Anti-Sweep-Jamming Method Based on the Averaging of Range Side Lobes for Hybrid Modulation Proximity Detectors

QILE CHEN1, XINHONG HAO1, ZHIJIE KONG2, AND XIAOPENG YAN1

1Science and Technology on Electromechanical Dynamic Control Laboratory, Beijing Institute of Technology, Beijing 100081, China
2China Academy of Launch Vehicle Technology, Beijing 100081, China

Corresponding author: Xinhong Hao (haoxinhong@bit.edu.cn)

This work was supported in part by the National Natural Science Foundation of China under Grant 61871414, and in part by the Equipment Pre-Research Foundation under Grant 61406190101.

ABSTRACT Hybrid modulation continuous waveform (HMCW) proximity detectors emit binary phase-coded chirp (BPCC) waveforms. Therefore, they possess the advantages of binary phase-coded modulation and chirp modulation simultaneously. To enhance the anti-sweep-jamming performance of HMCW detectors, this paper analyses the response of an HMCW detector under sweep jamming and proposes an anti-jamming method based on the novel average range side lobe (ARS) approach. Similar to the instant correlation with harmonic demodulation (ICHD) method, the ARS method also combines time-domain instantaneous correlation with frequency-domain harmonic demodulation to obtain the range information of the target. The difference is that the ARS method uses a fast Fourier transform (FFT) instead of a band pass filter (BPF) to extract the harmonic envelope and averages multiple harmonic coefficients obtained via the FFT. Because the correlation side lobes of chaotic codes follow a Gaussian distribution, the output of an HMCW detector under sweep jamming is suppressed after ARS processing. The proposed method is proven to be feasible and effective through numerical simulations. The results show that the ARS method offers improved anti-sweep-jamming performance.

INDEX TERMS Electronic warfare, HMCW detectors, electronic counter-countermeasures (ECCM), anti-sweep jamming.

I. INTRODUCTION

Continuous waveform (CW) detectors maintain high signal-to-noise ratios while requiring much less peak power than corresponding pulse sensor systems [1]. They are widely used in vehicle collision warning systems [2], radar altimeters [3], [4] and other related applications. However, their performance is seriously affected by interference and jamming [5]. In particular, sweep jamming [6]–[8], in which a narrowband jamming signal is rapidly sweeping over a wide frequency band, has a significant effect on CW detectors. Electronic countermeasure (ECM) simulation tests have verified that sweep jamming has become a serious threat to CW detectors [9]. Thus, suitable attention should be paid to the anti-sweep-jamming performance of CW detectors.

The associate editor coordinating the review of this manuscript and approving it for publication was Filbert Juwono. 

CW detectors can be classified into many types depending on their modulation modes, which include linear frequency modulation (LFM), binary phase shift keying (BPSK) modulation, and hybrid modulation (HM) [10]. With its large bandwidth and high range resolution, LFM is one of the most popular modulation modes. However, LFM detectors can be easily intercepted and jammed [1], [11]. Because of their low probability of interception, the anti-jamming performance of BPSK detectors is better than that of LFM detectors. However, BPSK detectors tend to exhibit high sensitivity to Doppler shifts, which limits their application for high-speed moving targets [12], [13]. As a combination of LFM and BPSK modulation, HM possesses the advantages of both modulation modes [11], [14], [15]. HM was first used in communication systems to overcome the limited waveform variability of chirp signals in a restricted time-frequency region [16]. Because of its good performance, it has been
increasingly widely applied in the field of detection. The ambiguity function of an HM waveform is a “thumbtack” function, indicating a good range resolution [17]. In addition, HM waveforms have larger time-bandwidth products than either LFM or BPSK waveforms and consequently have a lower probability of interception. Moreover, their Doppler tolerance is greater. With all these beneficial characteristics, HM has recently been attracting increasing attention and is being gradually applied in a number of CW detectors, leading to the emergence of a growing class of HMCW detectors [18], [19]. Reference [18] adopts the time-domain correlation as the ranging method. However, this method does not address the cross modulation of HM waveforms. Reference [19] introduced an instant correlation with harmonic demodulation (ICHD) ranging method for HMCW detectors. The ICHD method combines time-domain instantaneous correlation with frequency-domain harmonic demodulation to obtain the range information of the target. It makes the most of the correlation performance of the HM waveform and has good noise suppression ability. However, this method does not consider the effect of the intended jamming type, such as sweep jamming.

The key to enhancing anti-jamming performance, especially anti-sweep-jamming performance, is to suppress the range side lobes. The correlation side lobes of HM waveforms may lead to the appearance of ghost targets and degrade the jamming suppression performance of such a detector [20]. Many previous studies have addressed side lobe suppression. References [21]–[24] proposed several sequence design methods for generating sequences with side lobes lower than \(-80\) dB. However, these sequences consist of arbitrary phase codes that are not suitable for HMCW detectors. References [25], [26] used a genetic algorithm to optimize chaotic codes. After optimization, the side lobes of the chaotic codes were approximately \(-20\) dB. However, the optimization algorithm was too complex for implementation on low-power platforms. Filter design can also enable side lobe suppression, at the cost of degradation in the range resolution. However, such methods will increase the algorithm complexity and resource consumption. References [27]–[29] designed a CW random binary phase modulation radar with side lobes of approximately \(-30\) dB. This radar took advantage of the statistical characteristics of random binary codes to achieve greatly improved anti-jamming performance. However, this radar was not suitable for high-speed targets, as its ranging principle was complex, and its real-time performance was poor.

In the present paper, to enhance the anti-sweep-jamming performance of HMCW detectors, we first analyse the anti-sweep-jamming performance of an HMCW detector that adopts the ICHD ranging method. Then, an average range side lobe (ARS) method that uses fast Fourier transform (FFT) to extract the harmonic envelope and averages multiple harmonic coefficients obtained via the FFT is proposed according to the statistical characteristics of the output of the HMCW detectors under sweep jamming. The ranging and anti-sweep-jamming performance of the ARS are quantitatively analysed and compared with those of ICHD. Finally, simulations are performed to prove the superiority of the ARS. The results of the simulations show that, compared with the ICHD, the ARS can suppress the effect of sweep jamming by more than 20 dB.

The remainder of the paper is organized as follows. The mathematical models of a BPCC waveform and sweep jamming are introduced in Section II. The principle of HMCW detectors and the response of such a detector under sweep jamming are introduced in Section III. Section IV introduces the principles of the ARS anti-jamming method. Section V reports numerical simulations. Our conclusions are given in Section VI.

II. MATHEMATICAL MODELS

A. BPCC WAVEFORM MODEL

The BPCC transmitted waveforms are illustrated in Figure 1. Here, we choose chaotic codes generated from the logistic map as the basis for generating an HM waveform, and the chaotic codes vary with the LFM period. Because the BPCC waveform inherits the correlation properties of the chaotic binary codes, its side lobes follow a random distribution [30], [31].

Taking up-ramp LFM as an example, the transmitted BPCC waveform can be mathematically expressed as

\[
s(t) = A_t \sum_{n=0}^{\infty} \sum_{k=0}^{P-1} c_{k,n} v(t - kT_c)
+nPT_c) e^{(2\pi f_0 (t-nPT_c) + \frac{1}{2} \beta (t-nPT_c)^2) \text{rect}(t-nPT_c)},
\]

where \(c_{k,n} = \pm 1\) are the chaotic codes, with \(c_{k,n}\) being the kth element of the nth sequence, \(T_c\) is the chip epoch, \(P\) is the length of each sequence, \(f_0\) is the carrier frequency, \(\beta = \frac{B}{PT_c}\) is the LFM slope, with \(B\) being the LFM bandwidth, \(A_t\) is the amplitude of the transmitted signal, \(v(t) = \begin{cases} 1, & 0 < t < T_c \\ 0, & \text{else} \end{cases}\) is a gate function with a duration of \(T_c\), and \(\text{rect}(t) = \begin{cases} 10 < t < PT_c \\ 0, \text{else} \end{cases}\) is a gate function with a duration of \(PT_c\). The transmitted BPCC waveform is illustrated in figure 1.

Let \(u_n(t)\) denote the complex envelope of the nth segment of the transmitted BPCC waveform. Then, by normalizing the
amplitude to unity, \( u_n(t) \) can be expressed as

\[
u_n(t) = \sum_{k=0}^{P-1} c_{k,n}v(t - kT_c - nPT_c)e^{j\pi\beta(t-nPT_c)^2}. \tag{2}\]

Let \( F \{ \cdot \} \) represent the Fourier operator. Then, it is obvious that the spectrum of \( u_n(t) \) can be expressed as

\[
F\{ u_n(t) \} = \frac{1}{2\pi} F\left\{ \sum_{k=0}^{P-1} c_{k,n}v(t - kT_c - nPT_c) \right\} \otimes F\{ e^{j\pi\beta(t-nPT_c)^2} \}. \tag{3}
\]

As seen from Equation (3), the spectrum of the transmitted BPCC waveform is the convolution of the LFM and BPSK modulation spectra. Its bandwidth is jointly determined by the chip epoch and the LFM bandwidth. It retains the noise-like characteristics of the chaotic codes while having a wider bandwidth.

\[B. \quad \text{SWEEP JAMMING MODEL}\]

In sweep jamming, the jammer sweeps its frequency from one value to the next and does not share its power among multiple frequencies. Suppose that the sweep jamming parameters are as follows: the start frequency is \( f_{j0} \), the stop frequency is \( f_{jN} \), the frequency step is \( \Delta f \), the dwell time is \( T_{dw} \), and the number of steps is \( N = \frac{f_{j0}-f_{jN}}{\Delta f} \). The jamming signal can then be expressed as

\[
s_j(t) = A_j e^{j\pi f_{j0}(t)+\varphi_j}, \tag{4}\]

where \( f_j(t) = f_{j0} + \Delta f \sum_{n=0}^{N-1} kP_{dw}(t-kT_{dw}) \) is the instantaneous frequency of the jamming signal, \( A_j \) is the amplitude, \( \varphi_j \) is the initial phase, and \( P_{dw}(t) = \begin{cases} 1, & 0 < t < T_{dw} \\ 0, & \text{else} \end{cases} \).

\[\text{III. RESPONSE OF AN HMCW DETECTOR UNDER SWEEP JAMMING}\]

\[A. \quad \text{PRINCIPLE OF HMCW DETECTORS}\]

A block diagram of an HMCW detector is shown in Figure 3. The detector consists of two modules: a radio frequency module and a signal processing module. The radio frequency module includes a voltage-controlled oscillator (VCO), a phase modulator, and a circulator, among other components. It generates the transmitted signal and mixes the target echo signal with a local reference signal (LRS) to obtain the beat signal.

\[
s_{id}(t, \tau) = \begin{cases} A_{ref}A_1A_k(\tau) \sum_{n=0}^{P-1} c_{k,n}v(t - kT_c - nPT_c - \tau)e^{j2\pi [f_0 + \beta(t-nPT_c)](t - nPT_c)^2} & , \quad 0 \leq t - nPT_c < \tau; \\
A_{ref}A_1A_k(\tau) \sum_{n=0}^{P-1} c_{k,n}v(t - kT_c - nPT_c - \tau)e^{j2\pi [f_0 + \beta(t-nPT_c)](t - nPT_c)^2} & , \quad \tau \leq t - nPT_c < T 
\end{cases}
\tag{7}\]

\[
A_1 = A_2 = 1, \quad A_3 = 1, \quad A_0 = 1.
\]

FIGURE 2. Illustration of time-frequency analysis for the sweep jamming signal.

The main purpose of the signal processing module is to extract the range and velocity information, for which purpose the ICHD ranging method is adopted. After sampling, the beat signal from the radio frequency module is first multiplied by a delayed replica of the chaotic codes to realize instantaneous correlation. Then, a band pass filter (BPF) is used to extract the harmonic information of the beat signal. Finally, the harmonic information obtained from the BPF is mixed with a reference signal from a direct digital synthesizer (DDS) to obtain a Doppler signal that carries the range and velocity information of the target.

The reflected signal is an attenuated and delayed version of the transmitted signal and is expressed as

\[
s_i(t, \tau) = A_c(\tau)s_i(t - \tau), \tag{5}\]

where \( A_c(\tau) \) is the attenuation factor, which varies with the time delay, and \( \tau \) is the round-trip time needed for the signal to propagate from the detector to the target and back. For a specified target with a velocity \( v_0 \) and at an initial distance \( R \), the time delay of the target echo signal can be expressed as \( \tau = \frac{2(2R - vt_0)}{c}, \) where \( c = 3 \times 10^8 \text{m/s} \) is the speed of electromagnetic waves.

The LRS is a chirp signal generated by the VCO which is expressed as

\[
s_{ref}(t) = A_{ref} \sum_{n=0}^{\infty} e^{j2\pi f_0(t-nPT_c)+n\beta(t-nPT_c)^2} \text{rect}(t - nPT_c). \tag{6}\]

By mixing the target echo signal with the LRS, the beat signal is obtained, which can be expressed as (7), shown at the bottom of this page.
In general, the detection range of an HMCW detector is short, which means that $\tau \ll PT_c$, the beat signal can be ignored in the time range of $0 \leq t - nPT_c < \tau$. In addition, the amplitude of the reflected signal has little impact on the performance of the detector, therefore, its amplitude can be considered to be constant. Thus, we can obtain the following simplified beat signal:

$$s_{\text{id}}(t, \tau) = \frac{A_{\text{ref}}A_Ce^{j\phi_0}}{2}\sum_{n=0}^{\infty}\sum_{k=0}^{P-1}c_k n e^{j2\pi f_0 t - nPT_c - \tau}e^{j2\pi\beta t^2},$$

(8)

After sampling, the beat signal is sent to the signal processing module, and the ICHD method is used to obtain the information of the target. The ICHD method, which combines time-domain instantaneous correlation with harmonic demodulation, is implemented in two steps:

(1) Under the assumption that the predetermined range of the detector is $R_o$, the beat signal is first multiplied by a replica of the chaotic codes that is delayed by $\tau_o = \frac{2R_o}{c}$. By defining $t_n = (t - nPT_c)$, the instantaneous correlation signal obtained in this first step can be expressed as

$$s_{\text{corre}}(t_n, \tau) = \frac{A_{\text{ref}}A_Ce^{j\phi_0}}{2}\sum_{n=0}^{P-1}\sum_{k=0}^{P-1}c_k n e^{j2\pi f_0 t_n - kT_c}$$

$$e^{j2\pi\beta\tau^2}e^{j2\pi\beta\tau^2}.$$  

(9)

(2) Then, $s_{\text{corre}}(t_n, \tau)$ is passed through a BPF to obtain the $m_o$ harmonic, where $m_o$ satisfies $m_o = B\tau_o$. The $m_o$ harmonic of $s_{\text{corre}}(t_n, \tau)$ is

$$s_{m_o}(t_n, \tau) = \frac{A_{\text{ref}}A_Ce^{j\phi_0}}{2}e^{2\pi(jf_0 t - \frac{1}{2}\beta t^2)}X_A(\tau - \tau_o, 0)e^{2\pi m_o\omega_m t + \phi_0},$$

(10)

where $X_A(\tau - \tau_o, 0) = \int_0^{PT_c} s(t)\chi(t - \tau_o)\chi(t - \tau_o)dt$. As the motivation of the target, the Doppler information is also carried by $s_{m_o}(t_n, \tau)$. After mixing with the signal generated by the DDS at the frequency $m_o\omega_m$, the Doppler signal can be obtained as

$$s_{\text{id}}(t_n, \tau) = A_{\text{ref}}e^{2\pi(jf_0 t - \frac{1}{2}\beta t^2)}X_A(\tau - \tau_o, 0)e^{2\pi m_o\omega_m t + \phi_0}. $$

(11)

The Doppler signal inherits the envelope of $s_{m_o}(t_n, \tau)$. The main lobe of the signal, where the amplitude is highest, is formed when the distance between the target and the detector is approximately $R_o$, and side lobes are formed at other distances. Therefore, we can apply a threshold to obtain the range information. By varying the time delay of the replica and the corresponding harmonic number, the predetermined range of the detector can be controlled.

B. RESPONSE OF AN HMCW DETECTOR UNDER SWEEP JAMMING

When the HMCW detector receives a sweep jamming signal, it mixes the sweep jamming signal with the LRS.
The resulting beat signal under jamming can be expressed as

\[ s_{\text{id}}(t) = r(t) \times s_n^m(t) = \frac{A_j A_i}{2} \sum_{n=0}^{\infty} \sum_{k=0}^{P-1} c_{k,n} v(t - nP_T_c - \tau_o) e^{i(2\pi (f_0 - f_j(u))(t - nP_T_c)) + \pi \beta (t - nP_T_c)^2} \text{rect}(t - nP_T_c). \]  

(12)

After instantaneous correlation, the beat signal can be expressed as (13), shown at the bottom of this page.

In general, the dwell time is on the order of milliseconds, i.e., \( T_{\text{dw}} \gg PT_c \), therefore, we can still assume that \( t_n = t - nPT_c \) and \( s_{\text{id}}(t_n) \) can be simplified to

\[ s_{\text{id}}(t_n) = \frac{A_j A_i}{2} \sum_{k=0}^{P-1} c_{k,n} v(t_n - kT_c - \tau_o) e^{i(2\pi (f_0 - f_j(u))t_n + \pi \beta t_n^2 + \pi \beta (t_n)^2} \text{rect}(t_n - nPT_c). \]

(14)

\( s_{\text{id}}(t_n) \) is the BPCC signal, whose carrier frequency follows the sweep of the jamming frequency, and its spectrum is as

\[ X_{\text{id}}(m, u) = \frac{A_j A_i}{2} F \left\{ \sum_{k=0}^{P-1} c_{k,n} v(t_n - kT_c - \tau_o) \right\} F \left\{ e^{i(2\pi (f_0 - f_j(u))t_n + \pi \beta t_n^2 + \pi \beta (t_n)^2} \text{rect}(t_n - nPT_c) \right\}. \]

(15)

where \( \bigcirc \) represents convolution.

With the variation of the jamming frequency, the spectrum of \( s_{\text{id}}(t_n) \) gradually spans the bandwidth of the BPF. Because the energy of the jamming signal is much higher than that of a real echo signal, the \( m_o \) harmonic energy is saturated, and the true target is obscured, causing the detector to be jammed.

IV. PRINCIPLES OF THE ARS METHOD

The basic principle of the ARS method is illustrated in figure 5. This method takes advantage of the statistical properties of chaotic codes. Unlike the ICHD ranging method, it realizes harmonic demodulation by means of an FFT after instantaneous correlation and takes the average of harmonic coefficients.

A. FFT-BASED HARMONIC EXTRACTION METHOD

Because the velocity of the target satisfies \( v_0 \ll c \), the target echo signal is constant over a time interval of \( PT_c \). The time delay of the target echo signal can therefore be discretely

\[ s_{\text{id}}(t) = \frac{A_j A_i}{2} \sum_{n=0}^{\infty} \sum_{k=0}^{P-1} c_{k,n} v(t - kT_c - nP_T_c - \tau_o) e^{i(2\pi (f_0 - f_j(u))(t - nP_T_c)) + \pi \beta (t - nP_T_c)^2} \text{rect}(t - nP_T_c). \]

(13)

\[ s_{\text{corr}}(t_n, i) = \frac{A_{\text{ref}} A_i A_e(\tau_i)}{2} \left\{ \sum_{k=0}^{P-1} c_k v(t_n - kT_c - \tau_o) \right\} \left\{ \sum_{k=0}^{P-1} c_{k,n} v(t_n - kT_c - nP_T_c - \tau_o) e^{i(2\pi (f_0 - f_j(u))(t_n - nP_T_c)) + \pi \beta (t_n - nP_T_c)^2} \text{rect}(t_n - nPT_c) \right\}. \]

(18)
B. RANGING PRINCIPLE OF THE ARS METHOD

Because the velocity of the target satisfies \( v_0 \ll c \), we can assume that the radial distance between the HMCW detector and the target is constant over a time interval of \( GPT_c \).

In the ARS method, the \( m_o \) harmonic values obtained via the FFT are averaged to suppress the range side lobes. After averaging, the output is as

\[
S_G(m_o, i) = \frac{A_{ref}A_e(t_\tau)}{2A_t} \sum_{i=1}^{i+G} \frac{PT_c}{G} \int s_i(t_n, t_\tau) s_i(t_n, t_\tau) dt_n.
\]  

Here, we assume that the resolution of the detector is \( \Delta R \). When \( |t_i - t_\tau| < \frac{2\Delta R}{c} \), the harmonic values obtained via the FFT all lie in the main lobe. Therefore, after averaging, the output of the ARS method also lies in the main lobe. When \( |t_i - t_\tau| \geq \frac{2\Delta R}{c} \), the harmonic values obtained via the FFT lie in the side lobes. Because of the diversity of chaotic codes, the range side lobes corresponding to these values are different, and they satisfy the following Gaussian distribution

\[
P(X) = \frac{1}{\sqrt{2\pi B\sigma^2}} e^{-\frac{X^2}{2B\sigma^2}}
\]  

where \( X \) is the value of the range side lobe, \( P(\cdot) \) is the probability density function, and \( B \) is the variance of \( X \). After averaging, \( |S_G(m_o, g)| \) satisfies

\[
|S_G(m_o, g)| = \begin{cases} 
\text{main lobe}, & |t_i - t_\tau| < \Delta R \\
0, & |t_i - t_\tau| \geq \Delta R.
\end{cases}
\]  

As seen from formula (23), the ARS method maintains the ranging performance of the ICHD method. In addition, it can significantly suppress the range side lobes with a suitable \( G \).

C. ANTI-SWEEP-JAMMING PRINCIPLE OF THE ARS METHOD

The spectrum of the HMCW detector under sweep jamming is as

\[
X_{G_{jcor}}(m_o, i, u) = \frac{A_iA_t}{2} F\left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] \otimes F\{ e^{i2\pi (f_0 - f_o) t_n + \pi \beta t_n^2} \}. 
\]  

Usually, the dwell time is on the order of milliseconds, \( T_{dw} \gg GPT_c \). Therefore, the output of the ARS method under sweep jamming can be expressed as

\[
X_{G_{jcor}}(m_o, i, u) = \frac{A_iA_t}{2G} F\left[ \sum_{i=1}^{i+G} \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] \otimes F\{ e^{i2\pi (f_0 - f_o) t_n + \pi \beta t_n^2} \}. 
\]  

Because of the randomness of chaotic codes, \( \lim_{G \to \infty} \frac{1}{G} \sum_{i=1}^{i+G} \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) = 0 \). As \( G \to \infty \), the jamming will be completely suppressed, and the ARS output under sweep jamming will be zero.

\[
X_{G_{jcor}}(m_o, i, u) = 0.
\]  

V. SIMULATIONS AND DISCUSSION

In this section, we report simulations and discuss the detection performance achieved with the ARS method. The simulation parameters are shown in Table 1. We choose to work in the S-band because it is easy and cheap to implement, and the carrier frequency selected here is 3 GHz. According to the relationships among range resolution, bandwidth and code width, because a narrower code width leads to a higher cost, we set \( T_c = 50 ns \). Furthermore, the bandwidth is set 50 MHz to get a high range resolution of approximately \( \pm 1.5 \) m. The target is assumed to be an approaching aircraft with a velocity of 500 m/s. The parameters of sweep jamming are based on [1], [5]–[7] and some ECM simulation tests we performed.

A. SPECTRA OF BPCC WAVEFORMS

The spectra of BPCC waveforms with the same code epoch but different LFM bandwidths are shown in figure 6. The BPCC spectrum retains the noise-like characteristics of the chaotic codes while having a wider bandwidth. The BPCC bandwidth is approximately 30 MHz when the LFM bandwidth is 10 MHz, and it is approximately 70 MHz when the LFM bandwidth is 50 MHz. In other words, the BPCC bandwidth significantly increases as the LFM bandwidth \( B \) increases. According to the simulation results, the BPCC bandwidth can be approximated as \( B + \frac{1}{\tau_c} \). For a time duration

\[
S(m_f, i) = \frac{A_iA_e(t_\tau)}{2} \int \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] e^{i2\pi (f_0 - f_o) t_n + \pi \beta t_n^2} dt_n
\]  

\[
= \frac{A_iA_e(t_\tau)}{2} \int \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] e^{i2\pi (f_0 - f_o) t_n + \pi \beta t_n^2} dt_n
\]  

\[
= \frac{A_iA_e(t_\tau)}{2} \int \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] \left[ \sum_{k=0}^{P-1} c_{k,n}v(t_n - kT_c - \tau_o) \right] e^{i2\pi (m-m_o) t_n + \pi \beta t_n^2} dt_n
\]  

(19)
of $T = \frac{P T_c}{B + T}$, the interception factor for a BPCC signal is
\[
\alpha = K \left( \frac{1}{\sqrt{B + T}} \right)^{1/2}.
\]

**B. RANGING PERFORMANCE OF THE ARS METHOD**

The ARS anti-jamming method suppresses the range side lobes and hinders sweep jamming by averaging the harmonic coefficients obtained via the FFT. The range resolution of this method is no different from that of the ICHD method. Simulation results for the ARS method with different values of $G$ are shown in figure 7. The blue line, as a contrast, represents the ICHD output, the green line represents the ARS output with $G = 10$, the red line represents the ARS output with $G = 50$, and the yellow line represents the ARS output with $G = 100$. The main lobe of the ARS output lies at a distance of 6 m, and its width is approximately 6 m. It has the same range resolution as ICHD and has nothing to do with $G$. However, the amplitude of the main lobe decreases with increasing $G$, as the radial distance between the HMCW detector and the target cannot be considered constant when $G$ is too large.

**C. SUPPRESSION OF RANGE SIDE LOBES**

To facilitate the analysis of the range side lobe suppression performance of the ARS method, we convert the ARS outputs into decibels, as shown in figure 8. As seen in figure 8, the range side lobes of the ICHD output are approximately $-22$ dBm. By contrast, the range side lobes of the ARS output are lower than $-31$ dBm when $G = 10$, and they are lower than $-40$ dBm when $G$ is larger than 50. It is apparent that the range side lobe suppression performance of the ARS method is related to the value of $G$. With a suitable $G$, the range side lobe suppression performance of the ARS method is 20 dB higher than that of the ICHD method.

**D. ANTI-SWEEP-JAMMING PERFORMANCE OF THE ARS METHOD**

The signal received by an HMCW detector under sweep jamming can be divided into two cases. When the target is far from the detector and is not within its working range,
the energy of the target echo signal is too low. In this case, the received signal can simply be approximated as the sweep jamming signal. By contrast, when the target is within the working range of the detector, the received signal is the sum of the sweep jamming signal and the target echo signal.

The ARS outputs under sweep jamming with a signal-to-jamming ratio (SJR) of \(-20\) dB are shown in figure 9 and figure 10. Figure 9 shows the case in which the target is far from the detector. In this case, the ARS outputs are similar to noise. As a contrast, the ICHD output under sweep jamming is approximately \(-5\) dBm; however, the ARS output is lower than \(-17\) dBm with \(G = 10\). When \(G\) is larger than 50, the output is suppressed to lower than \(-26\) dBm. Thus, by choosing an appropriate \(G\), the ARS output can be more strongly suppressed than the ICHD output by more than 20 dB.

When the target is within the working range of the detector, the received signal simultaneously includes both the echo signal and the sweep jamming signal. The ARS outputs in this case are shown in figure 10. Because of the limited jamming suppression performance, the target is completely obscured by the jamming signal for the detector based on the ICHD method. By contrast, the target can be identified successfully by the ARS-based detector. As \(G\) increases, the identification performance greatly improves.

The output of ARS and ICHD with different SJRs is shown in figure 11. From figure 11, the jamming suppression performance of ARS is better than that of ICHD for different SJNs. The output powers of both ICHD and ARS are negatively correlated with the SJR, and their sum is a constant. We define the reciprocal of this constant as the processing gain (PG), which reflects the jamming suppression performance of each method. The PG of ICHD is approximately 24 dB. By contrast, the PG of ARS with \(G = 10\) is approximately 35 dB, and the PG of ARS is larger than 44 dB when \(G\) is larger than 50. With a suitable \(G\), the PG of ARS is larger than ICHD by more than 20 dB.

**E. INFLUENCE OF G ON THE ARS RESULTS**

The value of \(G\) has an important influence on the behaviour of the ARS method. The relationships between \(G\) and various measures of performance are shown in figure 11. Here, we select the maximum value of the range side lobes as the evaluation criterion for side lobe suppression (blue line), the maximum value of the output under sweep jamming as the evaluation criterion for jamming suppression (red line), and the maximum SJR that the detector can tolerate as the evaluation criterion for PG (new definition) (purple line). The degeneration of the main lobe is also plotted in figure 11 (green line). As seen from this figure, the maximum value of the range side lobes and output under sweep jamming are both significantly suppressed with the ARS method. The suppression performance and PG improve with increasing \(G\). However, when \(G\) is greater than 100, the suppression performance shows no further significant improvement. Moreover, when \(G\) is larger than 100, the amplitude of the main lobe is seriously degraded, which may cause the detector to lose its ranging capabilities. Therefore, the most suitable value of \(G\) is between 50 and 100.

**VI. CONCLUSION**

To enhance the anti-sweep-jamming performance of HMCW detectors, this paper has analysed the response of an HMCW detector under sweep jamming and has proposed a novel anti-sweep-jamming method called the ARS method. Similar to the ICHD method, the ARS method combines time-domain instantaneous correlation with frequency-domain harmonic demodulation, and it achieves the same range resolution as
the ICHD method. However, the ARS method uses an FFT instead of a BPF to extract harmonic coefficients and averages multiple harmonic coefficients obtained via the FFT. Because the range side lobes follow a Gaussian distribution, the range side lobes and the output under sweep jamming are both significantly suppressed after averaging with a suitable G. With the selection of a suitable G, the ARS method can suppress the range side lobes and the effect of sweep jamming more strongly than the ICHD method can by more than 20 dB.

REFERENCES

[1] Z. Kong, P. Li, X. Yan, and X. Hao, “Anti-sweep jamming design and implementation using multi-channel harmonic timing sequence detection for short-range FMCW proximity sensors,” Sensors, vol. 17, no. 9, p. 2042, Sep. 2017. doi: 10.3390/s17092042.

[2] K. Yue, X. Hao, and P. Li, “An LFMCW detector with new structure and FRFT based differential distance estimation method,” SpringerPlus, vol. 5, no. 1, p. 922, Dec. 2016. doi: 10.1186/s40064-016-2611-9.

[3] J.-H. Choi, M.-S. Jung, and K.-W. Yeom, “A design and assessment of a direction finding proximity fuze sensor,” IEEE Sensors J., vol. 13, no. 8, pp. 3079–3089, Aug. 2013. doi: 10.1109/journals.2226326.

[4] J.-H. Choi, J.-H. Jang, and J.-E. Roh, “Design of an FMCW radar altimeter for wide-range and low measurement error,” IEEE Trans. Instrum. Meas., vol. 64, no. 12, pp. 3517–3525, Dec. 2015. doi: 10.1109/tim.2015.2450294.

[5] L. D. Adams, EW104: EW Against a New Generation of Threats. Beijing, China: Publishing House of Electronics Industry, 2017, pp. 33–39.

[6] Y. S. Choi, S. Y. Lee, H. H. Choi, S. J. Lee, and C. S. Park, “The improved spatial nuller with frequency sweep jammer,” in Proc. IEEE/ION Position, Location Navigat. Symp., Monterey, CA, USA, May 2014, pp. 1084–1087.

[7] D.-Y. Choi, W.-K. Kim, J.-H. Kim, and H. Cho, “Performance of analog and digital modulation schemes under sweep jamming,” in Proc. 8th Int. Conf. Ubiquitous Future Netw. (ICUFN), Vienna, Austria, Jul. 2016, pp. 13–15.

[8] S. Mighani, M. Mivehchy, and M. F. Sabahi, “Evaluating sweep noisy barrage jamming effect on tracking radar based on functioning destruction time,” in Proc. 7th Int. Symp. Telecommun. (IST), Tehran, Iran, Sep. 2014, pp. 400–404.

[9] Z. X. P. Yan Li Hao, “Similarity-based quantization method for anti-jamming performance of radio fuze,” Acta Armamentarii, vol. 38, no. 7, pp. 1282–1288, 2017, doi: 10.3969/j.issn.1009-1093.2017.07.005.

[10] H. C. Zhao, Fundamentals and Methodology of Radar Face. Beijing, China National Defense Industry Press, 2012, pp. 91–107.

[11] M. Soumekh, “SAR-ECCM using phase-perturbed LFM chirp signals and DRFM repeat jammer penalization,” IEEE Trans. Aerosp. Electron. Syst., vol. 42, no. 1, pp. 191–205, Jan. 2006, doi: 10.1109/taes.2006.1603344.

[12] A. Akbarpour and D. Mirzahosseini, “Improving range resolution in jammed environment by phase coded waveform,” in Proc. IEEE CIE Int. Conf. Radar, Chengdu, China, Oct. 2011, pp. 226–230.

[13] X. Yang and J. Lin, “A digitally controlled constant envelope phase-shift modulator for low-power broadband wireless applications,” IEEE Trans. Microw. Theory Techn., vol. 54, no. 1, pp. 96–105, Jan. 2006, doi: 10.1109/trmtt.2005.861669.

[14] Q. Liu, Z. Huang, Y. Kou, and J. Wang, “A low-ambiguity signal waveform for pseudolite positioning systems based on chirp,” Sensors, vol. 18, no. 5, p. 1326, Apr. 2018, doi: 10.3390/s18051326.

[15] L. Wang, F. Gao, J. Xu, D. Wang, M. He, and J. Yuan, “Orthogonal wideband hybrid-coding radar waveforms design,” Signal, Image Video Process., vol. 11, no. 1, pp. 103–111, Jan. 2017, doi: 10.1007/s11760-016-0905-6.

[16] M. Kowatsch and J. Lafferl, “A spread-spectrum concept combining chip modulation and pseudonoise coding,” IEEE Commun., vol. 31, no. 10, pp. 1133–1142, Oct. 1983, doi: 10.1109/com.1983.1095745.

[17] L. Jing, “Analysis of pseudo code and LFM compound modulation signal,” Acta Armamentarii, vol. 32, no. 10, pp. 1117–1222, 2011.

[18] D. Wei, S. Zhang, S. Chen, H. Zhao, and L. Zhu, “Research on deception jamming of chaotic composite short-range detection system based on bifractal analysis and genetic algorithm–back propagation,” Int. J. Distrib. Sensor Netw., vol. 15, no. 5, May 2019, Art. no. 1550147719847444, doi: 10.1177/1550147719847444.

QILE CHEN was born in Henan, China, in 1992. He received the B.S. degree in mechatronic engineering from the Beijing Institute of Technology, China, in 2015, where he is currently pursuing the Ph.D. degree in intelligent detection and control. His research interests include signal processing, new detection theory of proximity radar, ECM, and ECCM.

XINHONG HAO was born in Henan, China, in 1974. She received the Ph.D. degree in mechatronic engineering from the Beijing Institute of Technology, in 2007. She is currently an Associate Professor with the Beijing Institute of Technology. Her main research interests include target detection theory of radio signal sensor processing, real-time signal processing, and time-frequency analysis.
ZHJIE KONG was born in Henan, China, in 1990. He received the Ph.D. degree in mechatronic engineering from the Beijing Institute of Technology, in 2018. He is currently an Engineer with the China Academy of Launch Vehicle Technology. His main research interests include radar signal processing and time-frequency analysis.

XIAOPENG YAN received the B.E. degree in mechanical and electronic engineering, the M.E. degree in pattern recognition and intelligent system, and the Ph.D. degree in mechanical and electronic engineering from the Beijing Institute of Technology, Beijing, China, in 1999, 2003, and 2009, respectively. He is currently a Professor with the School of Mechatronical Engineering, Beijing Institute of Technology, where he has been a Faculty Member, since 2003. His research interests include radio detection and signal processing in proximity sensor.

* * *