and hence degrade the PSLR. Figure 5 shows the Doppler effect on both the MF and OP-F outputs in case of noise free signal when $f_d = 0.156F_r$, where F_r is the pulse repetition frequency. Table (3) lists the PSLR at the output of MF and OP-F for both fixed target and moving target.

Although, the Doppler effect results in degradation in the PSLR at out of both MF and OP-F, but still the PSLR at the out of OP-F is greater than that of the MF output.

Table 3. PSLR at MF and OP-F outputs of fixed and moving targets.

Code	PSLR (dB)			
	Fixed Target		Moving Target ($f_d=0.156F_r$)	
	MF	OP-F	MF	OP-F
Barker code	22.27	317.96	22.27	53.13
PolyBarker code	22.28	316.97	22.28	54.35
Polyphase P4 code	17.32	311.16	17.27	53.66

5.4. Case IV: Two close targets

In this case, two close targets with different SNR are represented to evaluate the detection performance of the optimum filter. For this purpose and to preserve constant Probability of False Alarm (P_{fa}) equal to 10^{-6} , Smallest of (SO) algorithm is used for CFAR processor as this algorithm is intended to ensure good detection for neighboring targets [22]. The first target has SNR = 10dB while the second one has SNR= 6dB. Figure 6 shows that, when the MF is used alone the target with high SNR is detected while the target with smaller S/N is miss detected. On the other hand, figure (7) shows that the two targets are successfully detected when the SLC optimum filter is used.

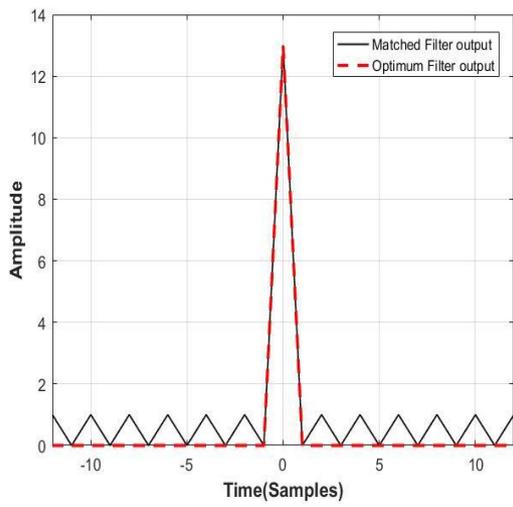
5.5. Receiver detection curves

To investigate the detection performance of the OP-F over the MF alone; the echo signals received from a fixed target is simulated and applied to both MF alone and MF followed by OP-F. Then, both outputs are applied to CFAR processor to evaluate the Probability of Detection (P_d) for each output at different SNR values preserving constant Probability of False Alarm (P_{fa}) equal to 10^{-6} . Figure 8 shows the detection curves for the three mentioned waveforms. These figures validate the outperforming detection performance of the MF with OP-F over MF alone.

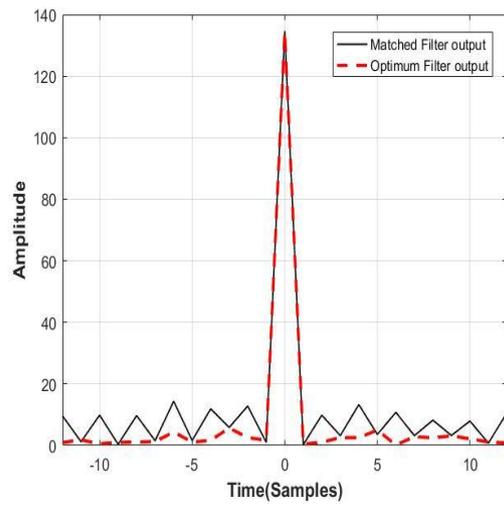
6. Conclusion

Range-time side-lobes cancellation techniques in phase coded pulse compression radars had been specially designed according to the type of the used waveform. In the present work, a mathematical formula has been derived for a generic optimum filter which is applicable to any kind of phase coded waveform with any length to completely cancel range-time side-lobes. The derived formula has been validated for Barker, polyBarker, and polyphase coded waveforms. Implementation of the proposed filter in frequency domain provides better performance than that of time domain.

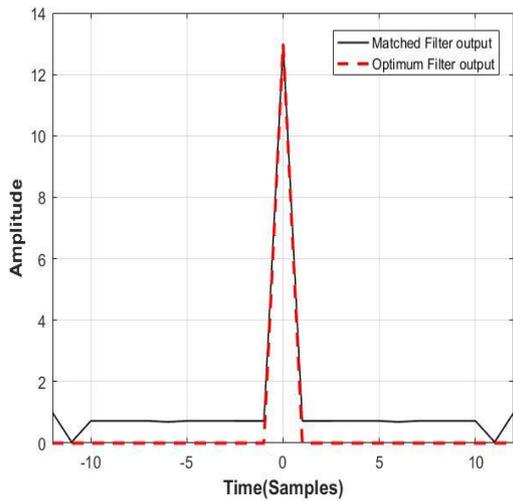
Performance comparison of the used waveforms through the derived formula has been carried out and its superiority compared to MF alone through detection curves has been validated.



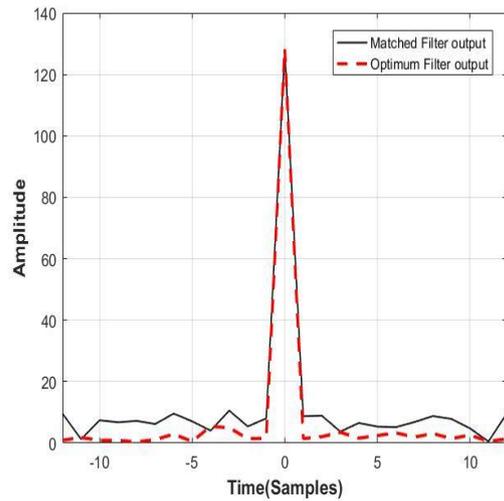
(a)



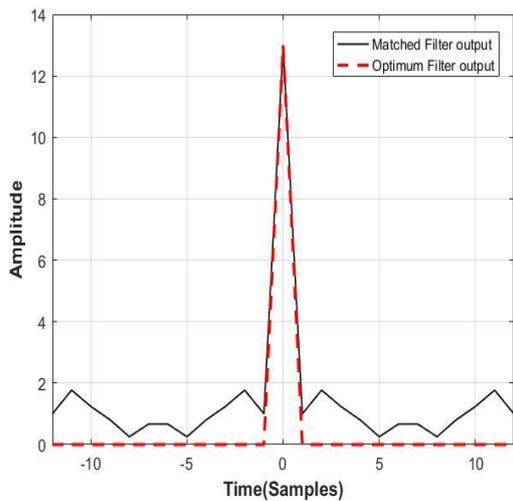
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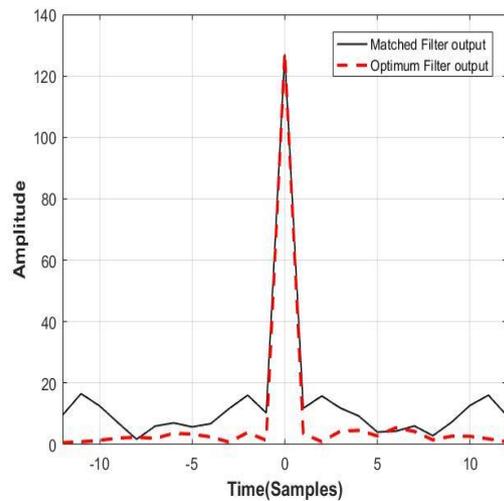
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(b)



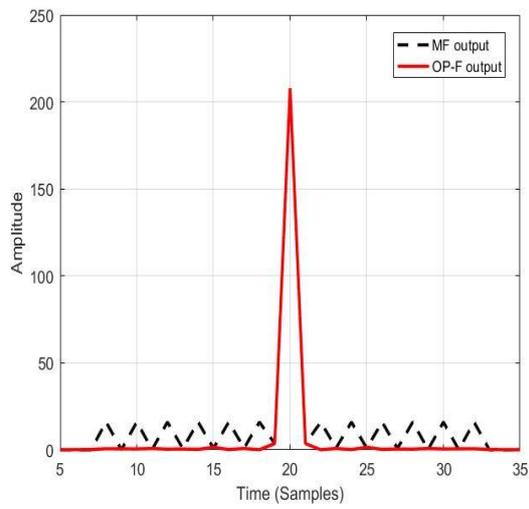
(c)



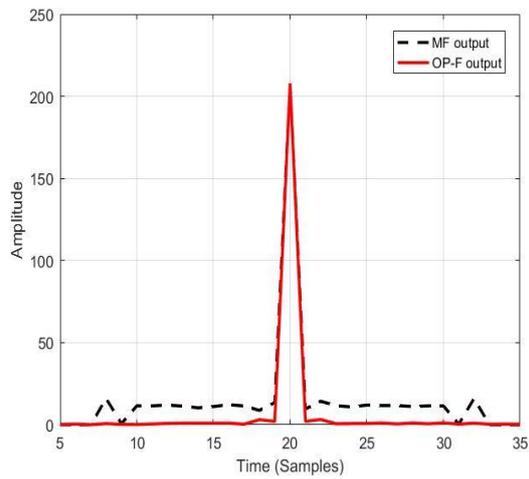
(c)

Figure 3. ACF and OP-F outputs of ideal signals (a) Barker, (b) PolyBarker and (c) Polyphase P4

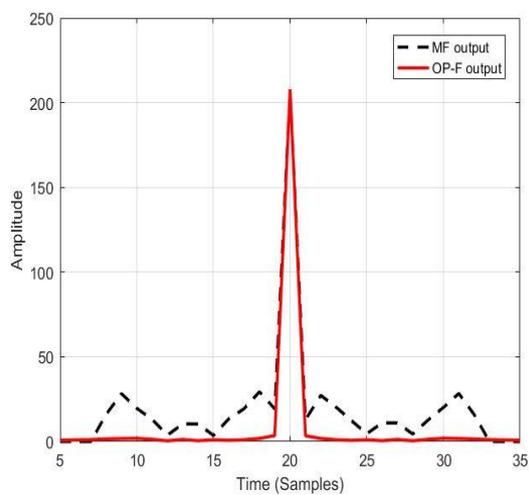
Figure 4. ACF and OP-F outputs of noisy signals(SNR=10 dB) (a) Barker, (b) PolyBarker and (c) Polyphase P4



(a)

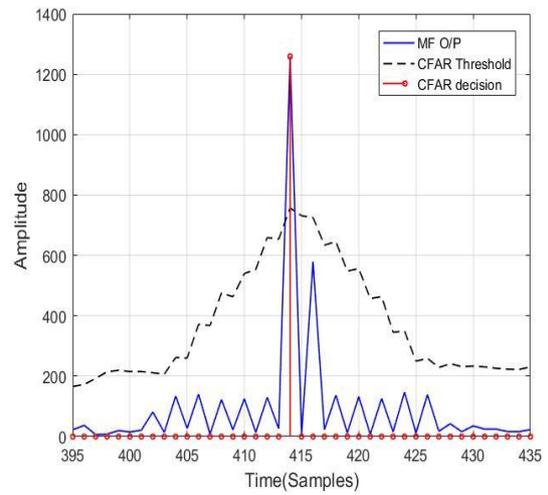


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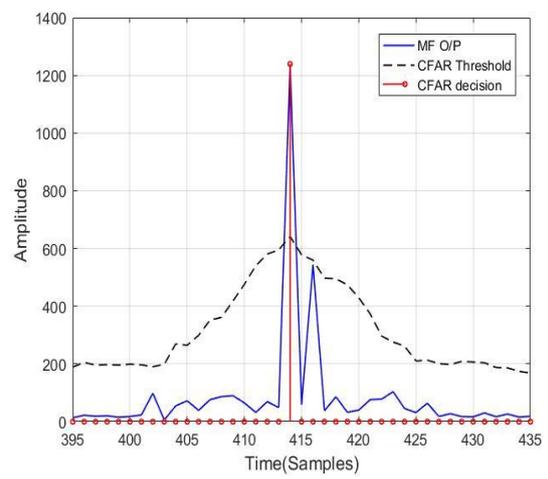


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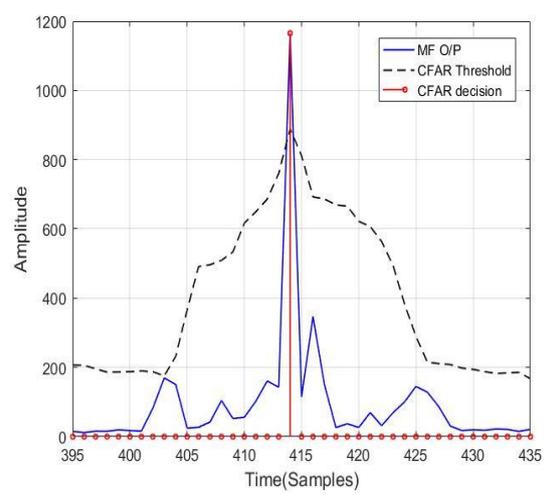
Figure 5. MF and OP-F outputs of moving target ($f_d=0.156F_r$) (a) Barker code, (b) polyBarker code and (c) polyphase P4



(a)

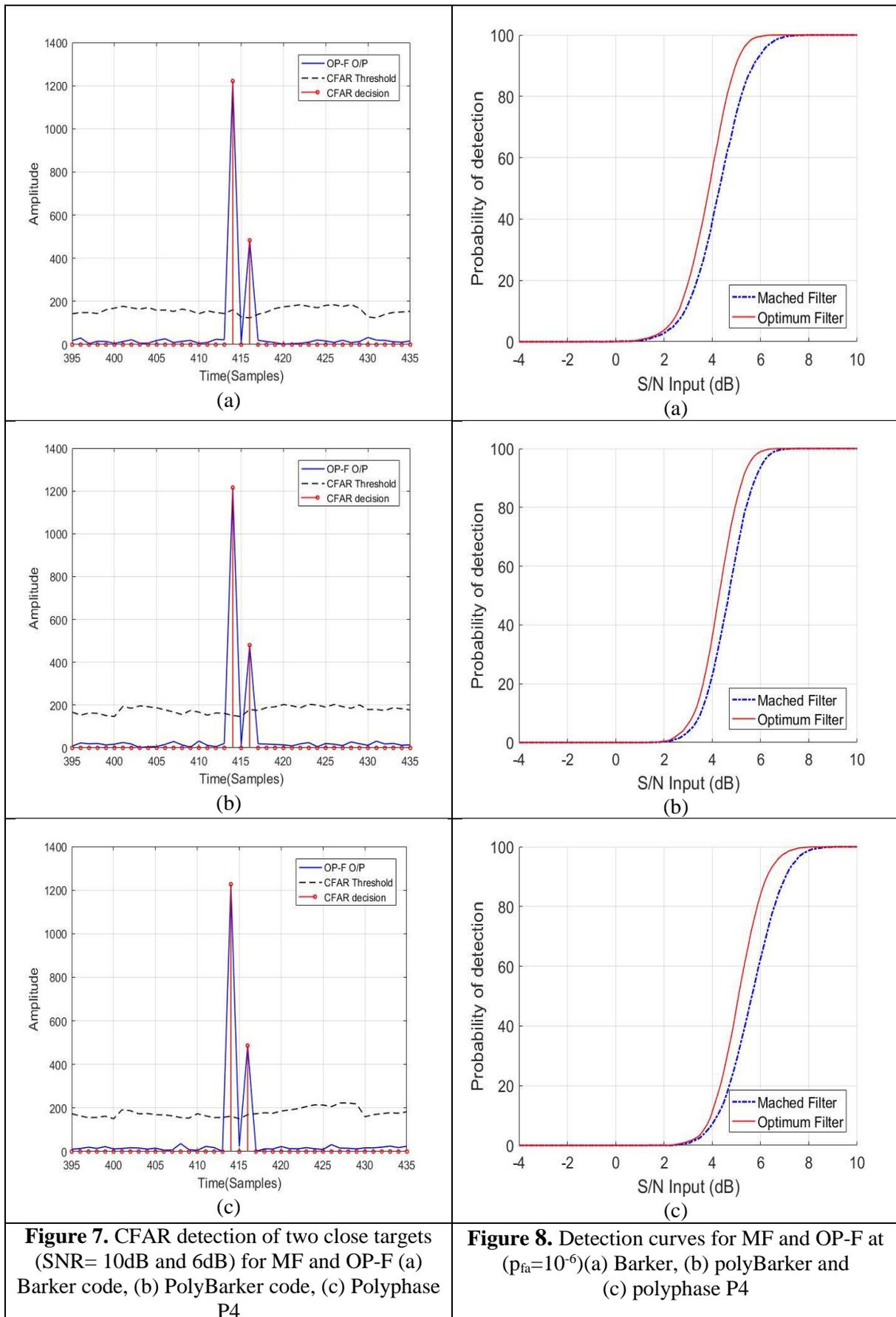


(b)



(c)

Figure 6. CFAR detection of two close targets (SNR= 10dB and 6dB) for MF alone (a) Barker code, (b) PolyBarker code, (c) Polyphase P4



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