

# Orthogonal zero correlation zone codes design for MIMO radar system

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This paper investigates the MIMO radar orthogonal waveform design problem. The result shows the ambiguity function (AF) property of the sum of transmitted signals is as important as the autocorrelation and crosscorrelation properties of transmitted signals. An orthogonal zero correlation zone (ZCZ) phase codes design method was proposed to simultaneously satisfy these property requirements. The idea is to simultaneously constrain the peak sidelobe of the AF of sum signal in ZCZ, and the peak sidelobe of aperiodic autocorrelation function (ACF) and peak of the aperiodic crosscorrelation function (CF) within ZCZ. We used the sequential quadratic programming (SQP) method to solve the nonlinear optimization problem with multi-variables and multi-constraints. Some of the designed results are presented, and their properties are improved significantly. The designed orthogonal ZCZ codes promise to be practically applicable.

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## 1. Introduction

Recently, multiple-input multiple-output (MIMO) radar has received more and more attention [1-6]. The waveform design problem is fundamental important for MIMO radar system. So far the optimal orthogonal phase codes for MIMO radar are the ones whose both peak sidelobe of the aperiodic autocorrelation function (ACF) and peak of the aperiodic crosscorrelation function (CF) are the minimum possible for a given code length, and have a minimum Doppler loss for a moving target. Therefore, in considerable literature on the design of MIMO radar orthogonal phase codes set, attention is only paid to the autocorrelation and crosscorrelation properties [16] and Doppler tolerance of the designed codes. As noted by prior authors, low autocorrelation and Doppler loss are critical because they facilitate high range resolution, and low cross-correlation levels are important for reducing mutual interference as well as maximizing independent information.

However, we argue that an additional metric is needed to achieve good performance for MIMO radar system. The additional metric is the property of ambiguity function (AF) of the sum of transmitted signals. It is known that MIMO radar employs multiple transmitting antennas to simultaneously transmit orthogonal probing signals and also uses multiple receiving antennas to receive the reflected signals from the targets. The orthogonal signals arriving from different transmitting antenna can be

separated by match-filtered processing and then be performed the equivalent transmit beamforming on the receive side “after the fact”. At the analysis stage, we presented the AF of the sum signal is equivalent to the equivalent transmit beamforming result of the signal component. Thus, low peak sidelobe level of the sum signal’s AF has become of increasing importance in the MIMO radar system design. Numerical examples of Section IV show the signals with low autocorrelation sidelobe level and low crosscorrelation level still might have poor AF sidelobe level of sum signal. Therefore, the property of AF of sum signal needs to be accounted for the orthogonal waveforms design of MIMO radar.

On the other hand, orthogonal codes set with low correlation level are needed only over a narrow window around the origin in some applications such as in the situation where the target distributions are sparse. A sequences set with the property that the autocorrelation sidelobe and crosscorrelation all vanish in a specified zone is referred to a zero correlation zone (ZCZ) sequences set. The ZCZ of this paper refers to very-low-correlation zone, that is, the correlation of ZCZ codes within the specified zone is very low, but not zero. This paper introduced an approach to orthogonal ZCZ phase codes design which can simultaneously satisfy all of these requirements. The idea is to simultaneously constrain the peak sidelobe of the AF of the sum of transmitted signals within the ZCZ, and the peak sidelobe of aperiodic autocorrelation function and peak of the aperiodic crosscorrelation function over the

ZCZ. To solve the nonlinear multi-variables, multi-constraints optimization problem, the sequential quadratic programming (SQP) method is used. Note that the designed result is continuous by SQP. We performed an additional design step which quantifies the continuous phase codes to get the discrete polyphase codes set which promises to be practically applicable. Some of the designed results are presented, and their properties are improved significantly.

The rest of this paper is organized as follows. Section II introduces the signal model and formulates the problem. In Section III the SQP method for orthogonal ZCZ codes set design is described. In Section IV the design results are presented. In Section V some conclusions are drawn.

## 2. Problem formulation

Consider a narrowband MIMO radar system with  $M$  transmitters and  $N$  receivers. We assume the two antenna arrays to be linear and parallel. The transmitter and the receiver are close enough that they share the same angle variable. Each array is composed of omnidirectional elements. The pulse of duration  $T_p$  is divided into  $L$  subpulses of identical duration. The phase coded waveform transmitted by the  $m$ th antenna can be expressed as

$$s_m(t) = \sum_{l=1}^L e^{j\phi_m(l)} \text{rect}\left(\frac{t - (l-1)T_p/L}{T_p/L}\right) \quad (1)$$

for  $m=1, 2, \dots, M$ , where  $t$  represents the time within the pulse (fast time) and

$$\text{rect}(x) = \begin{cases} 1 & 0 \leq x < 1 \\ 0 & \text{others} \end{cases} \quad (2)$$

and  $\phi_m(l)$  is the phase of the  $l$ th subpulse of the  $m$ th transmitted signal, which only can be selected from the phase codes set of  $P$  phase  $\{0, \frac{2\pi}{P}, 2 \cdot \frac{2\pi}{P}, \dots, (P-1) \cdot \frac{2\pi}{P}\}$ .

We assume a point target is located in the far field. Owing to the receiving array aperture is unrelated to the problem addressed here (because the phase differences in the reflected signals caused by the differences of the receiving antenna locations can be compensated by conventional receive beamforming), hence we only consider the scenario where only one receiving antenna is used to receive the reflected signal (namely  $N=1$  for simplicity). Thus, the demodulated received signal can be expressed as

$$\mathbf{x}(t, f_d) = \alpha_t \sum_{m=1}^M e^{j2\pi(m-1)\frac{d \sin \theta_t}{\lambda}} s_m(t - \tau) e^{j2\pi f_d(t - \tau)} + \mathbf{n}(t) \quad (3)$$

where  $\lambda$  is wavelength,  $d$  is the spacing of the transmitting antennas,  $f_d$  is Doppler shift associated with target,  $\theta_t$  is target direction,  $\tau$  is the propagation delay,  $\alpha_t$  is the corresponding path loss including the two-way propagation loss and the reflection coefficient, which is in general a complex number. The first term in (3) represents the signal reflected by the target. The second term is white noise. We assume that there is no antenna array misalignment. Throughout this paper we shall assume that  $d = \lambda/2$ . To simplify (3), we let  $\alpha_t = 1$ ,  $\tau = 0$ , and let

$$\mathbf{x}_s(t) = \sum_{m=1}^M e^{j2\pi(m-1)\frac{d \sin \theta_t}{\lambda}} s_m(t) \quad (4)$$

Then (3) can be rewrite as

$$\mathbf{x}(t, f_d) = \mathbf{x}_s(t) e^{j2\pi f_d t} + \mathbf{n}(t) \quad (5)$$

After being received, the echo signals can be performed to pass through a bank of matched filters, which is the conjugated time-reversed version of the transmit signals. Then the output of matched filter  $\mathbf{q}_m(t)$  ( $\mathbf{q}_m(t) = \mathbf{s}_m^*(-t)$ , where  $(\cdot)^*$  indicates Hermitian conjugate) can be written as

$$\mathbf{y}_m(t, f_d) = \int_{-\infty}^{\infty} \mathbf{x}_s(u) e^{j2\pi f_d(u)} \mathbf{s}_m^*(u - t) du + \tilde{\mathbf{n}}_m(t) \quad (6)$$

for  $m=1, 2, \dots, M$ , where  $\tilde{\mathbf{n}}_m(t) = \int_{-\infty}^{\infty} \mathbf{n}(u) \mathbf{s}_m^*(u - t) du$ .

After matched filtering, the  $m$ th row of the matrix  $\mathbf{y}(t, f_d) = [\mathbf{y}_1^T(t, f_d), \mathbf{y}_2^T(t, f_d), \dots, \mathbf{y}_M^T(t, f_d)]^T$  represents

the  $m$ th channel of data, where  $T$  is the transpose operator.

It is thus possible to linearly combine these channels, namely equivalent transmit beamforming. The weight vector of equivalent transmitting beamformer can be written in the form

$$\mathbf{a}_{\theta_0} = [1, e^{-j2\pi d \sin \theta_0 / \lambda}, \dots, e^{-j2\pi(M-1)d \sin \theta_0 / \lambda}]^T \quad \text{and further}$$

assume the direction pointed by the steering vector is the target direction, namely  $\theta_0 = \theta_t$ , then,

$$\begin{aligned} \mathbf{z}(t, f_d) &= \mathbf{a}_{\theta_0}^T \mathbf{y}(t, f_d) \\ &= \sum_{m=1}^M e^{-j2\pi \frac{d \sin \theta_t}{\lambda} (m-1)} \int_{-\infty}^{\infty} \mathbf{x}_s(u) e^{j2\pi f_d(u)} \mathbf{s}_m^*(u - t) du + \tilde{\mathbf{n}}(t) \\ &= \int_{-\infty}^{\infty} \mathbf{x}_s(u) \mathbf{x}_s^*(u - t) e^{j2\pi f_d(u)} du + \tilde{\mathbf{n}}(t) \end{aligned} \quad (7)$$

where  $\tilde{\mathbf{n}}(t)$  is the white noise after both matched filter

processing and equivalent transmit beamforming. Note that the second term in (7) is the AF of  $\mathbf{x}_s(t)$ :

$$|\chi(t, f_d)|^2 = \int_{-\infty}^{\infty} \mathbf{x}_s(u) \mathbf{x}_s^*(u-t) e^{j2\pi f_d u} du \quad (8)$$

From (4) we can see that  $\mathbf{x}_s(t)$  is the weighted sum of transmitted signals. To simplify the analysis, we focus here on the scenario where  $\theta_i = 0^\circ$ . The problem under the condition of any target direction will be explored in the future. Thus  $\mathbf{x}_s(t)$  represents the sum signal of transmitted signals. Equation (7) and equation (8) imply that the AF of the sum signal is equivalent to the equivalent transmit beamforming result of signal component. Thus, the low peak sidelobe of the sum signal's AF property has become of increasing importance in the design of the MIMO radars. However, the codes set with good autocorrelation and crosscorrelation properties doesn't guarantee whose sum signal still have good AF property, which will be shown in the simulation result of Section IV. Therefore, we need to add the AF property of sum signal which is equally important with ACF and CF properties to evaluate the performance of designed waveforms.

### 3. Waveform design

In some applications, orthogonal codes set with low correlations are needed only over a narrow window around the origin such as in the situation where the target distributions are sparse. The ZCZ of this paper refers to very-low-correlation zone, that is, the correlation of ZCZ codes within the specified zone is very low, but not zero. To describe conveniently, we first define the ZCZ. We assume  $2K$  ( $K < L-1$ ) range bins around the origin are ZCZ, which implies the correlation level of  $[-K, -1] \cup [1, K]$  range bins should be very low; we also assume  $(-F_D, F_D)$  Doppler bins are ZCZ. This Doppler range is divided to  $2D-1$  Doppler bins by Doppler resolution. It is sufficient to study the first quadrant of the AF according to the symmetry property of AF. Thus only the  $K \times D$  range-Doppler ZCZ needs to optimize, namely corresponding range bins set is  $[1, K]$ , and corresponding Doppler bins set is  $[0, D-1]$ . These Doppler shifts corresponding to Doppler bins are  $f_d = 0.5d/T_p$  for  $d = 0, 1, 2, \dots, D-1$ .

To describe conveniently, we define the following aperiodic ACF and aperiodic CF of transmitted signals for MIMO radar.

$$A(\phi_m, k) = \begin{cases} \frac{1}{L} \sum_{l=1}^{L-k} \exp j[\phi_m(l) - \phi_m(l+k)], & 0 \leq k < L \\ \frac{1}{L} \sum_{l=-k+1}^L \exp j[\phi_m(l) - \phi_m(l+k)], & -L < k < 0 \end{cases}, \quad m = 1, 2, \dots, M \quad (9)$$

$$C(\phi_p, \phi_q, k) = \begin{cases} \frac{1}{L} \sum_{l=1}^{L-k} \exp j[\phi_q(l) - \phi_p(l+k)] & 0 \leq k < N \\ \frac{1}{L} \sum_{l=-k+1}^L \exp j[\phi_q(l) - \phi_p(l+k)] & -N < k < 0 \end{cases}, \quad \begin{matrix} p \neq q, \\ p, q = 1, 2, \dots, M \end{matrix} \quad (10)$$

The AF of sum signal in (8) can be written as the following discrete form:

$$|\chi(k, d)|^2 = \sum_{l=-L}^L \mathbf{x}_s(l) \mathbf{x}_s^*(l-k) e^{j2\pi f_d l} \quad \begin{matrix} k = 1, 2, \dots, K, \\ d = 0, 1, \dots, D-1 \end{matrix}, \quad f_d = 0.5d/T_p \quad (11)$$

Therefore, the goal of this section is to design the orthogonal ZCZ codes set which can simultaneously satisfy the requirements of ACF, CF and the AF of sum properties, by minimize the following three figures of merit:

- peak sidelobe of aperiodic autocorrelation in ZCZ, that is,  $A(\phi_m, k)$  for  $k = -K, \dots, K$  and  $k \neq 0; m = 1, 2, \dots, M$ ;
- aperiodic crosscorrelation of transmitted signals within ZCZ, that is,  $C(\phi_p, \phi_q, k)$  for  $k = -K, \dots, K; p, q = 1, 2, \dots, M$ ;
- the AF peak sidelobe of sum signal over ZCZ, that is,  $|\chi(k, f_d)|^2$  for

$$k = 1, 2, \dots, K; f_d = 0.5d/T_p; d = 0, 1, \dots, D-1.$$

Consequently, the objective function can be given as follows:

$$\min_{\Phi} E = \min_{\Phi} \max \left\{ \max_{\substack{k=-K, \dots, K, k \neq 0 \\ m=1, \dots, M}} |A(\phi_m, k)|, \mu_1 \max_{\substack{k=-K, \dots, K \\ p, q=1, \dots, M}} |C(\phi_p, \phi_q, k)|, \mu_2 \max_{\substack{k=1, 2, \dots, K \\ f_d=0.5d/T_p \\ d=0, 1, \dots, D-1}} |\chi(k, d)|^2 \right\} \quad (12)$$

where  $\Phi$  is the size- $LM$  phase vector, the value of each element of  $\Phi$  is 0 to  $2\pi$  (not including  $2\pi$ );

$\max_{\substack{k=-K, \dots, K, k \neq 0 \\ m=1, \dots, M}} |A(\phi_m, k)|$  represents the peak sidelobe of the

aperiodic ACF of transmitted signals within ZCZ,

$\max_{\substack{k=-K,\dots,K \\ p,q=1,\dots,M}} |C(\phi_p, \phi_q, k)|$  is the peak of the aperiodic CF of

transmitted signals within ZCZ,  $\max_{\substack{k=1,2,\dots,K, \\ f_d=0.5d/T_p \\ d=0,1,\dots,D-1}} |\chi(k, d)|^2$  is the

peak of the AF of sum signal over ZCZ of range-Doppler plane,  $\mu_1$  and  $\mu_2$  are the weighting parameters which characterizes the importance of the crosscorrelation property and the AF of sum signal in the optimization process, typically,  $\mu_1 = \mu_2 = 1$ . The optimization problem in (12) can be changed into the following nonlinear optimization problem with multi-variables and multi-constraints.

$$\begin{aligned} & \min_{\phi, z} z \\ \text{s.t. } & |A(\phi_m, k)| \leq z, \quad k = 1, 2, \dots, K, m = 1, 2, \dots, M \\ & \mu_1 |C(\phi_p, \phi_q, k)| \leq z, \quad k = -K, \dots, 0, 1, \dots, K, p \neq q = 1, 2, \dots, M \\ & \mu_2 |\chi(k, d)|^2 \leq z, \quad k = 1, 2, \dots, K, f_d = 0.5d/T_p, d = 0, 1, \dots, D-1 \end{aligned} \quad (13)$$

where  $z$  is both the instrumental variable used for constraint and the objective function. Equation (13) first constrain the aperiodic ACF and aperiodic CF within ZCZ and the AF of sum signal over ZCZ and let each of constraints be less than or equal to the instrumental variable  $z$ , and then minimize  $z$ . These constraints make amplitude of both the autocorrelation sidelobe and crosscorrelation in ZCZ and the sidelobe of the AF in ZCZ distribute even, and then minimize instrumental variable  $z$ , which means simultaneously minimize the even-distributed autocorrelation sidelobe and crosscorrelation within ZCZ and the sidelobe of the AF within ZCZ in range-Doppler plane. The design requirements can be satisfied by this optimization means.

The nonlinear optimization problem with multi-variables and multi-constraints in (13) can be solved by Sequential Quadratic Programming (SQP) method. As with most optimization methods, SQP has arguably become the most successful method for solving nonlinearly constrained optimization problems. SQP is an iterative method for nonlinear optimization, which solve a sequence of optimization subproblems, each of which optimizes a quadratic model of the objective subject to a linearization of the constraints. If the objective function and the constraints are twice continuously differentiable, SQP methods can be used to solve the problem. The SQP methods have been implemented in the package of

MATLAB.

We assume the initial phase vector and the optimization result vector are denoted as  $\Phi_0$  and  $\Phi_{opt}$ , respectively, both of which are continuous phase vector. It should be noted that the initialization problem of phase vector, i.e., each element of  $\Phi_0$  must have a  $[0, 2\pi)$  uniformly distribution.

To apply the technique into practice, the designed continuous phase codes still need to quantify to discrete phase codes. According to the number of phase available  $P$  decided by the transmitter, the discrete orthogonal ZCZ codes can be obtained through the following quantify operator:

$$\tilde{\Phi}_{opt} = \lfloor \Phi_{opt} P / 2\pi \rfloor \frac{2\pi}{P} \quad (14)$$

where  $\lfloor \cdot \rfloor$  represents a ‘‘floor’’ rounding operation.

#### 4. Numerical results

Based on the method described in Section III, the phase coded waveforms set has been designed. Table I lists the designed polyphase codes set with  $L=128$ ,  $M=3$  and  $P=128$ . The other parameters include: the pulse duration  $T_p = 25.6\mu s$  and wavelength  $\lambda = 0.25m$ . Select 40 range bins adjacent to the origin (namely  $K=20$ ) and the Doppler shifts set  $[-2/T_p, 2/T_p]$  are selected as the ZCZ. The Doppler shifts set correspond to the target radial velocity range  $[-9765m/s, 9765m/s]$ . The positive Doppler range of  $[0, 2/T_p]$  is divided into  $D=5$  Doppler bins by the velocity resolution  $0.5/T_p$ . Let  $\mu_1 = \mu_2 = 1$ . Resolve the optimization problem in (13) by using the SQP method to get continuous phase codes set. We assume the number of phase available provided by transmitter is 128 ( $P=128$ ). Then we used (14) to quantify them and finally obtained the orthogonal 128-phase codes which possess low correlation within specified correlation zone.

Table 1. Phase sequences of a designed ZCZ polyphase code set with  $L=128$ ,  $M=3$ , and  $P=128$ .

No.	Phase sequences
Sequence 1	127,61,73,127,117,124,79,31,68,127,0,0,64,43,0,98,85,13,63,43,127,89,61,0,88,81,117,127,28,85,127,17,108,109,6,1,55,93,26,48,46,3,30,89,70,9,95,11,47,99,37,86,79,103,77,20,25,53,121,43,15,74,78,122,118,111,50,79,0,14,59,15,98,48,21,58,51,0,14,0,0,68,59,40,118,35,8,76,112,40,11,17,21,118,41,0,127,100,59,123,101,94,66,38,29,84,35,68,70,58,67,39,52,70,127,48,10,28,114,55,105,17,40,76,41,70,58,62
Sequence 2	100,55,12,92,68,76,56,10,104,127,115,101,18,54,88,98,127,19,127,74,74,23,73,78,0,58,83,87,69,93,32,116,100,81,105,61,68,8,54,104,81,90,106,30,120,69,74,32,39,118,57,5,83,102,86,31,43,6,38,47,18,116,28,61,127,125,7,62,112,12,91,122,85,11,60,126,86,69,82,37,103,118,88,15,61,44,121,1,64,127,0,108,102,116,28,51,40,44,50,0,52,80,0,56,73,100,97,70,9,107,50,44,79,113,127,98,52,82,55,0,62,66,85,83,90,11,111,28
Sequence 3	67,3,57,90,33,71,98,2,0,14,98,0,75,63,84,82,65,34,88,109,68,21,47,0,48,41,15,74,53,86,28,99,1,127,100,122,0,127,97,81,64,30,114,20,67,81,125,83,116,1,69,44,0,69,57,125,58,27,114,98,60,0,93,102,0,0,127,63,13,4,21,126,44,39,74,6,124,0,36,83,72,54,72,121,46,21,34,57,4,104,74,28,63,0,108,21,50,78,86,48,0,60,0,127,14,127,82,127,79,19,12,119,122,34,59,12,99,33,93,38,0,34,81,10,26,92,85,94

The ACFs of the polyphase Sequence 1, Sequence 2, and Sequence 3 of the ZCZ codes set are shown in the first row of Fig. 1, respectively. The CFs between Sequence 1 and Sequence 2, Sequence 1 and Sequence 3, and Sequence 2 and Sequence 3 are shown in the second row of Fig. 1, respectively. And the Fig. 2 is a larger version of Fig.1 in ZCZ.

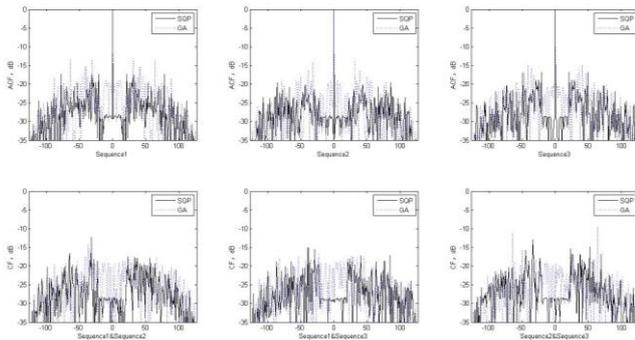


Fig. 1. The ACF and CF of polyphase sequences set designed.

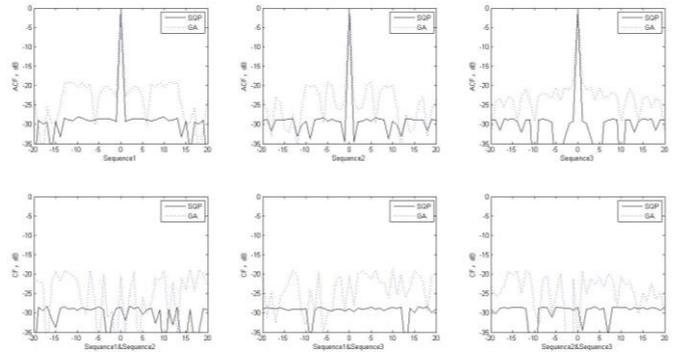


Fig. 2. A larger version of Fig.1 in ZCZ.

In Fig. 1 and Fig. 2, the solid black line depicts the autocorrelation and crosscorrelation of the sequences designed by the proposed method based on SQP method; the dashed blue line, in contrast, depicts the autocorrelation and crosscorrelation of designed results in Ref. [8] which used genetic algorithm (GA).

The autocorrelation peak sidelobe level (APSL) and crosscorrelation peak (CP) achieved by the proposed method in the ZCZ are  $-28.1\text{dB}$  and  $-28.2\text{dB}$ , respectively. While under the same simulation condition (the same  $M$  and  $L$ ), GA used in Ref. [8] yielded ZCZ codes set with both of APSL and CP being  $0.1154$  ( $-18.75\text{dB}$ ). By comparing them, we can find that the APSL and CP of the designed ZCZ polyphase codes set in this paper outperform the results of Ref. [8] approximately  $9.4\text{dB}$ .

It's important to note that the four phase coded set obtained by GA in Ref. [8] will not change much as the number of phase available  $P$  increases [12].

The simulation results of the signals after matched filter processing and equivalent transmit beamforming are

provided as follows. The AF of the sum of transmitted waveforms designed by the proposed method is shown in Fig. 3. The AF of sum signal designed by Ref. [8], in contrast, is shown in Fig. 4.

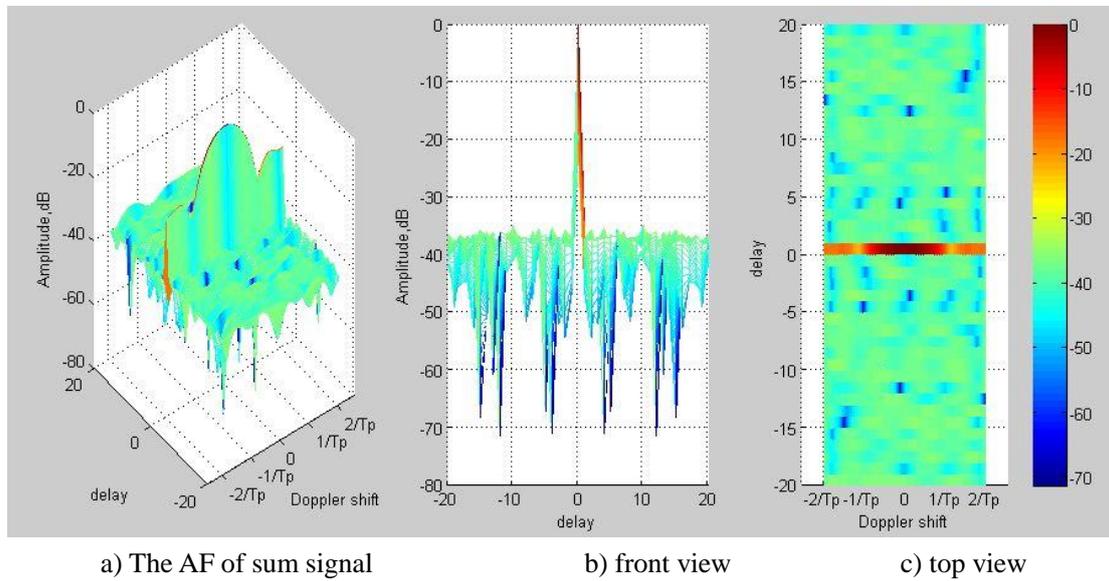


Fig. 3. The AF of the sum of transmit waveforms set designed by the proposed algorithm in this paper.

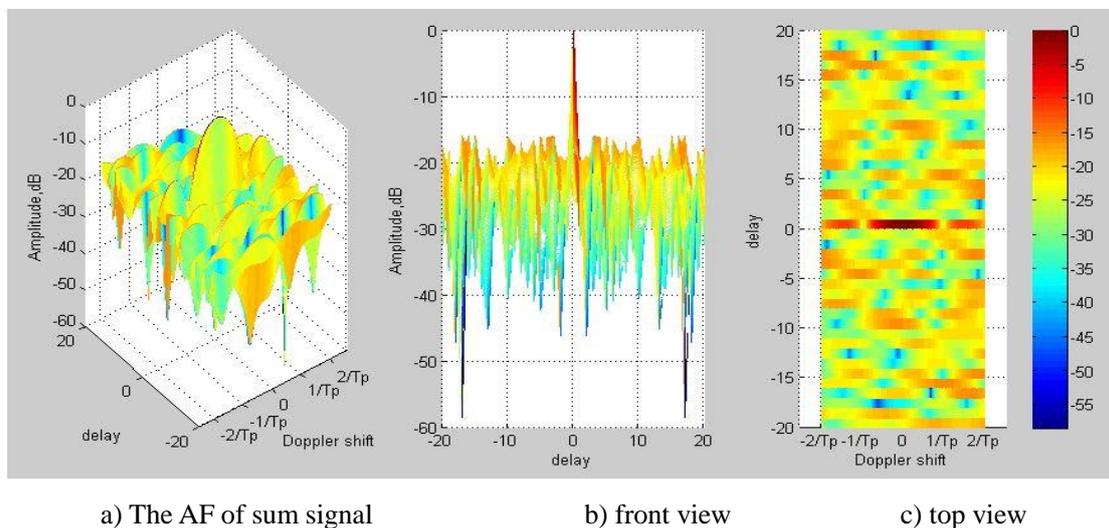


Fig. 4. The AF of the sum of transmit waveform sets designed by Ref. [8].

In Fig. 3, the peak sidelobe of the AF of sum signal in ZCZ is -34.84dB; while the peak sidelobe of the AF of sum signal in same zone is -15.83dB. Comparing the Fig. 3 and Fig. 4, it can be seen that the peak sidelobe of the AF achieved by the proposed algorithm outperforms the algorithm in Ref. [8] 19dB. From Fig. 4, it also can be seen that only optimizing the autocorrelation and crosscorrelation properties cannot guarantee the good

result after matched filter processing and equivalent transmit beamforming. Actually, the peak sidelobe of the AF of sum signal gets worse 2.92dB than its APSL and CP. Therefore, it is necessary to simultaneously optimize the properties of autocorrelation, crosscorrelation and the AF of sum signal.

## 5. Conclusions

To date, the theoretical literature on MIMO radar has focused largely on the use of orthogonal waveforms. These orthogonal waveforms always are separated by matched filter processing on the receiver and linearly combined by equivalent transmit beamforming “after the fact”. Through the analysis in Section II, we find that the AF of the sum of the transmitted waveforms is equivalent to the equivalent transmit beamforming result of signal component. Therefore, the peak sidelobe level of the AF of sum signal is equally important as ACF and CF properties. On the other hand, orthogonal waveforms set with low correlations are needed only over a narrow zone in some application. Thus, this paper proposed an orthogonal ZCZ phase codes design method to simultaneously satisfy these performance requirements. The idea is to simultaneously constrain the peak sidelobe of the AF of sum signal in ZCZ, and the peak sidelobe of aperiodic ACF and peak of the aperiodic CF within ZCZ. To satisfy all of these requirements, an efficient sequential quadratic programming (SQP) method is used to design the desired waveforms set. Note that the obtained phase codes are continuous. Therefore we performed an additional design step which quantifies the continuous phase codes to get the discrete polyphase codes set which promises to be practically applicable. Future research might concern the effect of target direction to orthogonal ZCZ codes set design.

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