Reducing the Sidelobes in Doppler-Range Beam Pattern and Controlling the Frequency Channel in SIAR

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1. Introduction

The synthetic impulse and aperture radar (SIAR) is a multi-input multi-output orthogonal and multicarrier frequency radar capable of detecting a variety of targets for military, medical, and transportation applications [1, 2]. This radar was initially developed by the Office National d’Etudes et de Recherches Aérospatiales in France (Figure 1) and later by the Xidian University in China [4, 5]. Microwave sparse array SIAR is an example of microwave band SIAR [6], while experimental radar from China’s Xidian University [7] and coastal radar [8, 9] are examples of metric band SIAR. The SIAR radar transmits like a radio station, with each antenna or the entire receiving array capable of revealing the target to the receiver. Since the SIAR radar has a longer coherent integration time [10] than phased array radars (PAR), it is able to integrate more pulses, compensating for a low signal-to-noise ratio. Additionally, because it employs orthogonal signals, it can monitor the space in all directions and does not produce a transmission beam, as orthogonal frequency diverse array does [11]. The beam of this radar is formed by compensating the transmitting and receiving apertures on the receiver side.

SIAR operates based on two main technical principles. Firstly, it employs multiple orthogonal carrier frequencies, resulting in each antenna and the entire transmitting array emitting in an omnidirectional fashion. This characteristic leads to the formation of a synthetic aperture at the receiver due to the delay between the transmitting and receiving arrays. Secondly, while the bandwidth of each transmitted signal is low, the aggregate bandwidth remains high. This allows for the creation of a synthetic impulse postprocessing, even without pulse compression.

The major contributions of this study are as follows:

(i) Propose a novel SIAR processing flow designed to reduce the sidelobes.
(ii) Explain the implementation of windowing and amplitude weighting in this new method.
(iii) Highlight the advantages offered by our proposed method.
(iv) Investigate the effect of various windowing techniques on sidelobe levels.

Lastly, we provide a comparison between our proposed method and existing methods for sidelobe reduction to highlight the benefits and improvements offered by our approach.

1.1. Issue and Goal. Despite its many capabilities, SIAR faces the challenge of a high sidelobe in the range-Doppler frequency [12]. A radar is disrupted by the presence of jammer, clutter, and false targets when sidelobe levels are high [13]. Thus, reducing the amplitudes of sidelobes is the objective that experts attempt to achieve. The use of different processes and appropriate waveforms [14] are among the methods of reducing the level of the sidelobe in this radar [15]. Following is a discussion of suitable waveforms and processes for this radar.

1.2. SIAR Processing Flow for PPFCA and PPFCNA. One of the most common waveforms in SIAR is the pulse waveform with different frequency codes. In SIAR radar, each transmitting antenna has a distinct frequency from other antennas. A frequency code determines the frequency of each antenna [16]. In this radar, two frequency coding methods, pulse-to-pulse frequency code non-agile (PPFCNA) and pulse-to-pulse frequency code agility (PPFCA), are used for different transmitted pulses in order to achieve the lowest sidelobe amplitude for the pulse waveform. The PPFCNA method employs a frequency code series for each pulse repetition interval (PRI), so that each pulse has the same code. In contrast, the PPFCA method employs a frequency code series for each PRI; consequently, the frequency code series must regenerate for each PRI. In the PPFCNA method, the sidelobe and main lobe are coherent, and the sidelobe and main lobe amplitudes are increased. In the PPFCA method, the sidelobes are incoherent due to the random generation of frequency codes in each PRI. After integration, the main lobe increases, but the amplitudes of the sidelobes do not increase. The two most important coding methods in SIAR are PPFCNA and PPFCA, which are used according to their applications.

The received signal is processed in Figure 2. This process is according to Chen and Wu [12] and consists of three main parts:

1. (1) Digital beamforming (DBF) compensates for the receiving aperture.
2. (2) The matching filter compensates for the transmitting aperture.
3. (3) Coherent integration of output pulses.

In the process, the time delay caused by each receiving antenna is compensated by DBF. All azimuth and elevation angles for each receiving antenna are checked. At the matched filter, the time delay caused by each transmitted antenna is compensated; thus, the azimuth angle, elevation angle, and ranges are verified. The process in a matched filter may occur in the time, frequency, or time–frequency domain [17]. The frequency domain process is used, and each pulse undergoes its own DBF and matched filter processes. In this article, the output of the Doppler-range is investigated.

1.3. Proposed Method. In array radars, the common weighting is based on the weighting of receiving antennas. Weighting the amplitude and phase of receiving antennas or
subarrays [18] can reduce the amplitudes of sidelobes or create a null in PAR [19, 20]. However, in SIAR radar, weighting can be done at a deeper level. In general, parts of a SIAR radar that can be weighted include the weighting of the receiving antenna and the weighting of the submatched filter. In this paper, the proposed method of amplitude weighting in the submatching filter is utilized to reduce the sidelobe amplitude and frequency channel control effects.

In the PPFCA method, frequency codes are randomly regenerated for each PRI. The proposed method must modify the processing flow in order to gain access to frequency channels. Consequently, SIAR provides direct access to each subsignal, allowing for the possibility of submatched filter weighting. In other words, the amplitude of each subsignal with a specific frequency is weighted. Additionally, this method is able to completely eliminate a subsignal, which means it can eliminate the effect of a particular frequency. The previous method combines all submatched filters, whereas the proposed method treats each submatched filter separately. Thus, the previous method cannot directly evaluate all subsignals or completely eliminate specific frequency effects. Simulations reveal that the proposed method reduces sidelobe levels more effectively than the previous method, while also providing superior frequency control. The following section discusses the mathematical aspects of SIAR, including the SIAR model, the previous processing flow, and the new method.

2. Data Model

In this section, the formulation of the problem from the transmitted signals to the process output for a moving target is investigated.

2.1. Transmitter Design. The transmitted signals from each antenna (subsignals) are:

\[ S_{n,k}(t) = g(t)e^{j2\pi f_{n,k}t}; \quad k = 1 \sim N_e, \quad n = 1 \sim N \]

\[ g(t) = \begin{cases} \frac{1}{\sqrt{T_e}}; & (n-1)T_r < t < nT_r + T_e \\ 0; & \text{else} \end{cases} \]

\[ f_{n,k} = f_0 + \Delta f_{n,k} = f_0 + c_{n,k} \Delta f, \]

where \( S_{n,k}(t) \) represents the subsignals transmitted from the \( k^{th} \) antenna during the \( n^{th} \) pulse, \( N_e \) is the number of transmitting antennas, \( n \) represents the number of transmitted pulses, \( N \) is the number of maximum transmitted pulses, \( g(t) \) denotes the signal’s envelope, \( T_r \) is the transmitted pulse width, \( T_e \) is the PRI, and \( f_{n,k} \) denotes the carrier frequency of the \( k^{th} \) antenna in the \( n^{th} \) pulse. Also, \( f_0 \) represents the main carrier frequency, \( \Delta f \) is the frequency interval, and \( c_{n,k} \) is the frequency code of the \( k^{th} \) antenna in the \( n^{th} \) pulse. If a subsignal has different frequency codes in different PRIs, the PPFCNA is used; otherwise, the PPFCNA is employed. The PPFCNA method derives its optimal frequency codes from the genetic algorithm (GA) described in Chen and Wu [12]. In the PPFCNA method, frequency codes are randomly regenerated for each PRI.

2.2. Array Geometry. The array geometry is illustrated in Figure 3. The transmitting and receiving arrays are arranged into two concentric rings sharing the same center. The positions of the transmitting antennas are defined by radius \( d_c \) and angles \( \theta_{ck} \), while the positions of the receiving antennas are determined by radius \( d_r \) and angles \( \theta_{rl} \).

\[ \theta_{ck} = \left[ 0, \frac{2\pi}{N_c}, \ldots, \frac{2\pi}{N_c} - 1 \right] \frac{2\pi}{N_c} \]

\[ \theta_{rl} = \left[ \frac{\pi}{N_r}, \frac{2\pi}{N_r}, \ldots, 2\pi \right] \]

where \( d_c \) represents the radius of the transmitting ring (transmitting array), \( d_r \) is the radius of the receiving ring (receiving array), \( \theta_{ck} \) represents the azimuth angle of the \( k^{th} \) transmitting antenna position, and \( \theta_{rl} \) the azimuth angle.
of the \(i\)th receiving antenna position. Both the transmitters' and receiver’s ring antennas have a uniform distribution, with an interelement angle space of \(2\pi/N_r\) rad. Furthermore, all antennas are omnidirectional. \(N_r\) is the maximum number of the transmitting antenna and \(N_e\) is the maximum number of the receiving antenna. The center of the two arrays is identical. \(\theta_0\) is azimuth, \(\phi_0\) elevation, and \(R_0\) distance from center to target. In this paper, the receiving array is considered to be smaller than the transmitting array.

2.3. Orthogonality Condition. In each transmitted pulse, the \(k\)th subsignal that reaches the target is equal to:

\[
S_k(t - \tau_k) = g(t)e^{j2\pi f_k(t - \tau_k)},
\]

(3)

where \(\tau_k\) represents the delay of the \(k\)th transmitter antenna to the target. To establish the orthogonality condition, there is Wang et al. [13]:

\[
\begin{align*}
\int_{-\infty}^{\infty} s_k(t - \tau_k) s^*_l(t - \tau_l) dt &= 0 \\
\int_{-\infty}^{\infty} s_k(t - \tau_k) s^*_l(t - \tau_l) dt &= \frac{\sin (\pi(c_k - c_l)\Delta f. T_x)}{\pi(c_k - c_l)\Delta f. T_x} e^{j2\pi \phi} = 0
\end{align*}
\]

\[
\phi = \left(\frac{c_k - c_l}{2}\right)(f_k - f_l) - (f_k\tau_k - f_l\tau_l)
\]

\[
\Rightarrow (c_k - c_l)\Delta f. T_x = \text{int} \neq 0
\]

\[
\Rightarrow \Delta f. T_x = \text{int}.
\]

(4)

This equation is the orthogonality condition. Also, the frequency codes should be integer.

2.4. Received Signal. The receiving signal at the \(i\)th receiver in the \(n\)th pulse is:

\[
x_{n,i}(t) = g(t - \tau_{00}) e^{j2\pi f_n t} e^{j2\pi f_0 n T_r}
\]

\[
\times \sum_{k=1}^{N_e} g^{2\pi f_{lk} T_r} e^{j2\pi f_{lk} n T_r} e^{j2\pi f_{lk} \tau_{lk}}
\]

\[
f_{00} = \frac{2\nu_0 t}{c},
\]

(6)

where \(f_{00}\) is the Doppler frequency of the target. Coherent processing interval causes \(\Delta f T_r/\nu_0 < 1\); thus, \(e^{j2\pi f_{lk} \tau_{lk}} \approx e^{j2\pi f_{lk} \tau_{lk}}\). The narrowband transmitted signals result in:

\[
e^{j2\pi f_{lk} \tau_{lk}} \approx e^{j2\pi f_0 \tau_{00}}.
\]

(7)

With omitting of constant \(e^{-j2\pi f_0 \tau_{00}}:

\[
x_{n,i}(t) = g(t - \tau_{00}) e^{j2\pi f_n T_r} e^{j2\pi f_0 n T_r}
\]

\[
\times \sum_{k=1}^{N_e} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \Delta f.T_r}.
\]

(8)

According to the explained diagram (Figure 2), DBF weight function is:

\[
{w}_r(\theta, \varphi) = e^{-j2\pi f_0 \tau_{rl}}, \frac{d_r \cos (\varphi) \cos (\theta - \theta_0)}{c},
\]

(9)

and the matched filter is:

\[
H_n(t, \theta, \varphi) = \frac{1}{N_e} g(t) \sum_{k=1}^{N_e} e^{j2\pi f_{lk} T_r} e^{-j2\pi f_{0k} T_r}.
\]

(10)

The matched filter is used in the frequency domain. Fast Fourier transform of the signal and the matched filter is used and the result of both multiplies to each other to make the output of the matched filter in the frequency domain. Then, inverse fast Fourier transform makes the output of the matched filter, \(y_n(t)\).

After that, range-Doppler is calculated as:

\[
Z(i, \tau) = \frac{1}{N_e} \sum_{n=1}^{N} y_n(t) e^{-j2\pi f_{0k} (\tau - \tau_{00})}, \quad i = 1 \sim N.
\]

(11)

2.5. SIAR Processing Flow. The signal received from each receiver in baseband and after the low-pass filter for the \(n\)th pulse is:

\[
x_{n,i}(t) = g(t - \tau_{00}) e^{j2\pi f_0 \tau_{00}} e^{j2\pi f_n T_r}
\]

\[
\times \sum_{k=1}^{N_e} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \Delta f.T_r}.
\]

where \(f_{00}\) is the Doppler frequency of the target. Coherent processing interval causes \(\Delta f T_r/\nu_0 < 1\); thus, \(e^{j2\pi f_{lk} \tau_{lk}} \approx e^{j2\pi f_{lk} \tau_{lk}}\). The narrowband transmitted signals result in:

\[
e^{j2\pi f_{lk} \tau_{lk}} \approx e^{j2\pi f_0 \tau_{00}}.
\]

(7)

With omitting of constant \(e^{-j2\pi f_0 \tau_{00}}:

\[
x_{n,i}(t) = g(t - \tau_{00}) e^{j2\pi f_n T_r} e^{j2\pi f_0 n T_r}
\]

\[
\times \sum_{k=1}^{N_e} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \tau_{lk}} e^{j2\pi f_{lk} \Delta f.T_r}.
\]

(8)

According to the explained diagram (Figure 2), DBF weight function is:

\[
{w}_r(\theta, \varphi) = e^{-j2\pi f_0 \tau_{rl}}, \frac{d_r \cos (\varphi) \cos (\theta - \theta_0)}{c},
\]

(9)

and the matched filter is:

\[
H_n(t, \theta, \varphi) = \frac{1}{N_e} g(t) \sum_{k=1}^{N_e} e^{j2\pi f_{lk} T_r} e^{-j2\pi f_{0k} T_r}.
\]

(10)

The matched filter is used in the frequency domain. Fast Fourier transform of the signal and the matched filter is used and the result of both multiplies to each other to make the output of the matched filter in the frequency domain. Then, inverse fast Fourier transform makes the output of the matched filter, \(y_n(t)\).

After that, range-Doppler is calculated as:

\[
Z(i, \tau) = \frac{1}{N_e} \sum_{n=1}^{N} y_n(t) e^{-j2\pi f_{0k} (\tau - \tau_{00})}, \quad i = 1 \sim N.
\]

(11)
The proposed method uses weighting of submatched filters in each channel separately instead of weighting the whole channels. This technique employs a fact that subsignals in SIAR have different and separate frequencies, each subsignal has a specific frequency. Therefore, each subsignal can be separated and weighted; consequently, weighting in the submatched filter significantly reduces sidelobes and increases frequency channel control.

### 2.7. Proposed Method

The proposed method uses weighting of submatched filters in each channel separately instead of weighting the whole channels. This technique employs a distinct processing flow that is introduced as follows. First, each subsignal must be separated in each receiver. Due to the fact that subsignals in SIAR have different and separate frequencies, each subsignal can be separated at the receiver. The output of the summation for each received channel is then passed to the matched filter. In this section, amplitude weighting is performed for the submatched filter. Finally, the output of each channel’s matched filter is added together. This procedure is shown in the diagram of Figure 4.

Received subsignal is:

$$x_{n,l,k} = g(t - \tau_{00})e^{j2\pi f_{0}nT \tau_{l}}e^{j2\pi f_{0}T_{r} \tau_{l,k}}e^{j2\pi f_{0}T_{a} \tau_{l,k}}e^{j2\pi f_{0}(t-\tau_{0})}.$$  

(13)

DBF for receiving antenna is:

$$w_{n,r} = e^{-j2\pi f_{0} \tau_{d}}.$$  

(14)

The PAR amplitude weighting does not apply in the proposed method. The output of the DBF is:

$$x'_{n,l,k} = g(t - \tau_{00})e^{j2\pi f_{0}nT \tau_{l}}e^{j2\pi f_{0}T_{r} \tau_{l,k}}e^{j2\pi f_{0}T_{a} \tau_{l,k}}e^{j2\pi f_{0}(t-\tau_{0})}e^{j2\pi f_{0} \Delta f(t-\tau_{0})}.$$  

(15)

In this part, each subsignal from each antenna is added to the same subsignal in other antennas. The submatched filter and weight are:

$$h_{n,k} = a_{n,k}(t)e^{-j2\pi f_{0} \Delta f}e^{-j2\pi f_{0} \tau_{d}},$$  

(16)

where $h_{n,k}$ is the submatched filter, $a_{n,k}(t)$ is the weight of each submatched filter that can be changed adaptively. The proposed method weights the submatched filter proportionally to the frequency channel of the transmitted signal. In other words, amplitude weighting corresponds to increasing or decreasing the effect of a subsignal with a particular frequency. When the PAR weight is applied, all submatched filters are assigned the same weight. Consequently, the PAR amplitude weighting model cannot eliminate the specific frequency channel in the SIAR process. However, if one of the submatched filters in the new method has zero weight, then the subsignal is completely eliminated from the process. In addition, the new method cannot be implemented in the previous matched filter diagram due to the summation of all submatched filters used in the previous method, which prevents separate access to submatched filters. Thus, each submatched filter cannot be matched with subsignals separately or eliminate the specific channel entirely. The proposed method has a great flexibility, and the amplitude weighting can be proportional to the frequency effect. If a target at a particular frequency has a high sidelobe in the proposed process, the sidelobe can be weakened by weighting to have minimal effects on the output. The output of each submatched filter is:

$$y_{n,k} = x''_{n,k}(t) \times h_{n,k}(t, \theta, \varphi).$$  

(17)

The summation of the whole submatched filter, called matched filter output, is:
The transmitting ring radius is $r_t = 45$ m and receiving ring radius is $d_r = 22.5$ m. The number of transmitting antennas is equal to 25, from $\theta_{d1} = 0$ to $\theta_{d25} = 0.24 \times 2\pi / 25 \text{ rad}$ by the step of $2\pi / 25 \text{ rad}$, so the transmitters’ angle positions are $\theta_{d1} = [0, 2\pi/25, 4\pi/25, \ldots, 24 \times 2\pi/25]$, and the number of receiving antennas is equal to 25, from $\theta_{d1} = 2\pi/25 \text{ rad}$ to $\theta_{d25} = 2\pi \text{ rad}$ by the step of $2\pi / 25 \text{ rad}$, so the receivers’ angle positions are $\theta_{d1} = [\pi / 25, 3\pi / 25, \ldots, 2\pi]$. The range of frequency codes is $(-12, 12)$. In the mentioned interval, the GA frequency code resulting from Chen and Wu [12] is $c_k = [-6, 12, -4, 5, -9, 1, -2, 8, -7, 10, -11, 3, 0, -3, 11, 10, 7, -8, 2, -1, 9, -5, 4, -12, 6]$. This code is used for the PPFCNA method. Also, the PPFCA frequency code is $c_{n,k} = \text{random}(-12, 12)$. Unique frequency codes are assigned to each antenna in each PPFCA pulse. The central carrier frequency is 100 MHz, the variable frequency is 20 kHz, and the sampling frequency is 0.64 MHz. The PRI is 1 ms, all transmitted pulses are 64 pulses, and the pulse width is 50 $\mu$s. The target is at a distance of 90 km, at azimuth angle of 0, and at elevation of 5 with a velocity of 300 m/s. The outputs of the new method without weighting are shown in Figure 7.

This simulation demonstrates that the proposed method reduces sidelobes more than Chen and Wu [12]. The new method reduces the sidelobes of the PPFCA and PPFCNA methods, although the PPFCA reduction is marginal. The reason is that PPFCA is one of the best methods to reduce the sidelobes. The proposed amplitude weighting by
windowing is used to assist the proposed diagram in further reducing the sidelobes.

Weights are states in Table 1, and relations between weights, frequency codes, and antenna numbers are presented in Figures 8–11.

Figure 8 shows that the 13th transmitting antenna has \( a_{13} = 1 \) and \( c_{13} = 0 \). This type of representation assists in determining the impact of each channel in the process. Consequently, these plots are useful for extracting amplitude weight futures.

3.1. Simulation One. The coding method used in this simulation is PPFCNA. Figure 12 depicts the output of the matched filter for various pulses and the output of the matched filter for a single pulse. As illustrated in Figure 12(a), the output of the

<table>
<thead>
<tr>
<th>Window</th>
<th>( a_k )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Kaiser</td>
<td>[0.78984, 0.82170, 0.85134, 0.87862, 0.90341, 0.92558, 0.94503, 0.96165, 0.97536, 0.98610, 0.99381, 0.99845, 1, 0.99845, 0.99381, 0.98610, 0.97536, 0.96165, 0.94503, 0.92558, 0.90341, 0.87862, 0.85134, 0.82170, 0.78984]</td>
</tr>
<tr>
<td>Chebyshev</td>
<td>[1, 0.36947, 0.42919, 0.48877, 0.54687, 0.60213, 0.65324, 0.69891, 0.73800, 0.76952, 0.79264, 0.80676, 0.81151, 0.80676, 0.79264, 0.76952, 0.73800, 0.69891, 0.65324, 0.60213, 0.54687, 0.48877, 0.42919, 0.36947, 1]</td>
</tr>
<tr>
<td>GA for PPFCNA</td>
<td>[0.77554, 0.27728, 0.60214, 0.82555, 0.86968, 0.57236, 0.50255, 0.86567, 0.38734, 0.48532, 0.78331, 0.64384, 0.62716, 0.32427, 0.72314, 0.75298, 0.77657, 0.76799, 0.51833, 0.50235, 0.72675, 0.80054, 0.49516, 0.63164]</td>
</tr>
<tr>
<td>GA for PPFCNA</td>
<td>[0.20948, 0.60891, 0.41899, 0.34343, 0.35207, 0.69325, 0.93496, 0.21405, 0.89354, 0.45222, 0, 0.72873, 0.53405, 0.10186, 0.12426, 0.69524, 0.63169, 0.29735, 0.71444, 0.40958, 0.55533, 0.91008, 0.06004, 0.85006, 0]</td>
</tr>
</tbody>
</table>

**Figure 8:** The Kaiser window.

**Figure 9:** The Chebyshev window.
matched filter in each pulse has nearly identical amplitudes and sidelobes due to the application of the PPFCNA technique. Therefore, the level of the sidelobes increases by summing the output of the matched filters. In Figure 12(b), the delay compensates independently for each channel in the new method. This method does not permit sidelobes to rise as much as the previous method because it checks the subsignals separately. This approach functions similarly to a local compensation and reduces sidelobes after channel summations for each pulse or even for the output of the entire matched filter.

The output of the simulations is in Figure 13.

The amplitude of the sidelobe has been significantly reduced. The output of the Doppler-range for the optimal window of the GA is in Figure 14.

By comparing the output of Figures 6 and 14, the amplitude of sidelobes has decreased significantly.

3.2. The Second Simulation. This simulation uses the PPFCA coding method. The output of the matched filter for a variety of pulses and the output of the matched filter for a single pulse are in Figure 15.

In Figure 15(a), the output of the matched filter for various pulses has different sidelobes, but the main beam is almost identical. This is the result of using the PPFCNA technique. Although sidelobes are not optimal in a single pulse, the sidelobes are reduced when multiple pulses are combined. However, the output of the submatched filter in a single pulse is obtained in Figure 15(b). Similar to the PPFCNA mode, each channel’s delay compensation is handled independently. Consequently, the new strategy utilizes the benefits of PPFCA earlier because it compensates for the delay in subsignals. Then, subsignals combination and pulses combination are used to reduce sidelobes. The output of the simulation is in Figure 16.

Although PPFCA is one of the effective ways to reduce the sidelobes, amplitudes of sidelobes in this method are almost halved by windowing. The output of the simulation using the GA window is in Figure 17.

3.3. Simulations’ Results. The output of the simulations is shown in Table 2:

The results presented in Table 2 indicate that the SIAR windowing method is suitable for reducing the sidelobe in both the PPFCNA and PPFCA methods. The maximum reduction of sidelobe in PPFCNA using the GA window is 8.7908 dB, while the minimum reduction of sidelobe using the Chebyshev window is 2.7606 dB. The sidelobe reduction
in the PPFCA method utilizing the GA is 3.3779 dB. Additionally, the Chebyshev window has reduced the sidelobe level by 0.9032 dB. In the new method, sidelobe reduction and channel control are obtained simultaneously. Moreover, it is predicted that if the amplitudes of the output’s sidelobes in the matching filter are different at any moment \( a_n, \beta(t) \neq a_\ell \), they cause the lower level of the sidelobe.

Finally, we examine the proposed method PPFCNA and reference PPFCNA to show the benefits of the proposed method. Five targets are chosen in different ranges and velocities. As shown in Figure 18, the previous method has lots of high sidelobe level while the new method reduces the sidelobes and shows the five targets clearly. This simulation, however, can be improved by other proper windows.

4. Conclusion

SIAR is an orthogonal multicarrier frequency radar that can reduce the level of sidelobes in a variety of ways, including the use of different frequency codes, waveforms, and coding methods. One of the main challenges in this radar is the level of sidelobes, which was attempted to be reduced using PPFCNA and PPFCA frequency coding methods. This article introduces weighting in SIAR as a suitable method for
FIGURE 14: Range-Doppler output for PPFCNA with GA.

FIGURE 15: Matched filter for PPFCNA. (a) Whole matched filter outputs for each pulse (previous method) and (b) submatched filter’s outputs in a pulse (proposed method).

FIGURE 16: The output of the simulation.
reducing the sidelobe amplitude at the deeper level of the matched filter. This weighting is different from the weighting of PAR and the previous SIAR process, necessitating a unique process flow. In PAR, the signals sent from the transmitter are inaccessible to the receiver because all transmitters operate at a fixed frequency, whereas in SIAR, each antenna transmits at a unique frequency. Since the transmitted signals are orthogonal, a submatch filter can be used independently. This paper demonstrates that due to the flexibility of the matched filter and the weighting of the SIAR, not only can sidelobes be reduced, but connections between the frequency and amplitude at the receiver can also be established. If the selected weight in the submatched filter is zero, the output channel in the corresponding submatched filter will be eliminated from the processing flow. Hence, amplitude weighting functions as a frequency filter. Simulations indicate that the proposed windowing method for PPFCNA and PPFCA with Chebyshev, Kaiser, and GA windows reduced sidelobes more effectively than the windowless method.

![Figure 17: The output for GA.](image)

<table>
<thead>
<tr>
<th>Weighting method</th>
<th>PPFCNA SLL (dB)</th>
<th>PPFCA SLL (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>No weighting</td>
<td>-7.1590</td>
<td>-11.7858</td>
</tr>
<tr>
<td>Chebyshev window</td>
<td>-9.9196</td>
<td>-12.6890</td>
</tr>
<tr>
<td>Kaiser window</td>
<td>-12.1347</td>
<td>-12.7700</td>
</tr>
<tr>
<td>GA window</td>
<td>-15.9498</td>
<td>-15.1637</td>
</tr>
</tbody>
</table>

![Figure 18: The output of the multitargets simulation for PPFCNA.](image)
Data Availability
No data has been used.

Conflicts of Interest
The authors declare that they have no conflicts of interest.

References


