An Improved Method for Reflectivity Estimation in Frequency-Sharing Weather Radar System

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***Abstract*—In frequency-sharing weather radar system where multiple radars reside in the same frequency channel, the reflectivity estimation error is dominated by the noise and the interference signals. Prior to the reflectivity estimation process, the CLEAN algorithm iteratively estimates the reflection channel, and therefore the stopping criterion of the CLEAN algorithm determines the reflectivity estimation performance. We propose the noise and interference power estimation scheme from the received signal in the presence of target, interference, and noise signals. The estimated noise pulse interference power is taken to the threshold in the stopping criterion of the CLEAN algorithm. Simulation results show that the proposed scheme outperforms the conventional scheme in terms of the reflectivity estimation performance.**

***Index Terms*—frequency-sharing weather radar, CLEAN algorithm, reflectivity estimation**

I. Introduction

The Center for Collaborative Adaptive Sensing of the Atmosphere (CASA), was established in 2003 to develop high spatial density networks of weather Doppler radars for sensing the lower atmosphere [1]. Such weather radar system (WRS) is composed of multiple weather radars, which are located at remote sites and measure the weather parameters (e.g., reflectivity and radial velocity) nearby. The weather parameters from multiple radars are combined to mitigate the problems arising from radar geometry (cone of silence, beam spreading, beam height, beam blockage, etc.).

Due to the significant data traffic increase of mobile communications, spectrum resource becomes more precious. However, in the current WRSs, radars operate in distinct frequency bands and thus the required bandwidth for a WRS increases linearly with the number of radar deployments. Recently, for the enhancement of spectrum efficiency, the frequency-sharing WRS using nearly orthogonal pulse compression codes was proposed in [2]. More specifically, by applying the matched filter, the inter-site interference between radars was suppressed owing to the nearly orthogonal pulse compression codes. Then, the modified CLEAN algorithm [3] was applied to the matched filter output for removing the sidelobe interference, which occurs even when there is only one radar operating in the frequency band. The modified CLEAN algorithm iteratively removes the target signals (including both mainlobe and sidelobe components) from the matched filter output, and concurrently reconstructs the impulse response, until the residual signal power of the matched filter output is less than a threshold. With an over-large threshold, some real target signals may be misrecognized as noise. On the other hand, with an over-small threshold, the suppressed inter-site interference and noise signals may be misidentified as target signals. However, the WRS taken the threshold to be the noise level in [2] without considering the interference power, and thus the noise and interference signals may be misidentified as target signals.

In this paper, the threshold of the CLEAN algorithm is taken to be the noise plus interference power. Since the target, interference, and noise signals are concurrently received, their powers are difficult to be separately estimated. To solve the problem, we propose the scheme to estimate the noise and interference powers from the received signal in the presence of target, interference, and noise signals.

II. System Model

We consider a WRS, which is defined as a group of weather radars that reside in the same frequency channel. The weather scatterers (hydrometeors) in each range bin are assumed to be represented by a virtual point target, and thus the reflection model is greatly simplified while maintaining estimation accuracies on weather parameters [4]. Also, a target velocity does not change over the observation time, and this assumption is suitable for the case of, for example, (stable) rain fall or snow fall.

The WRS consists of N mono-static weather radars with pulse compression, where the n th radar is denoted by Radar-n. Each radar independently extract weather parameters in the presence of the interference from other radars. Since the N radars follow the same operations to

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extract weather parameters, we focus on a specific radar (we call this radar the main radar) and regard the other radars as interferers (we call these radars interfering radars). Without loss of generality, we consider Radar-1 as the main radar and Radars- $2,3,\cdots,N$ as interfering radars throughout this paper.

Each radar periodically switches between the transmission and reception phases. Specifically, in the transmission phase, the radar transmits its pulse in the current direction of the antenna; it captures radar signals in the reception phase. Then, it switches back to the transmission phase for transmitting the next pulse. Note that, in order to improve observation performance for a specific direction, a radar stays in the direction during a number of repetitions of the two phases: we denote this number of repetitions by K.

A. Pulse Compression Code Set

In conventional single radar systems, pulse compression was used for obtaining the high-resolution range profile under the limitation of peak power [5]. Recently, in the WRS, pulse compression was used for removing interferences from the interfering radars by employing mutually uncorrelated pulse compression codes in [2]. In this paper, the nearly orthogonal polyphase code set in [2] is used for the same purpose.

In the transmission phase, Radar- n transmits a pulse modulated with its own code, where each sub-pulse represents an element of the code. Let us denote the pulse compression code of Radar- n by

$$\mathbf{s}_n = [s_n[1], s_n[2], \cdots, s_n[L_c]] \tag{1}$$

where L_c is the code length. The design of the code set $[\mathbf{s}_1^T, \mathbf{s}_2^T, \cdots, \mathbf{s}_N^T]^T$ is not in the scope of this paper and is assumed to follow [2].

B. Received Signal Representation

Let $1 \le k \le K$, K, and l denote the pulse index, the number of emissions of pulses for observation of one direction, and the range bin index, respectively. Then, the received signal $r_k[l]$, $1 \le l \le L+(L_c-1)$, of Radar-1 is given by

$$r_k[l] = \sum_{n=1}^{N} (s_n * h_{k,n})[l+1] + w_k[l] \tag{2}$$

where $h_{k,n}[l]$, $1 \le l \le L$, is the impulse response of the reflection channel for the k th transmitted pulse of Radar- n, $w_k[l]$ is a white Gaussian noise, and * is the convolution operation defined as $(f*g)[m] = \sum_{j=-\infty}^{\infty} f[j]g[m-j]$.

The unknowns required to be estimated in this paper are as follows. The noise energy is given by

$$E_0 = \sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |w_k[l]|^2 \tag{3}$$

The target signal and interference signal energies are given by

$$\begin{aligned} E_n &= \sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |(s_n * h_{k,n})[l+1]|^2 \\ &= |\mathbf{s}_n|^2 \left(\sum_{k=1}^{K}\sum_{l=1}^{L} |h_{k,n}[l]|^2\right) + 2\mathrm{Re}\left\{\sum_{m=1}^{L_c-1} (s_n \otimes s_n)[m] A_n[m]\right\} \\ &\overset{(a)}{=} |\mathbf{s}_n|^2 \left(\sum_{k=1}^{K}\sum_{l=1}^{L} |h_{k,n}[l]|^2\right) \end{aligned} \tag{4}$$

where $\mathrm{Re}\{\cdot\}$ denotes the real part of its complex argument, $\otimes$ is the cross-correlation operation defined as $(f \otimes g)[m] = \sum_{j=-\infty}^{\infty} f^*[j]g[j+m]$, and $A_n[m] = \sum_{k=1}^{K}\sum_{l=1}^{L-m} h_{k,n}[l]h_{k,n}^*[l+m]$. The equality (a) follows from that $h_{k,n}[l]$ is non-stationary with respect to the range bin index l and the average of the complex term in $A_n[m]$ tends to zero, i.e., $A_n[m]=0$ [6].

III. Noise and Interference Power Estimation

Since target, interference, and noise signals are received at the same time, it is difficult to separately estimate their energies, i.e., $\{E_n\}_{n=0}^{N+1}$. If $N+1$ simultaneous equations are given, the $N+1$ unknowns are easily solved. First of all, the energy of $r_k[l]$ is given by

$$\begin{aligned} b_0 &= \sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |r_k[l]|^2 \overset{(b)}{=} \sum_{n=1}^{N}\sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |(s_n * h_{k,n})[l+1]|^2 \\ &\quad + \sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |w_k[l]|^2 \\ &= \sum_{n=1}^{N} |\mathbf{s}_n|^2 \left(\sum_{k=1}^{K}\sum_{l=1}^{L} |h_{k,n}[l]|^2\right) \\ &\quad + 2\mathrm{Re}\left\{\sum_{n=1}^{N}\sum_{m=1}^{L_c-1} (s_n \otimes s_n)[m] A_n[m]\right\} + \sum_{k=1}^{K} \sum_{l=1}^{L+(L_c-1)} |w_k[l]|^2 \\ &\overset{(c)}{=} \sum_{n=1}^{N} E_n + E_0. \end{aligned} \tag{5}$$

The equality (b) follows from the fact that $\{h_{k,n}\}_{n=1}^{N}$ and $w_k[l]$ are independent each other. The equality (c) follows from $A_n[m]=0$. Let us consider N matched filters, where the i th filter is fed by $r_k[l]$ and is matched to $\frac{\mathbf{s}_i}{|\mathbf{s}_i|}$. Then, by following the similar procedure as (5), the energy of the matched filter output is given by

$$b_i = \sum_{k=1}^{K}\sum_{l=1}^{L+2(L_c-1)} |y_k^{(i)}[l]|^2 = \sum_{k=1}^{K}\sum_{l=1}^{L+2(L_c-1)} |(\frac{s_i}{|\mathbf{s}_i|}\otimes r_k)[l-L_c]|^2$$
$$= \sum_{n=1}^{N} |\mathbf{t}_{i,n}|^2 (\sum_{k=1}^{K}\sum_{l=1}^{L} |h_{k,n}[l]|^2)$$
$$+2\mathrm{Re}\{\sum_{n=1}^{N}\sum_{m=1}^{2L_c-2}(t_{i,n}\otimes t_{i,n})[m]A_n[m]\}$$
$$+\sum_{k=1}^{K}\sum_{l=1}^{L+2(L_c-1)} \frac{|(s_i\otimes w_k)[l-L_c]|^2}{|\mathbf{s}_i|^2}$$
$$= \sum_{n=1}^{N}\frac{|\mathbf{t}_{i,n}|^2}{|\mathbf{s}_n|^2}E_n + E_0 \quad (6)$$

where $y_k^{(i)}[l]$, $1 \le l \le L+2(L_c-1)$, is the output of the filter matched to $\frac{\mathbf{s}_i}{|\mathbf{s}_i|}$, $t_{i,n}[l] = (\frac{s_i}{|\mathbf{s}_i|}\otimes s_n)[l-L_c]$, $\mathbf{t}_{i,n} = [t_{i,n}[1], t_{i,n}[2], \cdots, t_{i,n}[2L_c-1]]$.

From (5) and (6), N + 1 simultaneous equations are given by

$$\begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & \frac{|\mathbf{t}_{1,1}|^2}{|\mathbf{s}_1|^2} & \cdots & \frac{|\mathbf{t}_{1,N}|^2}{|\mathbf{s}_N|^2} \\ \vdots & \vdots & \vdots & \vdots \\ 1 & \frac{|\mathbf{t}_{N,1}|^2}{|\mathbf{s}_1|^2} & \cdots & \frac{|\mathbf{t}_{N,N}|^2}{|\mathbf{s}_N|^2} \end{bmatrix} \begin{bmatrix} E_0 \\ E_1 \\ \vdots \\ E_N \end{bmatrix} = \begin{bmatrix} b_0 \\ b_1 \\ \vdots \\ b_N \end{bmatrix} \quad (7)$$

and the solutions, denoted as $\{E_n^*\}_{n=0}^{N+1}$, can be easily obtained by the standard simultaneous equations methods. The estimated energies $\{E_n^*\}_{n=0}^{N+1}$ can be converted to power units, i.e., $P_n^* = \frac{E_n^*}{K(L+L_c-1)}$. Then, the threshold for the CLEAN algorithm is determined as

$$T = \frac{L+L_c-1}{L+2(L_c-1)}(\sum_{n=2}^{N}\frac{|\mathbf{t}_{1,n}|^2}{|\mathbf{s}_n|^2}P_n^* + P_0^*) \quad (8)$$

IV. Reflectivity Estimation

In order to suppress the interference and noise, the received signal $r_k[l]$ is fed into the filter matched to $\frac{\mathbf{s}_1}{|\mathbf{s}_1|}$, and the output $y_k^{(1)}[l]$ is given by

$$y_k^{(1)}[l] = (\frac{s_1}{|\mathbf{s}_1|}\otimes r_k)[l-L_c]$$
$$= \sum_{n=1}^{N}(t_{1,n} * h_{k,n})[l+1] + \frac{(s_1\otimes w_k)[l-L_c]}{|\mathbf{s}_1|^2} \quad (9)$$

To improve the performance further, $y_k^{(1)}[l]$ is fed into the modified CLEAN algorithm that was proposed in [3] to effectively eliminates sidelobe interference for binary coding radar signals with contiguous scattering targets. The modified CLEAN algorithm iteratively removes the target signals (including both mainlobe and sidelobe components) from $y_k^{(1)}[l]$, and concurrently estimates the impulse response from the targets, until the residual signal power of $y_k^{(1)}[l]$ is less than the threshold T. For the single radar system, T was taken to be the noise level in [3]. If an over-large T is selected, some real target signals are misrecognized as noise. On the other hand, if an over-small T is selected, noises may be misidentified as target signals. In the WRS we consider, noise plus interference power is the proper value for T.

Let $z_k[l]$ denote the output of the CLEAN algorithm, and then the reflectivity at l th range bin can be estimated as [2], [7]

$$R[l] = \frac{(\frac{1}{2}cl\Delta)^2}{k_R G_t G_r}\frac{1}{K}\sum_{k=1}^{K}|z_k[l]|^2 \quad (10)$$

where c is the speed of light, Δ is the sub-pulse duration time, G_t is the antenna gains for transmission, G_r is the antenna gains for reception, and the k_R depends on the radar parameters such as 3-dB beamwidth, pulse duration time, and wavelength.

V. Simulation Results

In this section, we examine the performance of the proposed scheme for frequency sharing WRS through computer simulations. We assume a simple radar network consisting of $N = 4$ weather radars, where Radar-1 operates as the main radar while Radar-2, Radar-3, and Radar-4 operate as interfering ones. Next, we construct four reflection channels, i.e., $\{h_{k,n}[l]\}_{n=1}^{4}$. Specifically, we design the power of the impulse responses and the velocities of the weather scatterers in the corresponding reflection channel as in Fig. 1 and Fig. 2, respectively. The every member function in Fig. 1 has been normalized with its maximum magnitude. By following [2], [7], [8], $\{h_{k,n}[l]\}_{n=1}^{4}$ are generated from the channel power and the velocity. We set $K = 50$, $L = 300$, and $L_c = 128$.

The performance requirement of the WSR-88D on reflectivity is that, when the signal-to-noise ratio (SNR) is higher than 10 dB, the reflectivity error should be less than 1 dB. We set the SNR of the WRS to 10 dB in simulations for considering the worst case. The Signal-to-Interference Ratio (SIR) in the WRS is evaluated as

$$SIR = \frac{\sum_{k=1}^{K}\sum_{l=1}^{L+(L_c-1)}|(s_1 * h_{k,1})[l+1]|^2}{\sum_{n=2}^{N}\sum_{k=1}^{K}\sum_{l=1}^{L+(L_c-1)}|(s_n * h_{k,n})[l+1]|^2} \quad (11)$$

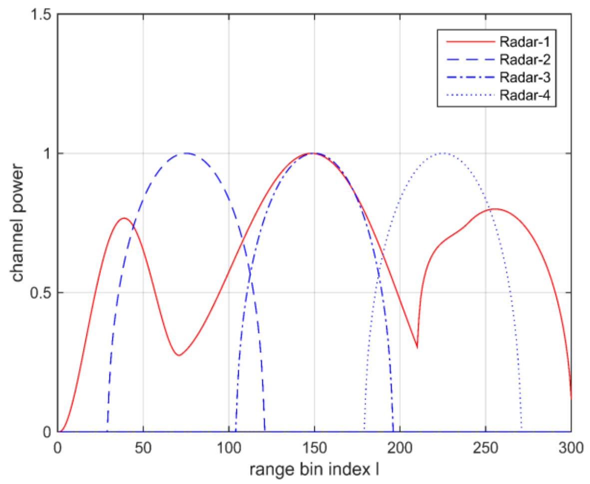

Figure 1. Normalized amplitudes of reflection channels $\{h_{k,n}[l]\}_{n=1}^{4}$.

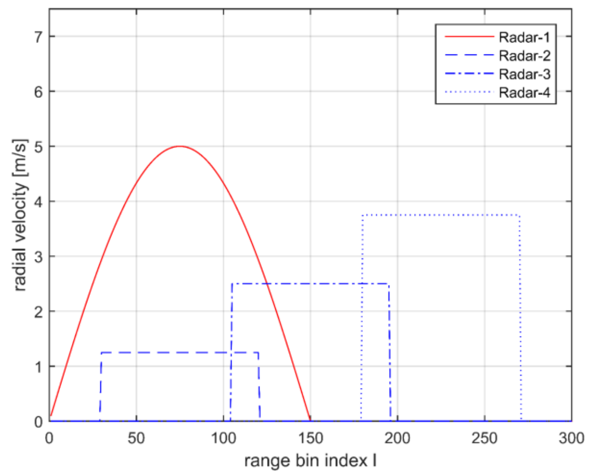

Figure 2. Velocities of the weather scatterers in the corresponding reflection channels $\{h_{k,n}[l]\}_{n=1}^{4}$.

Fig. 3 and Fig. 4 show the estimated channel power and the reflectivity error with respect to the range bin index, respectively. During $75 \leq l \leq 90$, while the conventional scheme in [2] can not satisfies the requirement, the proposed scheme satisfies the requirement.

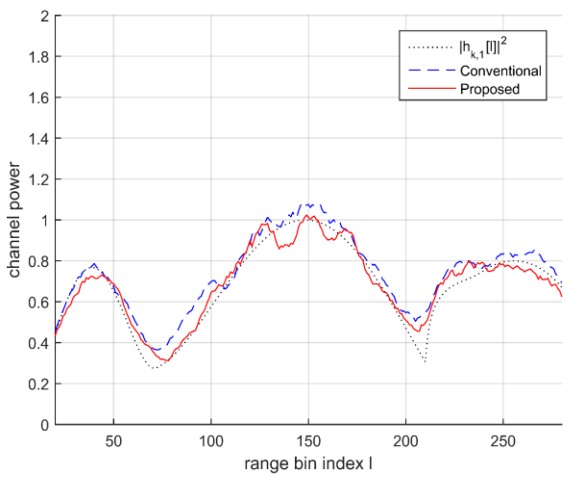

Figure 3. Comparison of the estimated channel powers as a function of range bin index l.

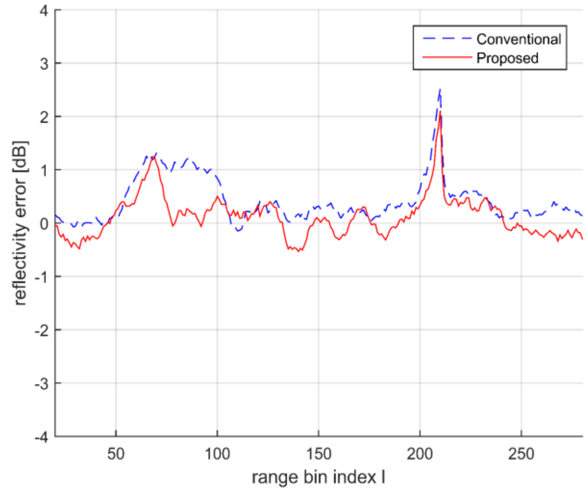

Figure 4. Comparison of the reflectivity errors as a function of range bin index l.

Fig. 5 shows the reflectivity error with respect to SIR. The reflectivity error showed in Fig. 5 is the error averaged value over the range. Although both of the conventional and proposed scheme satisfies the requirement in terms of the averaged reflectivity error, the proposed scheme outperforms the conventional scheme at the low SIR region.

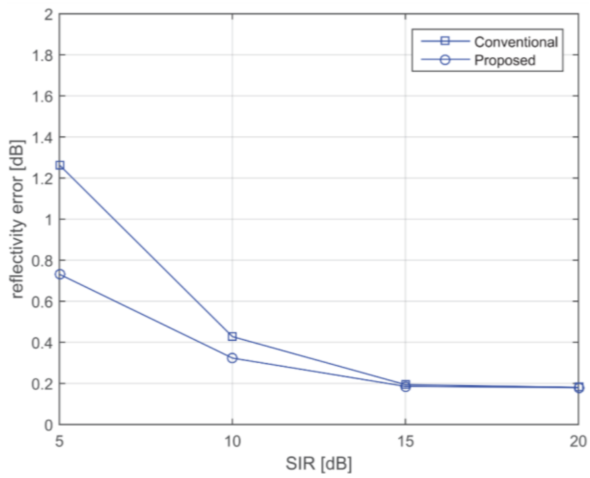

Figure 5. Comparison of the reflectivity errors as a function of SIR.

VI. Conclusion

In this paper, the noise and interference power estimation scheme was proposed using the known orthogonal pulse compression codes. The estimated noise and interference powers are applied to the stopping criterion of the CLEAN algorithm. Simulation results show that the proposed scheme outperforms the conventional scheme in terms of the estimated reflectivity.

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