Design of a coherent inverse synthetic aperture radar moving target simulator

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Abstract: Generally, the time delay of simulated pulses generated by inverse synthetic aperture radar (ISAR) moving target simulator (MTS) is changing discretely with time. In this case, there must be time delay errors between simulated pulses and real pulses reflected by real target. As a result, the coherence of the simulated pulses will be worsened and the ISAR image would be blurred due to the time delay errors. In order to solve the problem, a compensating technique consisting of two steps is proposed. Firstly, if the simulated pulse and the real pulse are located at different range cells, move the simulated pulse into the range cells at which the real pulse are located. Secondly, compensate the phase errors between simulated pulse and real pulse. After step two, at the efficient sampling time instants, the digital samples corresponding to the simulated pulse are equal to that constructed from the real pulse. So the radar will treat the simulated target as a real target. In other words, the coherence and the ISAR image quality of the simulated target are guaranteed. The simulation results demonstrate the efficiency of the proposed technique.

Keywords: moving target simulator, inverse synthetic aperture radar, time delay errors, time delay errors compensation

Classification: Microwave and millimeter wave devices, circuits, and systems

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1 Introduction

The ISAR radar is of paramount importance to military applications. The ISAR MTS plays an important role in the development of ISAR radar. Because the ISAR MTS can reduce the cost, shorten the development time, and easy the testing procedure. Besides, the ISAR MTS can be utilized as an active jamming source in some military applications, such as military maneuver.

In the recent decade, many diverse ISAR echoes generating methods have been presented. In [1], Digital Image Synthesizer (DIS) is employed to create false target images to deceive ISAR radar. In [1, 2], Digital Radio Frequency Memory (DRFM) is used to generate the simulated ISAR echoes. The methods mentioned above [1, 2, 3] are considered as active retransmission jamming and are mainly used in the scenarios of electronic countermeasure (ECM). In scenarios of radar testing, such as [4], the ISAR MTS is used to test the performance of ISAR radar and the MTS does not receive any signal. The ISAR MTS transmits simulated echoes on its own. Besides, computer simulation methods, such as [5, 6], are employed to establish the target models for ISAR applications. In ISAR imaging, the time delay of simulated echoes plays an important and crucial role in ISAR imaging. Because the Doppler information is directly related to the time delay value of echoes. In reality, the time delay of real target is changing continuously. While in case of ISAR MTS, such as [2, 3, 4], the time delay of simulated echoes varies discretely. The impact of discrete time delay on the coherence of simulated echoes is firstly discussed in [7] in the case of DRFM. Then more detailed discussion of the coherence of the simulated echoes is given in [8]. Due to the discrete time delay, the simulated echoes are not very coherent and the ISAR image would be blurred. But none of them [2, 3, 4, 7, 8] had proposed a feasible method of solving the problem caused by the discrete time delay of the ISAR MTS.

However, in this paper, we propose a compensating technique by which the problem caused by the discrete time delay can be solved. It consists of two steps. Firstly, because of the discrete time delay of the MTS, the simulated pulse generated by the MTS and the real pulse reflected by the real target may be located at different range cells. In this case, the MTS should move the simulated pulse into the range cells at which the real pulse are located. This process is denoted as range cell errors compensation. Secondly, although the two pulses are located at the same range cells, there may be phase errors between them. So the MTS must compensate the phase errors so that the digital samples from the simulated pulse are equal to that from the real pulse at the efficient sampling time instants. After step two, with respect to the radar, there is no difference between simulated target and real target. So the simulated target will equivalently move continuously like real target dose

and the coherence of simulated pulses is guaranteed. The method is demonstrated by the simulation results.

2 Architecture of the ISAR MTS system

Fig. 1. The block diagram of the simulation system.

The block diagram of the ISAR MTS system is shown in Fig. 1. The whole system including radar and ISAR MTS is controlled by a common ultra-high stable local oscillator. Its phase noise can be ignored. Therefore, all the synthesized clocks or frequencies in the system (both radar and ISAR MTS) have the same time base. This is the foundation of the proposed method shown below. The transmitter of the radar is shut down and there is no receiver in the MTS. That is, the MTS does not receive any signal transmitted from radar. The MTS generates the required simulated target according to the control information from the radar. The simulated pulses are transmitted from the antenna of the MTS at discrete time instants. The whole system can be placed in the microwave anechoic chamber and the MTS is situated at the far field of the radar. Unlike simulated pulses, the real pulses are reflected by real targets, such as real airplanes, satellites and missiles. In reality, it is a hard work and costs a lot to use real targets to test the radar performance in the early phase of radar development. Therefore, the MTS is employed to create the required simulated targets instead of real targets. In this paper, the terms simulated pulse, simulated echo and simulated target are used interchangeably to mean the same thing which are generated by the MTS. On the other hand, the terms real pulse, real echo and real target are also used interchangeably to mean the same thing.

3 Echo model of the real target

Suppose that the radar is chirp pulse-based. The transmitted linear modulation frequency (LFM) waveforms with pulse repetition interval (PRI) T_r are written as

$$s_t(\hat{t}, t_m) \sim \text{rect}\left(\frac{\hat{t}}{T_p}\right) e^{j\pi\mu\hat{t}^2} e^{j2\pi f_c t} = p(\hat{t}) e^{j2\pi f_c t} \quad (1)$$

where $p(\hat{t}) = \text{rect}(\hat{t}/T_p) e^{j\pi\mu\hat{t}^2}$ is the complex envelope, f_c is the center frequency, μ is the LFM coefficient, T_p is the pulse width, $t = t_m + \hat{t}$ is the full time, $\hat{t}$ is the fast time, $t_m = mT_r$ is the slow time (azimuth time) and m is the pulse number variable. For convenience, a target consisting of perfect point scatterers of equal reflectivity is chosen and it is situated at the far field of the radar. The received signal from a point scatterer at R away from the radar is given by

$$s_r(\hat{t}, t_m) \sim p\left(\hat{t} - \frac{2R}{c}\right) e^{j2\pi f_c(t - \frac{2R}{c})} \quad (2)$$

After coherent processing (down-converter), Eq. (2) is converted to base-band signal

$$s_{rb}(\hat{t}, t_m) \sim p(\hat{t} - \tau) e^{-j2\pi f_c \tau} \quad (3)$$

where $\tau = 2R/c$ is the real time delay between transmitted pulse and returned pulse. For a moving target with respect to radar, τ is changing continuously with time. Eq. (3) represents the echo model of the point scatterer on the real target.

4 Echo model of the simulated target with time delay errors

Due to the fact that digital delay line is widely utilized in MTS engineering, the simulated time delay of the simulated pulses generated by the MTS is discrete and it is the quantization of the real time delay τ in Eq. (3). So there are quantization errors between the simulated time delay and real time delay. The coherence of the simulated pulses is heavily affected by the magnitude of the quantization errors [7, 8]. Suppose that the step size (time delay resolution) of the simulated time delay is $T_{dl} = 1/f_{dl}$, then the simulated time delay is written as

$$\bar{\tau} = \left\lceil \frac{\tau}{T_{dl}} \right\rceil T_{dl} = DT_{dl} \geq \tau \quad (4)$$

where $\lceil \cdot \rceil$ denotes the ceiling operator and the integer $D = \lceil \tau/T_{dl} \rceil$. According to Eq. (4), the simulated pulse is transmitted by the MTS at starting time instant $\bar{\tau}$. So the time delay errors between real time delay and simulated time delay are written as

$$\Delta\tau = \bar{\tau} - \tau \quad (5)$$

The time delay error $\Delta\tau$ is a non-random variable, since both $\bar{\tau}$ and τ are deterministic variables with respect to MTS. Therefore, the compensation procedure shown below is easy to implement. Due to the time delay errors in Eq. (5), the simulated echo model of the simulated target corresponding to Eq. (3) is rewritten as

$$s_{rb}(\hat{t}, t_m) \sim p(\hat{t} - \bar{\tau}) e^{-j2\pi f_c \bar{\tau}} = p(\hat{t} - DT_{dl}) e^{-j2\pi f_c \tau} \quad (6)$$

As seen in Eq. (6), the time delay value in the complex envelope $p(\hat{t} - \bar{\tau})$ is not equal to the one in the phase factor $e^{-j2\pi f_c \tau}$. Since Eq. (5) is not always zero, the pulse repetition frequency (PRF) of simulated pulses will not vary like that of real target and will jitter around the real value. This will have an adverse effect on the coherence of simulated pulses. As a result, the Doppler information cannot be extracted correctly. So the coherent processing for the range-Doppler ISAR imaging will be affected and the ISAR images may be blurred and defocused. However, this adverse effect can be eliminated by using the proposed technique presented below.

5 Compensation of the time delay errors

Generally, a fine resolution digital delay line is applied to the MTS. So we can suppose that the step size of the simulated time delay is smaller than the sampling interval of the ADC in radar. That is, $T_{dl} < T_s$, where T_s is the sample rate of the digital receiver in radar. In the radar coherent integration time, the range gate is held still. According to Eq. (3) and Eq. (6), the position relationship between real pulse and simulated pulse in the range gate is shown in Fig. 2.

Fig. 2. The position relationship between real pulse and simulated pulse.

In Fig. 2, the shape of the two pulses is parabola which represents the quadratic phase of the chirp waveform. From Fig. 2 we can clearly see that, the two pulses are spaced by the time delay error $\Delta\tau$. Due to this error, they may be located at different range cells and their phases are different at the efficient sampling time instants. Fig. 2 shows the case that the two pulses are located at different range cells. Fig. 2 summarizes the impact of time delay errors on the coherence in terms of range cell errors and phase errors. So a compensating algorithm consisting of two steps is presented:

Step one: If the two pulses are located at different range cells (as shown in Fig. 2), move the simulated pulse along the time axis by several time delay steps so as to let them be located at the same range cells, as shown in Fig. 3. Otherwise, skip this step.

Step two: Compensate the phase of simulated pulse. As a result, at the efficient sampling time instants, the phase of simulated pulse is equal to that of real pulse. In other words, the two pulses are superposed by each other at the efficient sampling time instants.

In step one, the range cell error is given by

$$\Delta r = \left\lfloor \frac{\bar{\tau}}{T_s} \right\rfloor - \left\lfloor \frac{\tau}{T_s} \right\rfloor = \left\lfloor \frac{DT_{dl}}{T_s} \right\rfloor - \left\lfloor \frac{\tau}{T_s} \right\rfloor \geq 0 \quad (7)$$

where $\lfloor \cdot \rfloor$ denotes the flooring operator. In Fig. 2, $\Delta r = 1$. The purpose of step one is to let $\Delta r = 0$. The compensating process of step one is given by

$$\Delta r = \left\lfloor \frac{\bar{\tau} - KT_{dl}}{T_s} \right\rfloor - \left\lfloor \frac{\tau}{T_s} \right\rfloor = \left\lfloor \frac{(D - K)T_{dl}}{T_s} \right\rfloor - \left\lfloor \frac{\tau}{T_s} \right\rfloor = \left\lfloor \frac{\bar{\tau}'}{T_s} \right\rfloor - \left\lfloor \frac{\tau}{T_s} \right\rfloor = 0 \quad (8)$$

where $\bar{\tau}' = (D - K)T_{dl}$ and K is the total number of time delay steps by which the MTS must move the simulated pulse along the time axis. Eq. (8) is depicted in Fig. 3.

Fig. 3. The position relationship after compensating the range cells errors.

In Fig. 3, we can see that, after step one, the two pulses are located at the same range cells and the residual time delay errors are rewritten as

$$\Delta\tau' = \bar{\tau}' - \tau = (D - K)T_{dl} - \tau = LT_{dl} - \tau \leq T_{dl} < T_s \quad (9)$$

Where $L = D - K$. Like Eq. (5), $\Delta\tau'$ is also a non-random quantity, and

$$\lim_{T_{dl} \rightarrow 0} \Delta\tau' = 0 \quad (10)$$

In fact, the time delay errors expressed in Eq. (9) cannot be removed thoroughly with regard to the practical limitation $T_{dl} > 0$. According to Eq. (9), Eq. (6) can be rewritten as

$$\begin{aligned} s_{rb}(\hat{t}, t_m) &\sim p(\hat{t} - \bar{\tau}') e^{-j2\pi f_c \tau} \\ &= \text{rect}\left(\frac{\hat{t} - \bar{\tau}'}{T_p}\right) e^{j\pi\mu(\hat{t} - \tau)^2} e^{-j2\pi f_c \tau} e^{j\pi\mu[-2(\hat{t} - \tau)\Delta\tau' + (\Delta\tau')^2]} \\ &= \text{rect}\left(\frac{\hat{t} - \bar{\tau}'}{T_p}\right) e^{j\pi\mu(\hat{t} - \tau)^2} e^{-j2\pi f_c \tau} \varphi \end{aligned} \quad (11)$$

where

$$\varphi = e^{j\pi\mu[-2(\hat{t} - \tau)\Delta\tau' + (\Delta\tau')^2]} \approx e^{j2\pi\mu(\Delta\tau')\tau} \quad (12)$$

Eq. (11) is the echo model of simulated pulse after compensating the range cell errors. Compare Eq. (11) with Eq. (3) we can find out that the phase factor φ is the reason that the coherence of the simulated pulses is worsened. As a result, the Doppler information in the exponential $e^{-j2\pi f_c \tau}$ cannot be extracted correctly. Based on Eq. (10), the limit of Eq. (12) when $T_{dl} \rightarrow 0$ is $\varphi = 1$. So the limit of Eq. (11) is given by

$$\lim_{T_{dl} \rightarrow 0} p(\hat{t} - \bar{\tau}') e^{-j2\pi f_c \tau} = p(\hat{t} - \tau) e^{-j2\pi f_c \tau} \quad (13)$$

From Eq. (13), we can see that Eq. (11) is equal to Eq. (3) when $T_{dl} \rightarrow 0$. It means that φ can be ignored when T_{dl} is small enough. How far the phase factor φ should be taken account is determined by practical requirement. According to radar engineering experience [9], the random initial phase of each pulse affects the coherence of radar system. So the magnitude of the random initial phase must be small enough to guarantee the coherence. Although the phase in Eq. (12) is a non-random variable, its values corresponding to different pulses are different. Therefore, it can be regarded as additional initial phase added to each simulated pulse. Therefore, in order to guarantee the coherence of simulated pulses, the phase in φ must be small enough.

Now consider the worst situation in Eq. (9). That is, $\Delta\tau' = T_{dl}$. In this case, Eq. (12) is rewritten as $\varphi \approx e^{j2\pi\mu T_{dl}\tau}$. Set $\eta = \mu T_{dl}\tau$. So the condition of ignoring φ is that the quantity η must satisfy the following inequality

$$\eta \ll \frac{\alpha}{360^\circ} \quad (14)$$

where α is the maximum random initial phase in degrees of a coherent system. As an example, define $\mu = 20$ MHz/us, $\tau = 200$ us and $f_{dl} = 50$ GHz. Then η is 0.08. In this case, the maximum additional initial phase caused by Eq. (9) is about $0.08 * 360 = 28.8$ degrees. So the maximum random initial phase $\alpha \geq 28.8$ degrees. The coherent radar system cannot tolerate such a value. Suppose that the maximum

random initial phase of a coherent system cannot exceed 1 degrees, then, $\eta \leq 1/360$ and $f_{dl} \geq 1440$ GHz. It is obvious that such a high frequency is impossible to be achieved in MTS. Therefore, in reality, the phase factor φ cannot be ignored when Eq. (14) is not true.

In order to remove the adverse impact of Eq. (12) on the coherence, the step two, phase compensation is proposed to guarantee the coherence of the ISAR MTS system. Firstly, set a phase compensation factor

$$\phi = (\varphi)^* = e^{-j\pi\mu[-2(\hat{t}-\tau)\Delta\tau' + (\Delta\tau')^2]} \approx e^{-j2\pi\mu(\Delta\tau')\tau} = e^{-j2\pi\mu(LT_{dl}-\tau)\tau} \quad (15)$$

In Eq. (15), we can see that all the variables, such as L , τ , μ and T_{dl} , are all deterministic quantities. They are all known with respect to ISAR MTS. In practical usage, the fast time $\hat{t}$ and the high order item $(\Delta\tau')^2$ in Eq. (15) can be ignored. So the phase factor φ and the phase compensation factor ϕ are easy to measure. Then the step two can be done by multiplying Eq. (15) by Eq. (11)

$$\begin{aligned} s_{rb}(\hat{t}, t_m) &\sim \text{rect}\left(\frac{\hat{t} - \bar{\tau}'}{T_p}\right) e^{j\pi\mu(\hat{t} - \bar{\tau}')^2} e^{-j2\pi f_c \tau} \phi \\ &= \text{rect}\left(\frac{\hat{t} - \bar{\tau}'}{T_p}\right) e^{j\pi\mu(\hat{t} - \tau)^2} e^{-j2\pi f_c \tau} \end{aligned} \quad (16)$$

Eq. (16) is the final echo model of the ISAR MTS and it is depicted in Fig. 4. Compare Fig. 4 with Fig. 3, one can see that at the efficient sampling time instants, the two pulses are superposed by each other and their phases are equal to each other. So the sampled data delivered to the DSP (Digital Signal Processor) in radar from the simulated target is the same as that from the real target. So the radar cannot distinguish the two targets and the simulated target will be treated as a real target. In other words, the ISAR MTS with discrete time delay can equivalently generate a continuously moving target by using the technique presented in this paper.

Fig. 4. The result after compensating the phase errors.

In order to implement Eq. (16), replace the quantity $\bar{\tau}'$ with LT_{dl} and rewrite it in another form

$$s_{rb}(\hat{t}, t_m) \sim \text{rect}\left(\frac{\hat{t} - LT_{dl}}{T_p}\right) e^{j\pi\mu(\hat{t} - LT_{dl})^2} (e^{-j2\pi f_c \tau} \phi) \quad (17)$$

In Eq. (17), the phase factor φ is concealed in the exponential $e^{j\pi\mu(\hat{t} - LT_{dl})^2}$. For pulse m , the integer L can be obtained according to Eq. (4) and Eq. (8). Then set a digital counter with initial value equal to L . The counter is counting down at f_{dl} . During the counter running, the DSP in MTS computes the total initial phase factor $e^{-j2\pi f_c \tau} \phi$ and other parameters for pulse m . These parameters are sent to the DDS (Direct Digital Synthesizer) in the transmitter of MTS. Once the counter is

timing out, a trigger is issued to the transmitter immediately and a pulse is radiated in space, just as shown in Fig. 1.

6 Simulation setup and results

The supposed range-Doppler ISAR imaging scenario [10] is illustrated in Fig. 5(a). On the 2D coordinate system, the radar is situated at the origin and the targets center is located at some (x_0, y_0) point. The target is assumed to be composed of a number of perfect point scatterers of equal reflectivity. Their locations in the 2D coordinate system are plotted in Fig. 5(b). The target is flying along the x direction at a constant velocity of v_x . The target and radar parameters with respect to the scenario are listed in Table I.

Fig. 5. (a) ISAR imaging scenario. (b) Fictitious target of perfect point scatterers of equal reflectivity.

Table I. Target and radar parameters

parameter name	symbol	value
target parameters		
initial position in x	x_0	0 m
initial position in y	y_0	24 km
velocity along x	v_x	125 m/s
radar parameters		
center frequency	f_c	20 GHz
frequency bandwidth	B	2 GHz
sampling frequency	f_s	4 GHz
pulse width	T_p	0.5 us
pulse repetition frequency	PRF	6000 Hz
ISAR MTS parameters		
time delay resolution	$T_{dl} (f_{dl})$	20 ps (50 GHz)
signal-to-noise ratio	SNR	-17 dB

From Fig. 5(a), we can see that the distance, R , is changing with time. For some (x, y) point scatterer, it can be written as

$$R = ((x_0 + x - v_x t)^2 + (y_0 + y)^2)^{1/2} \quad (18)$$

In real-world scenarios, it is important to take additive noise into consideration. Therefore, the total received signal is the summation of returned signal plus the additive noise $n(t)$ as

$$h_{rb}(\hat{t}, t_m) = s_{rb}(\hat{t}, t_m) + n(t) \quad (19)$$

The additive noise is white Gaussian and the SNR is chosen to be -17 dB which is not good. But after applying the range compression, the range profile can be clearly observed, as shown in Fig. 6(a), Fig. 7(a) and Fig. 8(a). Due to the high PRF rate of 6000 Hz, the range profiles are well aligned, as depicted in Fig. 6(a), Fig. 7(a) and Fig. 8(a). Therefore, no range alignment algorithms is needed before applying the azimuth compression. The final ISAR image is obtained by applying a Fast Fourier Transform (FFT) operation along the azimuth dimension. In order to suppress the side lobes, hamming window function is chosen for ISAR imaging. The resultant ISAR images corresponding to Eq. (3), Eq. (11) and Eq. (17) are plotted in Fig. 6(b), Fig. 7(b) and Fig. 8(b), respectively.

On the range-Doppler plane as depicted in Fig. 6(b), the point scatterers on the real target are well-resolved in both directions. In Fig. 7(b), we can see that, in the range direction, the point scatterers on the simulated target can be resolved. But in the azimuth dimension, they cannot be resolved. It is because that the time delay errors affect the extraction of Doppler information of each point scatterer. In Eq. (11), the middle exponential $e^{-j2\pi f_c \tau}$ determines the Doppler frequency shift values of different point scatterers. Due to Eq. (12), the coherence of received signal is destroyed and the Doppler frequency information for each scatterer cannot be extracted exactly. Therefore, the final ISAR image cannot be obtained by

Fig. 6. (a) Range compressed data of real target. (b) ISAR image of real target.

Fig. 7. (a) Range compressed data of simulated target. (b) ISAR image of simulated target.

Fig. 8. (a) Range compressed data of simulated target after error compensation. (b) ISAR image of simulated target after error compensation.

Fig. 9. ISAR image of simulated target before time delay errors compensation. (a) $f_{dl} = 3600$ GHz. (b) $f_{dl} = 10000$ GHz.

applying a FFT operation along the pulse index and the image is blurred seriously. After utilizing the proposed phase compensation algorithm, in Fig. 8(b), we can see that the ISAR image is as good as that plotted in Fig. 6(b). Thanks to the phase compensation factor ϕ , the Doppler information is extracted correctly and the point scatterers on the simulated target are well-resolved in both dimensions. This demonstrates the efficiency of the proposed compensating algorithm. In other words, the ISAR MTS can simulate a continuously moving target. Besides, Fig. 9 is plotted to intuitively explain that, for a given scenario, how to choose the parameter f_{dl} when Eq. (12) can be ignored. From Fig. 9(a) one can see that when $f_{dl} = 3600$ GHz, though the plane can be observed, the ISAR image of simulated target is serious blurred. From Fig. 9(b), we can find out that even if f_{dl} is very high, the ISAR image is not as good as that in Fig. 8(b). So the best way of guaranteeing the coherence is to use the method presented in this paper.

7 Conclusion

We propose a method of eliminating the adverse impact of time delay errors on the coherence of simulated pulses and the quality of ISAR image. The method consists of two steps. The first one is range cell errors compensation and the other is phase errors compensation. After applying the proposed method, the problems caused by the time delay errors can be removed. Therefore, the ISAR MTS can generate a real target which is moving continuously for radar. The simulation results demonstrate the method.