High precision phase-domain radial velocity estimation for wideband radar systems

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Abstract: Radial velocity estimation used in wide-band radar systems is investigated. By analyzing the signal of cross-correlation output of adjacent echoes, it is found that the frequency and phase of the cross-correlation output are related to the target’s radial velocity. Since the precision of the phase estimation is higher than that of the frequency, a phase-based velocity estimator is proposed. However, the ambiguity problem exists in the phase estimators, and thus the estimation of the cross-correlation of adjacent echoes (CCAE) is used to calculate the ambiguity number. The root-mean-square-error (RMSE) of the proposed estimator is derived. Simulation results show that the performance of the proposed method is better than that of the frequency-based estimator.

Keywords: radial velocity estimation, ambiguity, wideband radar system.

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1. Introduction

In modern wideband radar systems, high-precision motion parameter estimation of targets is crucial, which attracts much more attention, and can be used both in tracking and inverse synthetic aperture radar (ISAR) imaging. Since the range resolution for wideband linear frequency modulation (LFM) signals is relatively high, compared with the narrow band signals, the ISAR imaging performance can be improved [1 – 8]. High-precision motion parameter estimation is also very important to the micro-Doppler extraction for automatic target recognition (ATR) [9 – 12]. In general, the parameter estimation methods for wideband radars can be divided into two categories: parameter-searching methods and non-parameter-searching methods.

The parameter-searching method uses some criteria to maximize or minimize the objective function, by searching the parameter space of each motion parameter. A large number of traditional methods fall into this category. For instance, in [13], the radar echo signal is projected onto a set of basis functions composed of different parameters. Then, the estimated parameters are obtained by maximizing the projection values. The correlation function between two echo signals’ envelopes with respect to different delays was introduced in [14]. The amount of envelope shift is determined by searching the maximum of the correlation function. In order to achieve high-precision motion parameter estimation, the Radon-Fourier transform (RFT) methods based on multi-pulse energy accumulation were used in [15 – 17].

On the other hand, the parameters to be estimated do not need to be traversed for the non-parameter-searching methods and can be directly estimated. In [18], a novel motion parameter estimation method based on keystone transform (KT) was proposed, in which the adjacent cross-correlation function (ACCF) was defined. The ACCF based method is further studied in [19 – 22]. In the modified methods, the KT is no longer required, and the choice of scaling factors has been optimized. As a result, the algorithm performance is improved, especially under lower signal-to-noise ratio (SNR) conditions. In [23], a fast algorithm based on the cross-correlation of adjacent echoes (CCAE) was proposed. The frequency of the cross-correlation results was used to estimate the velocity of the target. This method can be directly applied in radar systems using the stretched signals as probing waveform, due to its efficiency and ease of use.

Through further investigation, we find that the phase of the cross-correlation output is also related to the target’s velocity, and furthermore the precision obtained from the phase estimation is higher than the CCAE estimator. Therefore, we use the phase term to estimate the target velocity value to obtain a higher precision estimate. Simulation results show that the proposed method is effective and can be used in real wideband LFM radars.

The rest of this paper is organized as follows. In Sec-
tion 2, the signal model and the stretch processing in wideband radar systems are introduced. The CCAE method proposed in [23] is briefly introduced in Section 3, and the cross-correlation results are derived. In Section 4, the proposed algorithm for solving phase ambiguity is described in detail. In Section 5, the theoretical performance of root-mean-square error (RMSE) is derived. In Section 6, the simulation results under different conditions are presented. Finally, some conclusions are given in Section 7.

2. Signal model

In order to obtain a high range resolution, a wideband radar system transmits LFM signals with a large time-bandwidth product. The transmitted signal can be modeled as

\[ s_{tr}(t) = \text{rect}\left( \frac{t}{T} \right) \exp(j\pi\gamma t^2) \exp(j2\pi f_c t) \]  

where

\[ \text{rect}(u) = \begin{cases} 1, & |u| \leqslant 1/2 \\ 0, & \text{otherwise} \end{cases} \]

denotes the rectangle pulse shape, \( T \) is the pulse width and \( f_c \) is the radar center frequency. \( \gamma = B/T \) is the chirp rate of the LFM signal, where \( B \) is the swept bandwidth.

In this paper, the scatterer model of the wideband LFM signal is used. Since the range resolution of the wideband radar is high, the scatterers of a target will distribute in many range cells. The radar periodically transmits the LFM pulse signal, the pulse echoes reflected by the individual scatterers are superposed to form the target’s echo signal, which can be expressed as

\[ s_{re}(t, t_m) = \sum_{p=1}^{N} \text{rect}\left( \frac{t - \tau_p(t_m)}{T} \right) A_p \cdot \exp\{j\pi\gamma(t - \tau_p(t_m))^2\} \exp\{j\pi f_c(t - \tau_p(t_m))\} + \omega_0(t, t_m) \]  

where \( t_m = mT_r \) is the slow time, \( m \) is the pulse number, and \( T_r \) is the pulse repetition interval. \( A_p \) denotes the reflection coefficient of the \( p \)th scatterer, \( N \) is the number of scatterers and \( \tau_p(t_m) \) is the time delay from the radar to the \( p \)th scatterer. \( \omega_0 \) is an additive white Gaussian noise with its mean and variance being 0 and \( \sigma^2 \) respectively. The time delay \( \tau_p(t_m) \) is proportional to the instantaneous distance at time \( t_m \) between the radar and the scatterer, which can be expressed as

\[ \tau_p(t_m) = \frac{2R_p(t_m)}{c}, \]  

where \( c \) is the speed of light.

In this paper, the target’s rotation is not taken into consideration, so that the displacement of each scattering point between two adjacent pulses is equal. Thus the relationship of the distance and time delay change between different scatterers satisfy

\[ \begin{align*}
R_p(t_{m+1}) - R_p(t_m) &= R_q(t_{m+1}) - R_q(t_m) \\
\tau_p(t_{m+1}) - \tau_p(t_m) &= \tau_q(t_{m+1}) - \tau_q(t_m)
\end{align*} \]  

where \( p, q \in [1, N] \).

Since the interval between two adjacent pulses is commonly short, the instantaneous velocity of the target at time \( t_m \) can be approximated as

\[ v(t_m) = \frac{R_p(t_{m+1}) - R_p(t_m)}{T_r}. \]  

Substituting (3) into (5), we have

\[ v(t_m) = \frac{c(\tau_p(t_{m+1}) - \tau_p(t_m))}{2T_r}. \]  

To reduce the bandwidth of the echo signals, the dechirp operation is commonly used in wideband radar systems, where the echo signals are mixed with the local reference LFM signal, which can be written as

\[ s_{LO}(t) = \text{rect}\left( \frac{t}{T} \right) \exp\{-j(2\pi f_c t + \pi\gamma t^2)\} \]  

where \( \bar{T} \) is the duration of the reference signal. To cover the received signal, the duration of the transmitted signal is usually slightly larger than that of the received signal. The stretched signal can be expressed as

\[ s_{st}(t, t_m) = s_{re}(t, t_m) \cdot s_{LO}(t) = \sum_{p=1}^{N} \text{rect}\left( \frac{t - \tau_p(t_m)}{T} \right) A_p \exp(-j2\pi\gamma\tau_p(t_m)t) \cdot \\
\exp(-j2\pi f_c \tau_p(t_m)) \exp(j\pi\gamma\tau_p(t_m)^2) + \omega_1(t, t_m) \]  

where \( \omega_1(t, t_m) = \omega_0(t, t_m) \cdot s_{LO}(t) \) is the noise after stretching operation. From (8), it is observed that the stretched signal is formed by the superposition of several single-frequency signals. The frequency of each sinusoidal signal is determined by the delay \( \tau_p(t_m) \). After stretching operation, the bandwidth of the signal is greatly reduced. Therefore, the sampling rate can be greatly reduced. It is noted that with the stretch operation, not only the amount of the subsequent calculation is reduced, but also the SNR of the time-domain received signals increases.

The motion information of the target can be observed from the frequency and the phase of the echoes, thus the motion parameters can be estimated using suitable algorithms. In the following, the CCAE estimator [23] is first introduced briefly and then an improved algorithm is proposed.

3. Velocity estimation based on phase

In [23], a novel fast non-parameter-searching method based on CCAE for wideband LFM signals was proposed.
To calculate the time delay difference between the adjacent echo signals, the cross-correlation operation is performed, which can be expressed as

\[ s_{ac}(t, t_m) = s_{st}(t, t_m) \cdot \text{conj}(s_{st}(t, t_{m+1})). \]  \hfill (9)

By substituting (8) into (9), we obtain

\[ s_{ac}(t, t_m) = s_{se}(t, t_m) + s_{cr}(t, t_m) + \omega_2(t, t_m). \]  \hfill (10)

Detailed derivation of the cross-correlation function can be found in [23]. After that, we perform fast Fourier transform (FFT) operation on \( s_{ac}(t, t_m) \), the result can be written as

\[ S_{sc}(f, t_m) = \sum_{p=1}^{N} A_p^2 \text{sinc}\{T[f - \gamma(t_p(t_{m+1}) - t_p(t_m))]\} \cdot \exp\{j[2\pi f_c(t_p(t_{m+1}) - \tau_p(t_m)) + \pi\gamma(t_p^2(t_m) - t_p^2(t_{m+1}))]\}. \]  \hfill (11)

Thus the cross-term \( s_{cr}(t, t_m) \) will be accumulated, which is simplified as

\[ S_{sc}(f, t_m) = \left(\sum_{p=1}^{N} A_p^2\right) \cdot \text{sinc}\{T[f - \gamma(t_i(t_{m+1}) - \tau_i(t_m))]\} \cdot \exp\{j[2\pi f_c(t_i(t_{m+1}) - \tau_i(t_m)) + \pi\gamma(t_i^2(t_m) - t_i^2(t_{m+1}))]\}. \]  \hfill (12)

Since \( \tau_p^2 = 4R_p^2/\lambda^2 \) is small, the quadratic term \( \tau_i^2(t_{m+1}) - \tau_i^2(t_m) \) in (11) and (12) can be ignored. Since some of the energy of the cross-term are cancelled each other, the cross-term is much smaller than the self-term. Thus the cross-term \( s_{cr}(t, t_m) \) in (10) can be ignored. According to (4), we can see that the energy of the self-term in (11) will be accumulated, which is simplified as

\[ S_{se}(f, t_m) = \left(\sum_{p=1}^{N} A_p^2\right) \cdot \text{sinc}\{T[f - \gamma(t_i(t_{m+1}) - \tau_i(t_m))]\} \cdot \exp\{j[2\pi f_c(t_i(t_{m+1}) - \tau_i(t_m)) + \pi\gamma(t_i^2(t_m) - \tau_i^2(t_{m+1}))]\}. \]  \hfill (13)

where \( 1 \leq i \leq N \) is the scatterer index. The peak position of the spectrum of the self-term can be expressed as

\[ \hat{f}_m = \gamma(t_p(t_{m+1}) - \tau_p(t_m)). \]  \hfill (14)

By substituting (14) into (6), we have

\[ \hat{v}_m = c\hat{f}_m / 2\gamma T_r. \]  \hfill (15)

In [23], (15) is used for velocity estimation, i.e., the CCAE estimator. By observing (13), we find that the phase of the spectrum peak is also related to the velocity of the target and can also be used to estimate the velocity parameter. Moreover, as we know, the precision of the phase estimation is higher than that of the frequency. However, the phase estimators suffer from the \( 2\pi \) ambiguity problem, which needs to be solved before a phase estimator can be used.

When \( 2\pi f_c(\tau_p(t_{m+1}) - \tau(t_m)) < 2\pi \), the phase at the peak can be written as

\[ \hat{\phi}_m = 2\pi f_c(\tau_p(t_{m+1}) - \tau(t_m)). \]  \hfill (16)

By substituting (16) into (6), the velocity is estimated as

\[ \hat{v}_m = \frac{\lambda \hat{\phi}_m}{4\pi T_r}. \]  \hfill (17)

where \( \lambda = c / f_c \) is the wavelength of the transmitted signal.

However, when \( 2\pi f_c(\tau_p(t_{m+1}) - \tau_p(t_m)) > 2\pi \), the phase at the peak will be ambiguous, that is,

\[ \hat{\phi}_m = \text{mod}(2\pi f_c(\tau_p(t_{m+1}) - \tau_p(t_m)), 2\pi) = \text{mod}(\hat{\phi}_m, 2\pi) \]  \hfill (18)

where \( \hat{\phi}_m \) denotes the unambiguous phase. Unfortunately, the phase at the peak is generally ambiguous, due to the fact that the target velocity is large. An algorithm for solving the phase ambiguity is proposed in the next section.

After obtaining the unambiguous phase \( \tilde{\phi}_m \), we can estimate the velocity more precisely, which can be written as

\[ \tilde{\hat{v}}_m = \frac{\lambda \tilde{\phi}_m}{4\pi T_r} \]  \hfill (19)

### 4. Phase ambiguity resolving

In this section, a method of phase ambiguity resolving is introduced, and for convenience, the proposed algorithm is referred to as phase ambiguity resolving estimator (PARE). According to (18), it is clear that the relationship between \( \hat{\phi}_m \) and \( \tilde{\phi}_m \) is

\[ \begin{align*}
\hat{\phi}_m &= \text{mod}(\hat{\phi}_m, 2\pi) \\
\phi_m &= 2\pi n + \hat{\phi}_m
\end{align*} \]  \hfill (20)

where \( n \) denotes the ambiguity number, \( \hat{\phi}_m \) is the ambiguous phase of the spectrum and \( \phi_m \) is the desired unambiguous phase. Therefore, it is required to calculate the ambiguity number \( n \), and then \( \tilde{\phi}_m \) can be obtained according to (20).

As mentioned above, the precision of the CCAE estimator is coarse. However, it can be used to determine the
ambiguity number \( n \). Observing (14) and (19), it is shown that \( \hat{\phi}_m \) is related to \( \hat{f}_m \), which can be expressed as

\[
\hat{\phi}_{m\text{(rough)}} = \frac{2\pi f_s \hat{f}_m}{\gamma}.
\]  

(21)

Then the ambiguity number can be obtained by

\[
n = \left\lfloor \frac{\hat{\phi}_{m\text{(rough)}}}{2\pi} \right\rfloor
\]

(22)

where \( \lfloor z \rfloor \) means rounding \( z \) to the nearest integer in the direction of negative infinity. Substituting (21) and (22) into (20), we obtain the unambiguous phase

\[
\hat{\phi}_m = 2\pi \left( \frac{f_s \hat{f}_m}{\gamma} \right) + \hat{\phi}_m.
\]

(23)

Finally, substituting the unambiguous phase (23) into (19), we can obtain a more precise velocity estimation of the target. The implementation flow of the proposed algorithm can be summarized as the following steps.

(i) Perform cross-correlation operation on two adjacent stretched signals.

(ii) Perform FFT operation on the multiplication results for energy accumulation.

(iii) Estimate the frequency \( \hat{f}_m \) and the phase \( \hat{\phi}_m \) of the peak.

(iv) Calculate the ambiguity number \( n \) according to (21) and (22).

(v) Calculate the unambiguous phase \( \hat{\phi}_m \) according to (20).

(vi) Estimate the target velocity using the unambiguous phase according to (19).

5. Performance analysis

In this section, the theoretical RMSE of the PARE estimator is derived, compared with that of the CCAE estimator [23]. Supposing \( \omega_0 \) is white Gaussian noise with its mean and variance being 0 and \( \sigma^2 \) respectively, the SNR of \( s_{rc}(t, t_m) \) can be written as \( \text{SNR}_{re} = A^2 / \sigma^2 \), where \( A \) is the amplitude of the signal. The stretching operation will decrease the bandwidth of the signal. The sampling frequency to the dechirp signal is \( f_s \), then the noise bandwidth becomes \( f_s \) after stretching processing. Then the SNR of the stretched signal can be expressed as

\[
\text{SNR}_{st} = \frac{A^2}{f_s^2} \frac{B}{\sigma^2} = \frac{B}{f_s \sigma^2} \text{SNR}_{re}
\]

(24)

where \( B \) is the swept bandwidth.

Equation (24) implies that after stretching processing, the SNR of the signal increases by \( B/f_s \) times. In (10), \( \omega_2(t, t_m) \) represents the noise term in the cross-correlation function \( s_{ac}(t, t_m) \). In [23], it is proved that \( \omega_2(t, t_m) \) can be approximately viewed as a Gaussian noise with the mean 0, and the variance \( 2A^2 \sigma^2 + \sigma^4 f_s / B \). Thus, the SNR of the signal after cross-correlation operation can be written [24] as

\[
\text{SNR}_{ac} = \frac{A^4}{f_s f_s \left( 2A^2 \sigma^2 + \frac{f_s}{B} \sigma^4 \right)} = \frac{\text{SNR}_{st}}{2 + \text{SNR}_{st}}.
\]

(25)

For the PARE, the velocity of the target is calculated by estimating the phase of the spectrum peak in \( s_{ac}(t, t_m) \). Hence, this procedure is similar to estimating the initial phase of a complex sinusoidal signal. For a complex sinusoidal signal with its amplitude, phase and frequency unknown, the RMSE of the initial phase estimation can be written [18] as

\[
\text{RMSE}_\phi = \sqrt{\frac{1}{\text{SNR} \cdot N}}
\]

(26)

where \( N \) is the number of sampling points, and SNR is the SNR value of the sinusoidal signal. According to (19), we can obtain the RMSE of PARE as

\[
\text{RMSE}_v = \frac{\lambda}{4\pi T_f} \sqrt{\frac{1}{\text{SNR}_{ac} N}}.
\]

(27)

Then, substituting (24) and (25) into (27), the RMSE of the target velocity can be written as

\[
\text{RMSE}_v \approx \frac{c}{4\pi f_s T_f} \sqrt{\frac{2B f_s + f_s^2 / \text{SNR}_{re}}{\text{SNR}_{re} \cdot N B^2}}.
\]

(28)

In [23], the RMSE of CCAE is expressed as

\[
\text{RMSE}'_v = \frac{c}{4\pi B T_f} \sqrt{\frac{6(2B f_s + f_s^2 / \text{SNR}_{re})}{\text{SNR}_{re} \cdot N B^2}}.
\]

(29)

From (28) and (29), we can see that the RMSE of PARE is smaller than that of CCAE. In the following, the difference between the RMSEs of these two estimators is compared using the parameters of a typical radar system with its parameters listed in Table 1. Substituting these parameters into (28) and (29), the performances of the CCAE and PARE are shown in Fig. 1.

| Table 1 Radar parameters |
|---------------------------|
| Parameter | Value |
| Bandwidth/GHz | 1 |
| Sampling frequency/MHz | 10 |
| Pulse repetition interval/s | 0.02 |
| Wavelength/m | 0.033 |
| Number of sampling points | 2,000 |
It can be seen that the RMSE of the proposed algorithm is about 14 dB lower than that of the CCAE algorithm. The reason is that, the proposed method uses the phase information of the spectrum peak to correct the velocity estimation calculated from the frequency term, which improves the estimation precision.

6. Simulation results

To verify the performance of the proposed method, simulations are conducted and compared with that of the CCAE. The simulation parameters are shown in Table 2. The length of the target is about 10 m, and the target moves at a constant velocity of 80 m/s.

| Parameter                  | Value |
|----------------------------|-------|
| Center frequency/GHz       | 9     |
| Sampling frequency/MHz     | 10    |
| Bandwidth/GHz              | 1     |
| Pulse width/μs             | 200   |

Then we evaluate the performance of the PARE under different parameters. The SNRs increase from −30 dB to 5 dB, and 1 000 Monte Carlo simulations are performed for each SNR to generate the RMSEs. The simulation results are shown in Fig. 2–Fig. 4, where CRLB refers to Cramer-Rao lower bound.

It is shown that the RMSE of the PARE is about 14 dB lower than that of the CCAE when the SNR is higher than the threshold SNR, i.e., 0 dB, for the PARE. There are two SNR thresholds for the PARE, with the lower one, i.e., −8 dB, as the same as the SNR threshold for the CCAE. For convenience, the SNR threshold 0 dB is denoted as \( SNR_{Th1} \), and the SNR threshold −8 dB is denoted as \( SNR_{Th2} \). When the SNR is lower than \( SNR_{Th1} \), the precision of the CCAE may be poor and is not enough for the ambiguity resolving for the PARE. Thus the precisions of the PARE for the SNR between \( SNR_{Th1} \) and \( SNR_{Th2} \) are the same as those of the CCAE.
those of CCAE. Thus in real applications, PARE can be utilized directly. When the SNR is higher than $\text{SNR}_{\text{Th2}}$, the estimates come from PARE. While the SNR is lower than $\text{SNR}_{\text{Th2}}$, the accuracy will be close to CCAE. The users do not need to choose the estimator according to the SNR.

Actually, if we count the number of the bad estimates, we will find that the number is not very large. The number of bad estimates for each SNR is shown in Table 3. The estimation results with the SNR $-5$ dB are shown in Fig. 3, which shows that only a few estimates are ambiguous. In real applications, these bad estimates which are called outliers can be found and thrown away. Fig. 4 shows the performance of the PARE without the outliers when calculating the RMSEs. It is shown that the performance without outliers can reach the CRLB. In the tracking state, through filtering it is easy to find these outliers, which are easy to be removed from the estimates.

Table 3 Number of bad estimates for each SNR

| SNR/DB | -10 | -9 | -8 | -7 | -6 | -5 | -4 | -3 | -2 | -1 | 0 |
|--------|-----|----|----|----|----|----|----|----|----|----|---|
| Bad estimates number | 154 | 109 | 99 | 72 | 38 | 17 | 14 | 6  | 2  | 1  | 0 |

7. Conclusions

High precision radial velocity estimation is investigated in this paper, which is important in wideband radar systems. The frequency and the phase of the CCAE contain the radial velocity information of the target. A phase-based velocity estimator, i.e., PARE, is proposed and in order to resolve the ambiguity problem, the estimation of the CCAE is used as the initial estimation and then the ambiguity number is obtained. The ambiguity of the PARE estimation is then resolved using the ambiguity number. The performance analysis is performed and simulation results show that the performance of the proposed method outperforms the frequency-based estimator. However, since the SNR threshold of the PARE is higher than that of the CCAE, the PARE under low SNR conditions needs to be further investigated.

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