Wideband RF Self-Interference Cancellation for FMCW Phased-Array Radar

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ABSTRACT In this paper, a segmented self-interference cancellation (SSIC) method is proposed for frequency-modulated continuous-wave phased-array radar for canceling wideband multipath self-interference (SI) signals. A radio frequency canceller based on the SSIC method that uses the time-frequency segmentation equivalence of the linear frequency-modulated (LFM) signal is designed that can convert wideband SI cancellation (SIC) into several narrowband SICs. Numerical and simulation results show that the proposed method can achieve better cancellation performance with lower hardware complexity than the conventional multi-tap SI cancellation (MTSIC) method. Furthermore, the superior performance of the SSIC method is verified by experiments with a 250 MHz LFM signal.

INDEX TERMS Frequency-modulated continuous wave, linear frequency modulated, phased array, self-interference cancellation, wideband.

I. INTRODUCTION

FREQUENCY-modulated continuous-wave (FMCW) phased-array radar is widely used for automotive applications such as intelligent cruise control and collision warning/avoidance because it can provide information on the distance and velocity of multiple targets and estimate their azimuth and elevation angles [1], [2]. Due to the simultaneous transmission and reception at the same frequency, however, its performance significantly suffers from strong multipath self-interference (SI), including the direct-path component and the clutter-path components reflected by environmental objects both with shared antennas and separate antennas [3]–[6]. In general, the SI signal is much stronger than the echo signal reflected from the target, the sidelobes of which are strongly coupled in range and Doppler in the application of the matched filter technique, thereby affecting the correct reception of the echo signal and, more importantly, saturating the low noise amplifier of the receiver [4]. An effective way to combat the above impacts is to cancel the SI signal at the receiver’s radio frequency (RF) front end.

Most previous research focuses on the cancellation of the leakage power caused by circulators for shared antennas [7]–[9]. For multipath SI composed of clutter, cancellation is usually carried out by an RF canceller with a multi-tap structure (also referred to as an adaptive filter) that consists of delayers, attenuators and phase shifters to emulate the actual wireless channel [10]–[16]. However, a larger signal bandwidth and more complex channel environments mean that more taps are needed for SI cancellation (SIC), which causes the following two problems. First, an input signal...
with a higher power is needed for SIC due to the power loss incurred by splitting the signal among the taps; Second, as shown in Fig. 1, each receive array needs an RF canceller, and the hardware complexity (the number of analog devices) increases proportionally with the number of taps, thus resulting in excessive economic budgets.

Due to the restrictions mentioned above, it is difficult to reach a compromise between the cancellation performance and hardware complexity for wideband SIC in FMCW phased-array radar. In terms of the linear frequency-modulated (LFM) signal extensively applied in FMCW radar, its frequency changes linearly with time [3]. Therefore, a wideband LFM signal with a bandwidth of \( B \) and a duration of \( T \) can be equally divided into \( K \) narrowband LFM signals with a bandwidth of \( B/K \) and a duration of \( T/K \) in different time segments by time-domain segmentation, which is called time-frequency segmentation equivalence in this paper. Based on this feature, we propose a segmented SIC (SSIC) method that cancels the time-domain segmented SI signals in different time segments with the same RF canceller. The SSIC method converts the wideband SIC into several narrowband SICs and outperforms the conventional multistatic SIC (MTSIC) method. The contributions of this paper are summarized as follows:

1) An RF canceller based on the proposed SSIC method is built for wideband LFM SIC, along with an estimation algorithm for the amplitudes and phases of the canceller.
2) The closed-form expressions of the cancellation performance of the SSIC method are formulated by minimizing the residual SI energy.
3) The performance of the SSIC method is verified by simulations and experiments, which both show that this method can achieve better cancellation performance with lower hardware complexity than the conventional MTSIC method, especially for the wideband scenario.

The rest of this paper is organized as follows. The system model with RF SIC for FMCW phased-array radar is described in Section II. In Section III, an RF canceller based on the proposed SSIC method, including the architecture of the RF canceller and the steps of the RF SIC, is presented, and the closed-form expressions of the cancellation performance of the SSIC method are formulated, followed by a comparison with the performance of the conventional MTSIC method. Additionally, the setup time for perfect cancellation of the SSIC method is discussed in detail. In Section IV, the simulation results and the experimental verification are presented. Finally, conclusions are presented in Section V.

II. SYSTEM MODEL WITH RF SIC FOR FMCW PHASED-ARRAY RADAR

As shown in Fig. 1, the RF canceller established between the transmit and receive chains has a reference source from the power amplifier (PA) output, which includes all the imperfections (nonlinearity and noise) of the transmit chain. Therefore, the transmitted signal after the PA can be given by

\[
s'_T(t) = s(t) + s'_{nl}(t) + n_T(t),
\]

where \( s(t) \) and \( s'_{nl}(t) \) are the linear and nonlinear components of \( s_T(t) \), respectively, and \( n_T(t) \) is the noise in the transmitted signal.

In (1), the linear component \( s(t) \) of the LMF signal used for transmission can be written as

\[
s(t) = \sqrt{2P} \cos(2\pi f_c t + \pi \mu t^2), -T/2 \leq t \leq T/2,
\]

where \( T \) is the sweep repetition interval, \( \mu \) is the chirp rate, \( f_c \) is the central frequency, and \( P \) is the transmit power. Moreover, the signal bandwidth \( B = \mu T \).

A transmitted signal with high power will leak to the receiver through the SI channel, thereby forming an SI signal. Let \( u(t) = \bar{s}_T(t) e^{-j2\pi f_c t} \) be the equivalent lowpass signal of \( s_T(t) \), where \( \bar{s}_T(t) = s_T(t) + j\tilde{s}_T(t) \) is the analytic signal of \( s_T(t) \) and \( \tilde{s}_T(t) = s_T(t) \ast (1/\pi t) \) is the Hilbert transform of \( s_T(t) \). Assuming that the SI channel with multiple propagation paths is a linear time-invariant system, the received SI signal is the sum of the light-of-sight (LOS) component and all the resolvable multipath components [17], that is,

\[
r_{sl}(t) = s_T(t) \ast h_{st}(t)
\]

\[
= \text{Re}\left\{\sum_{m=1}^{M} a_m u(t - \tau_m) e^{j[2\pi f_c (t - \tau_m) + \theta_m]}\right\}
\]

\[
= \sum_{m=1}^{M} a_m [s_T(t - \tau_m) \cos \theta_m - \bar{s}_T(t - \tau_m) \sin \theta_m],
\]

where \( \tau_m \) and \( a_m \) are the delay and amplitude of the \( m \)-th path, respectively, \( \theta_m \) is the phase shift caused by the Doppler frequency shift, \( M \) is the number of paths, \( \ast \) denotes the convolution operation, and \( \text{Re}(\cdot) \) denotes the real part of (\( \cdot \)).

From (3), the real bandpass impulse and frequency response of the SI channel can be written as

\[
h_{st}(t) = \sum_{m=1}^{M} a_m \delta(t - \tau_m) \cos \theta_m.
\]

\[
H_{st}(f) = \begin{cases} 
\sum_{m=1}^{M} a_m e^{-j(2\pi f \tau_m - \theta_m)}, & f_1 \leq f \leq f_2 \\
\sum_{m=1}^{M} a_m e^{-j(2\pi f \tau_m + \theta_m)}, & -f_2 \leq f \leq -f_1 
\end{cases}
\]

where the multiplication factor 0.5 is omitted, which will not have any effect on subsequent analysis, \( f_1 = f_c - B/2 \) and \( f_2 = f_c + B/2 \).

Considering the echo signal \( r_{echo}(t) \) reflected by the target, the received signal at the front-end of the receiver can be given by

\[
r(t) = r_{sl}(t) + r_{echo}(t) + n_R(t),
\]

where \( n_R(t) = n_{sp}(t) + n_i(t) \) is the noise of the received signal, \( n_{sp}(t) \) is the receiver noise floor, and \( n_i(t) \) is the additive noise of the SI channel.
During the RF SIC, the reference signal coupled from the PA output is fed into the RF canceller to form the reconstructed SI signal, which can be written as

\[ r_c(t) = s_T(t) * h_c(t) + n_c(t), \]

where \( h_c(t) \) is the impulse response of the RF canceller and \( n_c(t) \) is the noise of the reconstructed SI signal. Note that the power loss caused by the coupler is omitted here because it can be regarded as a part of \( h_c(t) \).

To perform the RF SIC at the receiver’s RF front-end, \( r_c(t) \) is subtracted from \( r(t) \), and the residual signal after cancellation is given by

\[ r_{sic}(t) = r(t) - r_c(t) = r_{res}(t) + r_{echo}(t) + n_{res}(t), \]

where \( n_{res}(t) = n_c(t) - n_c(t) \) is the noise of the residual signal and \( r_{res}(t) = s_T(t) * [h_{si}(t) - h_c(t)] \) is the residual SI signal due to imperfect cancellation.

III. RF SIC BASED ON THE PROPOSED SSIC METHOD

In [18], a reconfigurable bandpass filter (BPF) is inserted in each tap to divide the wideband SI into several narrowband SIs, which are then canceled. Although an SI signal with a smaller bandwidth is easier to cancel, designing the high-Q RF BPFs required for this method is difficult and costly.

The specific steps of the SSIC method are described as follows, where the power loss due to the power divider and the coupler is omitted.

**Step 1:** A power divider is used to split the reference signal \( s_T(t) \) coupled from the PA output into \( L + 1 \) branches, one branch of which and the received signal coupled from the low noise amplifier (LNA) are sent to the intermediate frequency (IF) mixer and sampled by the analog-to-digital converter (ADC) to get the discrete-time samples of reference signal \( s_T[n] \) and received signal \( r[n] \), \( n = 1, 2, \ldots, N \), where \( N \) is the number of samples.

**Step 2:** The discrete-time reference and received signal are fed into the parameter control module to calculate the amplitudes \( \hat{\alpha}_1, \hat{\alpha}_2, \ldots, \hat{\alpha}_L \) and the phases \( \hat{\theta}_1, \hat{\theta}_2, \ldots, \hat{\theta}_L \) of the taps, \( k = 1, 2, \ldots, K \), where \( K \) is the number of time segments in the sweep repetition interval, and \( L \) is the number of taps.

**Step 3:** The remaining \( L \) branches in step 1 are fed into \( L \) delays to generate the delayed reference signals \( s_T(t - \tau_1), s_T(t - \tau_2), \ldots, s_T(t - \tau_L), \) where the delay \( \tau_1 \) is estimated according to the distance of the antenna arrays and the actual channel environment.

**Step 4:** In the \( k \)-th time segment of the current sweep repetition interval, configure the attenuators and phase shifters of the analog taps with amplitudes \( \hat{\alpha}_1, \hat{\alpha}_2, \ldots, \hat{\alpha}_L \) and phases \( \hat{\theta}_1, \hat{\theta}_2, \ldots, \hat{\theta}_L \) respectively, and superimpose the time delay signals after amplitude and phase adjustment to generate the reconstructed SI signal \( r_{res}(t) \).

**Step 5:** In the \( k \)-th time segment, the RF SIC is performed by subtracting \( r_{res}(t) \) from the received signal.

**Step 6:** Repeat steps 2~5 until \( k = K \) to complete the RF SIC in the current sweep repetition interval, and continue to repeat steps 1~6 in each subsequent sweep repetition interval.

The architecture of the parameter control module described in step 2 is shown in Fig. 3(b). Moreover, we also...
present the algorithm used in the parameter control module in Algorithm 1.

**Algorithm 1 Algorithm for Parameter Control**

**Input:** The discrete-time reference signal $s_{r}[n]$ and received signal $r[n]$, $n = 1, 2, \ldots, N$.

**Output:** The amplitudes $a_{1}^{k}, a_{2}^{k}, \ldots, a_{L}^{k}$ and the phases $\hat{\theta}_{1}^{k}, \hat{\theta}_{2}^{k}, \ldots, \hat{\theta}_{L}^{k}$, $k = 1, 2, \ldots, K$.

1. $s_{T}^{k} = \{s_{r}[(k-1)P + 1], \ldots, s_{r}[kP]\}^{T}$, and $r_{k} = \{r[(k-1)P + 1], \ldots, r[kP]\}^{T}$ represent the reference vectors and the received vectors after time-domain segmentation, respectively, where $k = 1, 2, \ldots, K$, and $P = N/K$.

2. for $k = 1$ to $K$

3. Construct the delayed vectors $s_{T}^{k}, r_{T}^{k}$, $s_{T}^{k}, r_{T}^{k}$, \ldots, $s_{T}^{k}, r_{T}^{k}$ in $\mathbb{R}^{P \times 1}$ according to the delays $\tilde{\tau}_{1}, \tilde{\tau}_{2}, \ldots, \tilde{\tau}_{L}$;

4. Construct the delayed matrix $Q_{k} = [s_{T}^{k}, s_{T}^{k}, \ldots, s_{T}^{k}]_{\tilde{\tau}_{1}, \tilde{\tau}_{2}, \ldots, \tilde{\tau}_{L}}$;

5. Perform the Hilbert transform on $Q_{k}$ and $r_{k}$ to generate $\hat{Q}_{k} \in \mathbb{C}^{P \times L}$ and $\hat{r}_{k} \in \mathbb{C}^{P \times 1}$;

6. Compute the analytic matrix $Q_{k} = Q_{k} + j\hat{Q}_{k} \in \mathbb{C}^{P \times L}$ and the analytic vector $\hat{r}_{k} = r_{k} + jr_{k} \in \mathbb{C}^{P \times 1}$;

7. Compute the Moore-Penrose pseudoinverse $\hat{Q}_{k} = (\hat{Q}_{k}^{H}Q_{k}^{H})^{-1}Q_{k}^{H} \in \mathbb{C}^{L \times P}$, where $(\cdot)^{H}$ denotes the Hermitian transpose of $(\cdot)$;

8. Compute the weight vector $w_{k} = \hat{Q}_{k}^{H}\hat{r}_{k} \in \mathbb{C}^{L \times 1}$;

9. Determine $\hat{a}_{1}^{k}, \hat{a}_{2}^{k}, \ldots, \hat{a}_{L}^{k}$ and $\hat{\theta}_{1}^{k}, \hat{\theta}_{2}^{k}, \ldots, \hat{\theta}_{L}^{k}$ according to the amplitudes and phases of the elements in $w_{k}$, respectively;

10. end for

11. return $\hat{a}_{1}^{k}, \hat{a}_{2}^{k}, \ldots, \hat{a}_{L}^{k}$ and $\hat{\theta}_{1}^{k}, \hat{\theta}_{2}^{k}, \ldots, \hat{\theta}_{L}^{k}$, $k = 1, 2, \ldots, K$.

### B. PERFORMANCE ANALYSIS

For convenience, the notations used in this subsection are listed as follows. $x^{k}(t)$ denotes the $k$-th segmented signal after time-domain segmentation for $x(t)$, $E_{x}$ is the energy of $x(t)$, and $E_{x}^{k}$ is the energy of $x^{k}(t)$.

First, the reference signal and the received signal are divided into $K$ segments with an equal length in chronological order within the time $T$. Let $T$ be the length of each time segment, $T_{k}$ be the $k$-th time segment, and $B_{k}$ be the corresponding frequency segment, $k = 1, 2, \ldots, K$.

After the time-domain segmentation, the multipath features of the segmented SI signal have not changed. Therefore, the $k$-th SI signal can be given by

$$r_{res}^{k}(t) = s_{T}^{k}(t) \ast h_{si}(t).$$ (9)

The $k$-th reconstructed SI signal can be written as

$$r_{k}^{k}(t) = s_{T}^{k}(t) \ast h_{si}^{k}(t) + n_{c}^{k}(t),$$ (10)

where $h_{c}^{k}(t)$ is the impulse response of the RF canceller for the $k$-th SI signal cancellation and the corresponding frequency response is expressed as

$$H_{c}^{k}(f) = \left\{ \begin{array}{ll}
\sum_{l=1}^{L} \alpha_{l}^{k} e^{-j(2\pi f \tau_{l} - \hat{\theta}_{l})} = x^{T}w_{k}, & f \geq 0 \\
\sum_{l=1}^{L} \alpha_{l}^{k} e^{-j(2\pi f \tau_{l} + \hat{\theta}_{l})} = x^{T}w_{k}, & f \leq 0
\end{array} \right.$$ (11)

where

$$x = [ e^{-j2\pi f \tau_{1}}, e^{-j2\pi f \tau_{2}}, \ldots, e^{-j2\pi f \tau_{L}} ]^{T}.$$ (12)

$$w_{k} = [ \alpha_{1}^{k} e^{j\hat{\theta}_{1}}^{k}, \alpha_{2}^{k} e^{j\hat{\theta}_{2}}^{k}, \ldots, \alpha_{L}^{k} e^{j\hat{\theta}_{L}}^{k} ]^{T}.$$ (13)

The $k$-th residual SI signal after cancellation can be obtained by subtracting $r_{k}^{k}(t)$ from $s_{T}^{k}(t)$, that is,

$$r_{res}^{k}(t) = s_{T}^{k}(t) \ast [ h_{si}(t) - h_{c}^{k}(t) ].$$ (14)

To assess the cancellation performance, we define the cancellation ratio (CR), i.e., the ratio of the SI signal energy before cancellation to that after cancellation, as

$$G(\text{dB}) = 10 \log \frac{E_{res}}{E_{res} + E_{n_{res}}}.$$ (15)

where $\log(\cdot)$ denotes the base-10 logarithm.

Given that $n_{r}(t), n_{t}(t)$ and $n_{c}(t)$ are independent narrowband Gaussian white noise with bilateral power spectral densities of $N_{f}/2, N_{c}/2$ and $N_{c}/2$, respectively, the energy of $n_{res}(t)$ within the time $T$ can be given by

$$E_{n_{res}} = (N_{r} + N_{c})BT.$$ (16)

In this paper, since we consider only the in-band cancellation performance and the nonlinearity caused by the higher-order intermodulated components is out-of-band, the nonlinear component $s_{n}^{k}(t)$ can be omitted in (1).

The specific derivation of the energy of the $k$-th residual SI signal is presented in (17), in which the cross term between the noise and signal is omitted because they are independent, and the spectrum of $s(t)$ with a large time-bandwidth product can be approximated as [19]

$$S(f) \approx \left\{ \begin{array}{ll}
\frac{P}{2\mu} e^{-|f|^{2}(1 - f^{2})^{2} - \frac{f^{2}}{\mu}} |f|, & f_{1} \leq f \leq f_{2} \\
\frac{P}{2\mu} e^{-|f|^{2}(1 - f^{2})^{2} - \frac{f^{2}}{\mu}} \frac{f^{2}}{\mu}, & -f_{2} \leq f \leq -f_{1}
\end{array} \right.$$ (18)

Thus, the energy spectrum $|S(f)|^{2}$ can be approximated as a constant $\lambda = P/2\mu$, and the expressions of the matrices $R$, $R_{k} \in \mathbb{C}^{L \times L}$, the vectors $z, z_{k} \in \mathbb{C}^{1 \times 1}$ and the scalars $\eta_{si}, \eta_{si}^{k}$ are respectively given by

$$R = \int_{f_{c} - \frac{B}{2}}^{f_{c} + \frac{B}{2}} x^{*}x^{T} df = \sum_{k=1}^{K} \int_{B_{k}}^{K} x^{*}x^{T} df = \sum_{k=1}^{K} R_{k}.$$ (19)
\[ E_{r_{res}}^k = \int_{T_k} |r_{r_{res}}^k(t)|^2 dt = \int_{T_k} |s^k(t) + [h_{si}(t) - h^k_c(t)]|^2 dt + \int_{T_k} |n^k_{su}(t) + [h_{si}(t) - h^k_c(t)]|^2 dt \]

\[
\text{Parseval} = 2|S(f)|^2 \int_{B_k} |H_{si}(f) - H^k_c(f)|^2 df + N_r T \int_{f_c + \frac{B_r}{2}}^{f_c - \frac{B_r}{2}} |H_{si}(f) - H^k_c(f)|^2 df
\]

\[ = 2\lambda(\eta^k_{si} + w^H_k R_k w_k - w^H_k z_k - z^H_k w_k) + N_r T (\eta^k_{si} + w^H_k R_k w_k - w^H_k z - z^H w_k). \]  

\[ z = \int_{f_c - \frac{B_r}{2}}^{f_c + \frac{B_r}{2}} H_{si}(f)x^* df = \sum_{k=1}^{K} \int_{B_k} |H_{si}(f)x^* df = \sum_{k=1}^{K} z_k. \]  

\[ \eta^k_{si} = \int_{f_c - \frac{B_r}{2}}^{f_c + \frac{B_r}{2}} |H_{si}(f)|^2 df = \sum_{k=1}^{K} |H_{si}(f)|^2 df = \sum_{k=1}^{K} \eta^k_{si}. \]  

When the delays \( t^k_1, t^k_2, \ldots, t^k_B \) have been determined, the aim of SSIC is to minimize the energy of \( r_{r_{res}}^k(t) \) by adjusting the amplitudes and phases of the taps.

From (17), \( E_{r_{res}}^k \) is a quadratic function with respect to \( w_k \), and \( R_k \) and \( R \) in (19) are positive semidefinite matrices. Thus, \( E_{r_{res}}^k \) is a convex function whose minimum value can be obtained by setting its gradient to 0. Since \( w_k \) is a complex vector, the complex gradient vector of \( E_{r_{res}}^k \) can be given by

\[ \nabla E_{r_{res}}^k = \frac{\partial E_{r_{res}}^k}{\partial R}(w_k) + j \frac{\partial E_{r_{res}}^k}{\partial \text{Im}(w_k)} = 2\lambda(R_k w_k - z_k) + N_r T(R_k w_k - z). \]  

The minimum norm solution of the equation \( \nabla E_{r_{res}}^k = 0 \) gives the optimal value of \( w_k \) as

\[ w^{opt}_k = (2\lambda R_k + N_r T R)^{-1}(2\lambda z_k + N_r T z), \]  

where \((\cdot)^{-1}\) denotes the Moore-Penrose pseudoinverse of \((\cdot)\). The energy of \( r_{si}(t) \) can be simply expressed as

\[ E_{r_{si}} = \int_{-T/2}^{T/2} |r_{si}(t)|^2 dt = (2\lambda + N_r T)\eta_{si}. \]  

Thus, the maximum CR representing the optimal cancellation performance of SSIC can be expressed as

\[ G^S_{\text{MAX}} = 10 \log \frac{(2\lambda + N_r T)\eta_{si}}{(N_R + N_c)BT + \sum_{k=1}^{K} E_{r_{res}}^k|_{w_k = w^{opt}_k}}, \]  

Let \( \gamma_T \) be the signal-to-noise ratio (SNR) for the transmitted signal, \( \gamma_c \) be the SNR for the reconstructed signal, and \( \gamma_n \) be the interference-to-noise ratio (INR) for the SI signal, which are respectively defined as

\[ \gamma_T = 10 \log \frac{2\lambda}{N_r T}, \]  

\[ \gamma_c = 10 \log \frac{(2\lambda + N_r T)\eta_{si}}{N_c T B}, \]  

\[ \gamma_n = 10 \log \frac{(2\lambda + N_r T)\eta_{si}}{N_r T B}. \]  

Thus, \( G^S_{\text{MAX}} \) can be rewritten as a function related to \( \gamma_T, \gamma_c \) and \( \gamma_n \).

Obviously, the theoretical upper bound of CR be reached only when the SI signal is completely canceled, i.e., \( E_{r_{res}} = 0 \). The theoretical upper bound of CR is expressed as

\[ G_{UB} = 10 \log \frac{(2\lambda + N_r T)\eta_{si}}{(N_R + N_c)BT}. \]  

Since the RF canceller is composed of passive components, only thermal noise exists in \( n_c(t) \) and thus we have \( N_c \ll N_R \). Therefore, equation (29) can be approximately reduced to \( G_{UB} \approx \gamma_R \).

### C. PERFORMANCE COMPARISON

In particular, SSIC reduces to the MTSIC method in the case of \( K = 1 \). From (23), the optimal weight vector \( w^{opt} \) solved by MTSIC is

\[ w^{opt} = R^H z. \]  

From (17), the energy of residual signal energy in MTSIC is expressed as

\[ E_{r_{res}} = (2\lambda + N_r T)(\eta_{si} + w^H R w - w^H z - z^H w). \]  

Therefore, the maximum CR for MTSIC can be obtained by substituting \( w^{opt} \) in \( E_{r_{res}} \), that is,

\[ G^M_{\text{MAX}} \approx 10 \log \frac{\eta_{si}}{(\eta_{si} - z^H R^H z)}. \]  

From (17) and (31), \( E_{r_{res}} = \sum_{k=1}^{K} E_{r_{res}}^k \), \( E_{r_{res}}^{1}, E_{r_{res}}^{2}, \ldots, \) and \( E_{r_{res}}^{K} \) are convex functions [20] that have minimum values at \( w^{opt}_1, w^{opt}_2, \ldots, \) and \( w^{opt}_K \), respectively, such that

\[ E_{r_{res}}^k|_{w_k = w^{opt}_k} \leq E_{r_{res}}^k|_{w_k = w^{opt}_k}, k = 1, 2, \ldots, K. \]  

Thus,

\[ \sum_{k=1}^{K} E_{r_{res}}^k|_{w_k = w^{opt}_k} \leq \sum_{k=1}^{K} E_{r_{res}}^k|_{w_k = w^{opt}_k} = E_{r_{res}}|_{w = w^{opt}}. \]  

From (34), we have

\[ G^S_{\text{MAX}} \geq G^M_{\text{MAX}}. \]  

Therefore, the proposed SSIC method outperforms the conventional MTSIC method.
D. SETUP TIME FOR PERFECT CANCELLATION

For the proposed SSIC method, SIC is performed when the cancellation parameters have not been completely obtained, which causes mismatch between the cancellation parameters and the segmented SI signal so that the cancellation performance is not optimal.

To assess the time consumed before the cancellation performance reaches the optimal level, we define a setup time for perfect cancellation $t_{\text{setup}}$, which is equivalent to the time consumed for computing all the cancellation parameters, the amplitudes $\alpha_1^k, \alpha_2^k, \ldots, \alpha_L^k$ and the phases $\theta_1^k, \theta_2^k, \ldots, \theta_L^k$, $k = 1, 2, \ldots, K$.

We assume that the time required for the parameter module shown in Fig. 3(b) to calculate a set of amplitude and phase parameters is $t_p$. The time required to calculate each set of amplitude and phase parameters is shown in Fig. 4, where $T$ is the sweep repetition interval and $t_k = kT/K$, $k = 1, 2, \ldots, K$.

**Case 1:** $t_p \leq T/K$

![FIGURE 4. The time required to calculate each set of amplitude and phase parameters (the black downward arrow).](image)

**Case 2:** $t_p > T/K$

Thus, the time required to calculate $K$ sets of amplitude and phase parameters, that is, the setup time for perfect cancellation of SSIC, is

$$t_{\text{setup}} = \begin{cases} T + t_p, t_p \leq T/K \\ T/K + Kt_p, t_p > T/K \end{cases} \quad (36)$$

Equation (36) shows that the setup time for perfect cancellation $t_{\text{setup}}$ keeps $T + t_p$ constant when $t_p \leq T/K$, and this parameter begins to increase linearly with $K$ when $K$ increases to $t_p > T/K$, the trend of which is shown in Fig. 5 in detail.

As a comparison, the setup time of MTSIC is also presented in Fig. 5. Because the cancellation parameters need to be calculated only once, the setup time of MTSIC is always $T + t_p$. Thus, the setup time for perfect cancellation of SSIC is longer than that of MTSIC when the number of segments $K$ is large.

IV. SIMULATION AND EXPERIMENTAL VERIFICATION

A. SIMULATION RESULTS

This subsection compares the cancellation performance of the SSIC and MTSIC methods by simulations with the bandwidth $B$ and the number of taps $L$ as independent variables, respectively. Moreover, for the proposed SSIC method, the effects of the number of segments $K$ on the cancellation performance are discussed. The specific parameters of the LFM signal in the simulations are presented in Table 1.

**TABLE 1.** Parameters of the LFM signal for simulation.

| Parameter                  | Value  |
|----------------------------|--------|
| Center frequency $f_c$ (GHz) | 3.1    |
| Sweep repetition interval $T$ (us) | 20     |
| SNR for the transmitted signal $\gamma_r$ (dB) | 48     |
| SNR for the reconstructed signal $\gamma_c$ (dB) | 60     |
| INR for the SI signal $\gamma_R$ (dB) | 47     |

**TABLE 2.** Parameters of the multipath SI channel for simulation.

| Path No. | Delay (ns) | Amplitude | Phase (rad) |
|----------|------------|-----------|-------------|
| 1        | $1$        | $0.3604$  | $0.9036$    |
| 2        | $2.9134$   | $0.0867$  | $0.4963$    |
| 3        | $4.9086$   | $0.0787$  | $-2.3575$   |
| 4        | $6.0023$   | $0.0723$  | $1.8958$    |
| 5        | $9.2130$   | $0.0695$  | $-1.4812$   |
| 6        | $11.8644$  | $0.0475$  | $-2.5488$   |
| 7        | $14.1303$  | $0.0395$  | $2.9653$    |
| 8        | $15.9695$  | $0.0317$  | $2.3870$    |

The CM1 statistical model based on the modified S-V channel model recommended in the IEEE 802.15.3a standard, which can effectively model the ultrawideband indoor channel with a communication distance of 0~4 m [21], is adopted in the simulations. More specifically, according to the statistical parameters of the CM1 model, an SI channel with 8 paths is randomly generated in the simulations. The parameters of this channel are shown in Table 2, where the phase is taken as a uniformly distributed random variable in
the range of \([-\pi, \pi]\). During the simulations, the delays of the taps are uniformly configured from 1 ns to 16 ns according to the number of taps.

### FIGURE 6.
The cancellation performance \(G\) versus the bandwidth \(B\).

The cancellation performance with respect to the bandwidth \(B\) is illustrated in Fig. 6, where the bandwidth ranges from 100 MHz to 500 MHz. For both methods, the cancellation performance degrades with increasing bandwidth. For MTSIC, the cancellation performance rapidly decreases to 15 dB at 300 MHz, which means that this method is not effective for an SI signal with a larger bandwidth. As a comparison, the performance of SSIC for wideband SI signals significantly improves with increasing number of segments \(K\). For example, when the bandwidth is 500 MHz, SSIC with \(K = 4\) can still cancel an SI with approximately 37 dB. Since the SSIC method converts wideband SIC into several narrowband SICs by segmenting the LFM signal in the time domain, this method is more robust in wideband scenarios than the conventional MTSIC method.

### FIGURE 7.
The cancellation performance \(G\) versus the number of taps \(L\) in the case of \(B = 250\) MHz.

Fig. 7 shows the relationship between the number of taps and the cancellation performance. The performance improves with increasing number of taps for both methods, and SSIC outperforms MTSIC for a given number of taps. In other words, to achieve the same performance, the fewer taps are required in SSIC than in MTSIC, which means that SSIC has lower hardware complexity than MTSIC. Therefore, it can be concluded that SSIC can achieve better cancellation performance with lower hardware complexity than MTSIC.

Both Fig. 6 and Fig. 7 show that the cancellation performance improves with increasing number of segments \(K\), which is shown in Fig. 8 in detail. Due to the time-frequency segmentation equivalence of the LFM signal, the more segments there are, the smaller the bandwidth of the time-domain segmented signals, thereby readily resulting in perfect cancellation. Moreover, the simulation results imply that the required number of segments increases, achieving the desired performance, with a decrease in the number of taps or an increase in the bandwidth.

### B. EXPERIMENTAL VERIFICATION

To further verify the performance of the proposed SSIC method and the corresponding simulation results, we test the experimentally measured data with MATLAB software. An FMCW phased-array radar operating at a center frequency of 3.1 GHz with a sweep bandwidth of 250 MHz is built and tested, and a transmit array element and a receive array element are randomly selected to work simultaneously. A dual-channel oscilloscope is used to simultaneously collect the transmitted and received signals, whose spectra are inspected with a spectrum analyzer. Photographs of the experimental setup and block diagrams of the experiment are shown in Fig. 9(a) and (b), respectively.

The specific experimental parameters are shown in Table 3, where \(\gamma_c\) is a hypothetical value used in MATLAB to avoid the influence of the noise in the reconstructed signal on the

| Number of Segments \(K\) | Theory \(G\) | Simulation \(G\) | Upper Bound \(G\) |
|--------------------------|---------|----------------|----------------|
| \(K=1\)                  |         |                |                |
| \(K=2\)                  |         |                |                |
| \(K=3\)                  |         |                |                |
| \(K=4\)                  |         |                |                |

- Theory: Theoretical value
- Simulation: Simulated result
- Upper Bound: Upper limit value
The time-domain waveforms of the reference signal (the transmitted signal) and the SI signal (the received signal) are shown in Fig. 10. This figure shows that the SI signal suffers from serious fading and distortion due to the complicated wireless channel environment, causing the measured SI signal to contain abundant multipath components. These aforementioned disadvantages make cancellation increasingly difficult.

The power spectral density of the SI signal before and after cancellation is shown in Fig. 11, from which we can see that the cancellation performance of the conventional MTSIC method improves with increasing number of taps. However, the SI signal can be canceled only by approximately 26 dB with MTSIC even when the number of taps is as high as 64, and the level of the residual SI signal after cancellation is still much higher than the receiver noise floor, which means the receiver is easily saturated and blocked.

As a comparison, the SI signal is canceled by approximately 45 dB across a bandwidth of 250 MHz by using the SSIC method with 4 taps, and the residual SI signal after cancellation is only 2 dB above the receiver noise floor. Meanwhile, the out-of-band leakage caused by nonlinearity is improved to some extent. Note that the above performance is obtained when the number of segments $K$ is as high as 1800, which significantly exceeds the number of segments mentioned.

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### TABLE 3. Parameters of the experimentally measured data.

| Parameter                        | Value |
|----------------------------------|-------|
| Sample rate (GHz)               | 40    |
| Number of samples               | $8 \times 10^6$ |
| Transmit power (dBm)            | -10   |
| SI signal power (dBm)           | -42   |
| SNR for the transmitted signal $\gamma_T$ (dB) | 48    |
| SNR for the reconstructed signal $\gamma_c$ (dB) | 60    |
| INR for the received signal $\gamma_R$ (dB) | 47    |
TABLE 4. Experimental results with the SSIC and MTSIC methods, where \(L\) is the number of taps and \(K\) is the number of segments.

| CR (dB) | MTSIC | SSIC |
|--------|-------|------|
| \(L=1\) | \(K=1\) | 1.32 | 14.70 |
|        | \(K=10\) | 2.98 | 24.35 |
|        | \(K=100\) | 16.40 | 44.74 |
|        | \(K=1800\) | 27.34 | 46.91 |
| \(L=4\) | \(K=1\) | 5.40 | 16.90 |
|        | \(K=10\) | 16.40 | 46.80 |
|        | \(K=100\) | 22.52 | 46.91 |
|        | \(K=1800\) | 35.36 | 46.99 |
| \(L=16\) | \(K=1\) | 15.96 | 26.07 |
|        | \(K=10\) | 22.52 | 46.95 |
|        | \(K=100\) | 35.36 | 46.99 |
|        | \(K=1800\) | 44.74 | 46.99 |

CR: The cancellation ratio defined in Section II.

used in the simulation with the same performance due to the complex channel environments and the effect of the out-of-band nonlinearity. Additionally, too many segments result in a longer setup time for perfect cancellation. The specific experimental results are shown in Table 4.

As shown in Table 4, similar to the simulation results, the experimental results demonstrate that the SSIC method can achieve better cancellation performance with fewer taps than the conventional MTSIC method.

V. CONCLUSION

To address the fact that a wideband SI signal is difficult to cancel in FMCW phased-array radar, an RF SSIC method is proposed in this paper. The proposed SSIC method converts the wideband SIC into several narrowband SICs, and thus, this method is more robust in wideband scenarios than the conventional MTSIC method. Moreover, simulation and experimental results show that the SSIC method can achieve higher cancellation performance with lower hardware complexity than the MTSIC method.

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