LETTER Special Section on Recent Progress in Antennas and Propagation in Conjunction with Main Topics of ISAP2010

Doppler Centroid Estimation for Space-Surface BiSAR

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SUMMARY Phase synchronization is a crucial problem in Bistatic Synthetic Aperture Radar (BiSAR). As phase synchronization error and Doppler phase have nearly the same form, Doppler Centroid (DC) cannot be estimated with traditional method in BiSAR. A DC estimation method is proposed through phase-interferometry of Dual-channel direct signal. Through phase interferometry, phase synchronization error can be counteracted while Doppler phase is reserved and DC can be estimated from the reserved phase.

key words: *BiSAR, Doppler centroid estimation, direct signal*

1. Introduction

Space-Surface BiSAR (SS-BiSAR) is composed by space-borne Synthetic Aperture Radar and a ground-nearby receiver [1], [2]. Direct signal is quite suited for phase synchronization as it has simple form and high SNR. A Doppler Centroid (DC) estimation method was proposed in [3] and this method makes use of direct signals and Ephemeris data. As an amount of prior information are needed, its application is limited and inflexible. A new method through processing dual-channel direct signals is proposed in this paper. Through applying phase interferometry to the special dual-channel direct signals, Phase Synchronization Error (PSE) is suppressed and DC can be estimated from the residual phase. For this method, only few prior parameters are necessary to estimate DC and therefore, the proposed method can be applied widely.

2. DC Estimation through Dual-Channel Direct Signal

2.1 System Description

SS-BiSAR is composed of a space-borne transmitter and a ground-nearby receiver placed on a balloon, aircraft or tower. It has unsymmetrical architecture as there is great difference between transmitter and receiver in altitude and velocity.

2.2 Signal Model for BiSAR

BiSAR is coherent radar system. Monostatic and bistatic coherent radar are both composed of transmitter and receiver. Main difference is whether antenna, frequency synthesizer and timer are shared. Structure differences result in synchronization problem. Although there are different structures, echo of point target can be expressed by Eq. (1) for monostatic and bistatic SAR.

$$r(\eta, t) = u(t - \tau(\eta)) \exp(j\theta(\eta)) \tag{1}$$

In Eq. (1) t, η, $\tau(n)$, $u(t)$ and $\theta(n)$ denote fast-time, slow-time, transfer delay, baseband signal and baseband phase. According to the operating principle of SAR, $\theta(n)$ can be expressed by

$$\theta(\eta) = \varphi_c(t - \tau(\eta)) - \varphi_{LO}(t) \tag{2}$$

In Eq. (2) $\varphi_c(t)$ and $\varphi_{LO}(t)$ are transient phases of carrier and local oscillator. In ideal condition $\varphi_c(t)$ and $\varphi_{LO}(t)$ vary linearly with time and can be expressed by Eq. (3) and Eq. (4):

$$\varphi_c(t) = 2\pi f_c t + \varphi_c(T_0) \tag{3}$$

$$\varphi_{LO}(t) = 2\pi f_{LO} t + \varphi_{LO}(T_0) \tag{4}$$

In Eq. (3) and Eq. (4) f_c and f_{LO} are nominal frequency of carrier and local oscillator and $\varphi_c(T_0)$ and $\varphi_{LO}(T_0)$ are their initial phases. In monostatic radar f_c and f_{LO} can be supposed to equal so $\theta(n) = -2\pi f_c \tau(n)$. While in bistatic radar frequency error needs to be considered, Eq. (3) and Eq. (4) need to be modified as Eq. (5) and Eq. (6) [4]:

$$\varphi_c(t) = 2\pi \int_{T_0}^{t} \left(f_c + f_e^T(\zeta)\right) d\zeta + \varphi_c(T_0) \tag{5}$$

$$\varphi_{LO}(t) = 2\pi \left(\int_{T_0}^{t} f_{LO} + f_e^R(\zeta)\right) d\zeta + \varphi_{LO}(T_0) \tag{6}$$

f_e^T and f_e^R are time-varient frequency deviation. According to Eq. (5) and Eq. (6), in BiSAR $\theta(n)$ can be denoted by

$$\begin{aligned}\theta(\eta) &= -2\pi f_c \tau(\eta) + 2\pi \int_{T_0}^{t-\tau(\eta)} f_e^T(\zeta) d\zeta \\ &\quad - 2\pi \int_{T_0}^{t} f_e^R(\zeta) d\zeta + \varphi_c(T_0) - \varphi_{LO}(T_0)\end{aligned} \tag{7}$$

In Eq. (7) the first term is Doppler phase and the others are Phase Synchronization Error (PSE) and PSE $\theta_e(\eta)$ can be denoted by

$$\theta_e(\eta) = 2\pi \int_{T_0}^{t-\tau(\eta)} f_e^T(\zeta) d\zeta - 2\pi \int_{T_0}^{t} f_e^R(\zeta) d\zeta + \varphi_e(T_0) \tag{8}$$

Generally primary part of frequency deviation is constant

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bias so $\theta_e(\eta)$ can be approximated to be linear.

2.3 DC Estimation

As PSE has similar form with Doppler linear phase, traditional DC estimation method becomes invalid before PSE is compensated. PSE and Doppler phase have different origins: Doppler phase comes from the relative movement between radar and target while PSE is introduced by frequency deviation of carrier and local oscillator. Apparently Doppler Phase has close relation with space while PSE depends on time. Basing on this characteristic, a DC estimation method utilizing dual-channel direct signal is proposed. In this method, dual-channel direct signals are received at given place. Through phase interference processing in time-domain, PSE is cancelled and Doppler phase is reserved. DC can be obtained from the residual phase. In following content, relation between residual phase and DC is derived.

Firstly reference frame is founded. X-axis is the direction of radar velocity, Z-axis is perpendicular with ground and Y-axis composes a right-hand frame with X and Z just as shows. According to the transferring relation between orbit velocity and radar velocity, slant range history of space-borne SAR can also be approximated by hyperbola and orbit velocity should be replaced by radar velocity [5]. In Fig. 1, initial position of transmitter is $(0, 0, H)$, transmitter is H high and moves along the direction of x-axis with speed of v_r.

In SS-BiSAR, slant range can be expressed by Eq. (9) and for direct signal there is $R_R(x_0, y_0, 0) = 0$:

$$R_{sum}(\eta) = \sqrt{(v_r\eta - x_0)^2 + y_0^2 + H^2} + R_R(x_0, y_0, 0) \quad (9)$$

As Fig. 1 shows, two antennas are deployed along a line paralleling with flying path of transmitter with coordinates of $(x_0, y_0, 0)$ and $(x_0 + L, y_0, 0)$. The nearest distance between antenna and flying path is R_0 and distance between the two antennas is L. Assuming effective velocity of transmitter is v_r and wave beam's moving speed is v_b. Doppler phase histories of the two direct signals satisfy translational-invariant condition and interval of two direct signals' aperture centre moments is $\Delta\eta = L/v_b$ [5].

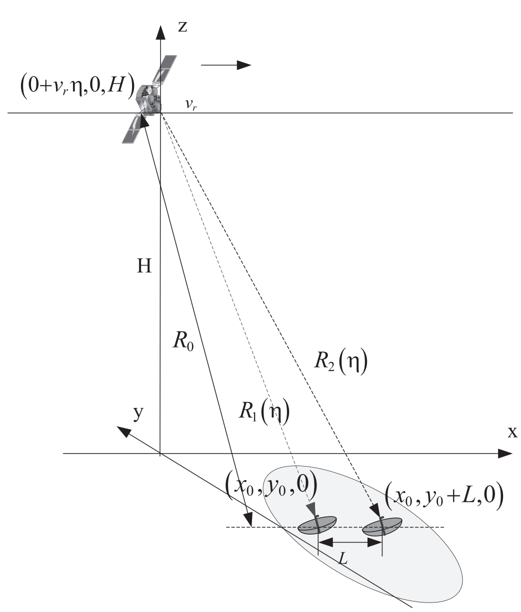

Fig. 1 Geometry of direct signal receiving.

From Fig. 1 transfer delay for the two direct signals can be expressed by Eq. (10) and Eq. (11):

$$\tau_1(\eta) = \frac{R_1(\eta)}{c} = \frac{\sqrt{(v_r\eta - x_0)^2 + y_0^2 + H^2}}{c} \quad (10)$$

$$\tau_2(\eta) = \frac{R_2(\eta)}{c} = \frac{R_1(\eta - \Delta\eta)}{c} = \frac{\sqrt{(v_r\eta - L - x_0)^2 + y_0^2 + H^2}}{c} \quad (11)$$

As $\tau_1(\eta)$ and $\tau_2(\eta)$ are nearly the same, PSEs are nearly the same according to Eq. (8) while Doppler phases are different. Through phase interference, PSE is eliminated while Doppler phase is reserved.

Phase histories of two direct signals can be obtained after direct signals are compressed and difference of two phase histories will be Eq. (12):

$$\theta_{in}(\eta) = -2\pi f_c(\tau_1(\eta) - \tau_2(\eta)) + \theta_{e1}(\eta) - \theta_{e2}(\eta) \quad (12)$$

From Eq. (8) it can be known that $\theta_{e1}(\eta) \approx \theta_{e2}(\eta)$ so Eq. (12) can be simplified to

$$\theta_{in}(\eta) \approx -2\pi f_c(\tau_1(\eta) - \tau_2(\eta)) = -2\pi f_c \Delta\tau(\eta) \quad (13)$$

In Eq. (13) $\Delta\tau(\eta) = \tau_1(\eta) - \tau_2(\eta)$. Generally Doppler phase history can be approximated by 2-order polynomial expansion and two direct signals can be expressed by Eq. (14) and Eq. (15):

$$\theta_{d1}(\eta) = -2\pi f_c \tau_1(\eta) = \theta_0 + 2\pi f_{dc}\eta + \pi f_{dr}\eta^2 \quad (14)$$

$$\theta_{d2}(\eta) = -2\pi f_c \tau_2(\eta) = \theta_{d1}(\eta - \Delta\eta) \quad (15)$$

In Eq. (14) f_{dc} and f_{dr} are DC and Doppler FM rate. Obviously $\theta_{in}(\eta)$ can be approximately denoted by

$$\begin{aligned}\theta_{in}(\eta) &= \theta_{d1}(\eta) - \theta_{d2}(\eta) \\ &\approx 2\pi f_{dc}\Delta\eta + 2\pi\Delta\eta f_{dr}\eta - \pi f_{dr}\Delta\eta^2\end{aligned} \quad (16)$$

In Eq. (16) $\theta_{in}(\eta)$ is composed by three terms: $2\pi f_{dc}\Delta\eta$ and $-\pi f_{dr}\Delta\eta^2$ are constant and $2\pi\Delta\eta f_{dr}\eta$ is linear. $\Delta\eta$ can be calculated when L and v_b are known. What's more, $2\pi\Delta\eta f_{dr}\eta$ can be obtained if f_{dr} is estimated. In fact PSE is dominated by linear phase and has little impact on Doppler FM rate, so f_{dr} can be easily estimated accurately through 2-order polynomial fitting the Doppler phase history as SNR of direct signal is quite high. When Δt and f_{dr} are known, f_{dc} can be calculated by Eq. (16).

Generally DC is quite high in space-borne SAR system and Doppler ambiguity will always occur when yaw steering isn't applied. In order to obtain unambiguous DC, phase

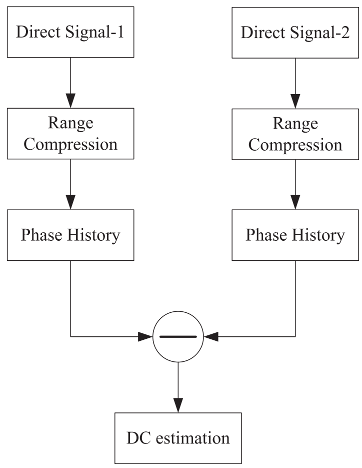

Fig. 2 Flow chart of dual-channel direct interference.

in Eq. (16) must be unwrapped and Δt must satisfy the following condition:

$$2\pi f_{dc}\Delta\eta \leq 2\pi \Rightarrow \Delta\eta \leq \frac{1}{f_{dc}} \quad (17)$$

In Eq. (17), the upper boundary of $\Delta\eta$ can be estimated from the prior information of DC. When Eq. (17) is satisfied, DC can be calculated by the following equation:

$$f_{dc} = \frac{\theta_{in}(\eta) - 2\pi\Delta\eta f_{dr}\eta + \pi f_{dr}\Delta\eta^2}{2\pi\Delta\eta} \quad (18)$$

2.4 DC Estimation Procedure

The procedure of the proposed DC estimation method is shown in Fig. 2 and this procedure includes three parts: dual-channel direct signal range compression and phase estimation, phase interference and DC estimation.

3. Performance Analysis

3.1 Precision Analysis

From Eq. (18), errors of f_{dr} and $\Delta\eta$ impact on the estimation precision of DC. As f_{dr} can be estimated accurately, DC estimation error is mainly caused by the error of $\Delta\eta$. Assume that the measurement of $\Delta\eta$ is $\Delta\hat{\eta}$, then error of $\Delta\eta$ is $\Delta\eta_e = \Delta\eta - \Delta\hat{\eta}$. According to the DC calculation equation, DC estimation error is

$$\varepsilon = \hat{f}_{dc} - f_{dc} = \frac{\theta_{in}(\eta)\Delta\eta_e}{2\pi(\Delta\eta - \Delta\eta_e)\Delta\eta} - \frac{f_{dr}\Delta\eta_e}{2} \quad (19)$$

As $\Delta\eta = L/v_b$, $\Delta\eta_e$ is mainly caused by the errors of space between antenna phase centers (hereinafter referred to as baseline) and v_b. In SS-BiSAR, satellite velocity can be calculated accurately from orbit elements, so baseline measurement error leads to main cause of $\Delta\eta_e$.

As shown in Fig. 3, A and B denote the phase centers of two antennas and length of $\overline{AB}$ is L. When $\overline{AB}$ is unparalleled to the flying direction or baseline measurement error exists, it can be known from the geometry relation of Fig. 3 that the effective length of baseline is

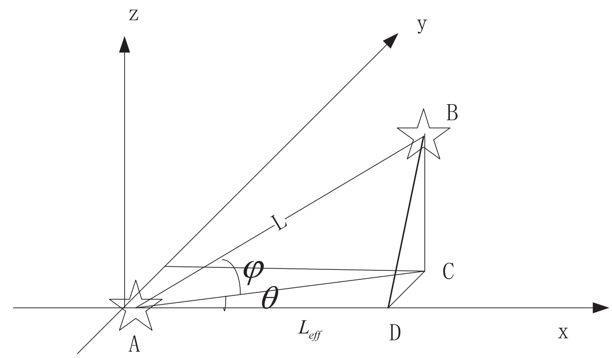

Fig. 3 Coordinate system of receiving antennas.

$$L_{eff} = |AD| = |AB|\cos\varphi\cos\theta = L\cos\varphi\cos\theta \quad (20)$$

When measurement value of the baseline is $\hat{L}$, $\Delta\eta_e$ can be expressed by

$$\Delta\eta_e = \frac{\left|\hat{L} - L_{eff}\right|}{v_b} = \frac{\left|\hat{L} - L\cos\varphi\cos\theta\right|}{v_b} \quad (21)$$

From Eq. (21), it can be known that the length and angle errors of the baseline will influence DC estimation precision. In an interferometric SAR, through precision machining and angle measurement, the length and angle errors of the baseline can be less than 1 cm and 1°, repectively. Under the above condition, the DC estimation precision is analyzed by numeral simulation. In simulation the well-known radar satellite Radarsat-I is chosen as illuminator and a stationary receiver is adopted. From the typical parameters of Radarsat-I echo, corresponding PRF, f_{dc} and f_{dr} are 1256.98 Hz, 2600 Hz and −866.5 Hz/s, respectively, in SS-BiSAR configuration. Simulation results are shown in Fig. 4 and Fig. 5. Figure 4 shows the DC estimation error under different baseline measurement errors at different baseline lengths. Figure 5 demonstrates the simulation result of DC estimation error caused by different angle measurement errors. From simulation results, it can be known that baseline measurement error is primary cause of DC estimation error and maximum DC estimation error is less than 30 Hz. The DC estimation precision can satisfy the imaging requirement, i.e., the DC error should be less than 5% PRF [5].

In proposed method, direct signals are received by a multi-channel receiver and phase inconsistency among different channels will influence DC estimation. Through making use of multi-channel radar testing technique phase inconsistency can be tested and well compensated, therefore impact of phase inconsistency between receiving-channels can be eliminated.

Thermal noise will also influence the final estimation precision. In SS-BiSAR system with a space-borne radar as illuminator, SNR of direct signals is quite high and can be improved after range compression, so impacts of thermal

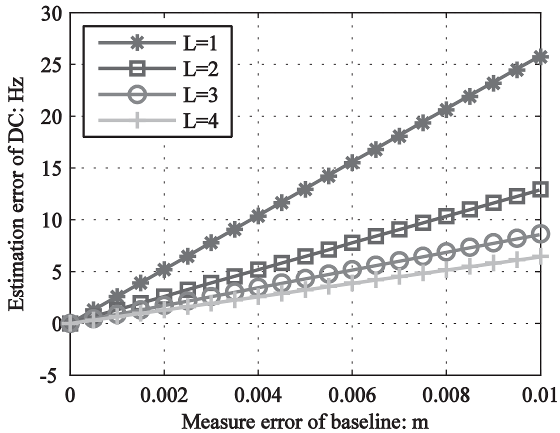

Fig. 4 DC error introduced by baseline error.

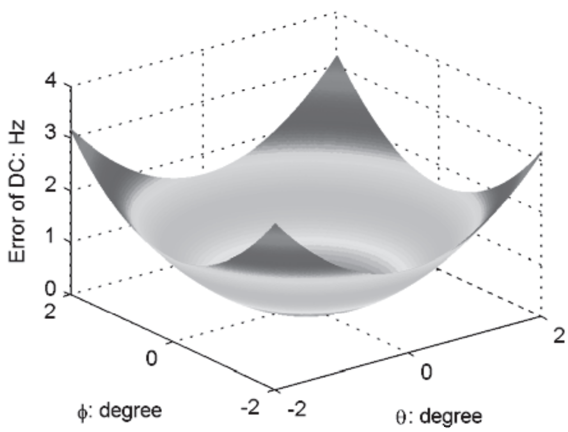

Fig. 5 DC error introduced by baseline direction error.

noise can be ignored.

3.2 Suppression of Phase Noise

As phase noise is difficult to suppress with general filtering method. The methodology introduced in [3] is sensitive to phase noise so high performance oscillator is needed to guarantee the method valid. However, the methodology proposed in this paper can well suppress the phase noise as phase error corresponding to phase noise will be cancelled by phase interference. The block diagram is shown in Fig. 6.

Assuming $\varphi_n(\eta)$ is the phase error introduced by phase noise, after synchronization processing, residual phase error is $\varphi_n(t-\tau_1(\eta)) - \varphi_n(t-\tau_2(\eta))$. As $\tau_1(\eta)$ and $\tau_2(\eta)$ vary with time, synchronization system is a time-variant system. However it can be approximated to be time-invariant during a pulse width as $\tau_1(\eta)$ and $\tau_2(\eta)$ are nearly changeless during this interval and frequency response can be expressed by

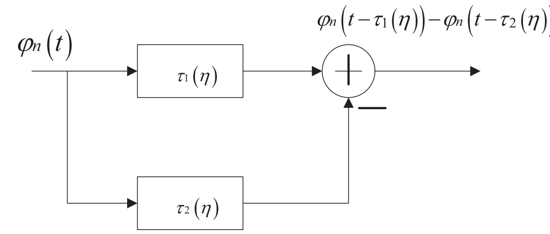

Fig. 6 Filter processing of phase noise.

$$H(f,\eta) = e^{-j2\pi f\tau_1(\eta)}\left(1 - e^{j2\pi f\Delta\tau(\eta)}\right) \quad (22)$$

It can be seen that this system is a delay system cascaded by a band-pass filter group. The center of pass band is

$$f_0(k) = \frac{k}{\tau_1(\eta) - \tau_2(\eta)} \quad (23)$$

As $\tau_1(\eta)$ and $\tau_2(\eta)$ vary with slow time, pass band will also change but low-frequency region will always be suppressed. As power spectrum of phase noise concentrates low-frequency region, phase noise will always be suppressed. It is obvious that the proposed methodology can accommodate severe phase noise condition.

4. Conclusion

Traditional DC estimation method becomes invalid in BiSAR as PSE exists. Through receiving dual-channel direct signal, DC can be estimated from the phase difference between the two direct signals. This method only needs little information about the space-borne radar and it is easy to implement. An important advantage of this method is that it is insensitive to phase noise because phase noise will be effectively suppressed and it will give high precision under severe phase noise case.

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