MIMO RADAR DETECTION IN COMPOUND
GAUSSIAN CLUTTER USING ORTHOGONAL
DISCRETE FREQUENCY CODING SPACE TIME
WAVEFORM

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ABSTRACT
This paper proposes orthogonal Discrete Frequency Coding Space Time Waveforms (DFCSTW) for
Multiple Input and Multiple Output (MIMO) radar detection in compound Gaussian clutter. The proposed
orthogonal waveforms are designed considering the position and angle of the transmitting antenna when
viewed from origin. These orthogonally optimized show good resolution in spikier clutter with Generalized
Likelihood Ratio Test (GLRT) detector. The simulation results show that this waveform provides better
detection performance in spikier Clutter.

KEYWORDS
Multiple Input and Multiple Output (MIMO), orthogonal Discrete Frequency Coded Space Time
Waveforms (DFCSTW), Generalized Likelihood Ratio Test (GLRT), Compound Gaussian Clutter

1. INTRODUCTION

Multiple Input and Multiple Output system (MIMO) transmits multiple linearly independent
probing signals via its transmit antennas and receives multiple coded waveforms from multiple
locations. MIMO radar systems have many advantages including increased angle resolution [1-6],
increased Doppler resolution [1, 7], reduced ground-based radar clutter levels [1], sharper
airborne radar clutter notches [2,7], Lower Probability of Intercept (LPI) [1,8], and relaxed
hardware requirements [1].

The performance of the transmitted waveforms is judged by their correlation properties [9-12].
The waveforms should have good autocorrelation properties for high range resolution and good
cross-correlation for multiple target return separability. So, there is a need to design MIMO radar
waveforms as orthogonal pulses with low correlation properties. In literature various algorithms
have been proposed to design the orthogonal sequences with low autocorrelation and cross-
correlation peak sidelobe levels. In [9], the focus is to design orthogonal Discrete Frequency
Coding Waveforms_Frequency Hopping (DFCW_FF) for netted radar systems using simulated
annealing (SA) algorithm to optimize the frequency sequences, [10] and [11] focus on orthogonal
DFCW_FF and orthogonal Discrete Frequency Coding Waveforms_Linear Frequency
Modulation (DFCW_LFM) to design multiple orthogonal sequences with good correlation using
a Modified Genetic Algorithm (MGA) technique. In [12] various Cyclic Algorithms (CA) for
unimodular MIMO radar waveforms are designed for good correlation properties. Target Radar
Cross Section (RCS) fades the received signal from the target. One way to maximize the system’s processing gain is by antenna spacing [13-15]. In [16], adaptive space time waveform was proposed to improve detection performance. To improve the system performance, antenna spacing is one of the important factors.

Statistical characterization of the clutter is necessary for designing the detector. The clutter echoes result from very large number of elementary scatterers due to which the radar system has relatively low resolution capability. As the radar resolution increases, the statistics of the clutter has no longer been observed to be Gaussian. There is experimental evidence that high resolution radar systems are now plagued by target-like “spikes” that gives rise to non-Gaussian heavy tailed observations [17-18]. The high resolution sea clutter is modelled by compound Gaussian clutter which is a sample of compound K -distribution clutter. In [19-23] the focus is on the Generalized Likelihood Ratio Test (GLRT) detector to yield excellent performance and it is very much attractive for radar detection in the presence of correlated non-Gaussian clutter model.

In this paper, the proposed Orthogonal Discrete Frequency Coded Space Time Waveforms (DFCSTW) can improve the target detection in compound Gaussian (spikier) clutter. Section II illustrates the signal model and GLRT detector, section III shows the numerical simulations for the model developed in section II. Section IV concludes the paper.

2. WAVEFORM MODEL

Consider a MIMO radar system with a $T_x$ transmitting antennas. Let $x_i$ where $i \in \{1,2,\ldots,T_x\}$ denote the position of $T_x$ transmitting antennas located at an angle $\theta_i$ when viewed from an origin. Each element may transmit $N$ coding frequencies on each subpulse of a waveform. Each $R_x$ receives and processes the signal from all the $T_x$ transmitters. The received signals are the reflected signals from a target and clutter. Each element transmits $N$ pulses with a Pulse Repetition frequency (PRF) of $f_r$.

2.1. Discrete Frequency Coding Space Time Waveform (DFCSTW)

Linear Frequency Modulation (LFM) is the first and probably the most popular pulse compression method. Discrete Frequency-coding Waveform (DFCW) has the large compression ratio. DFCW can lower correlation properties if sequences are coded properly. The basic idea is to sweep the frequency band (B) linearly during the pulse duration (T) and the time bandwidth product of the signal is BT. The spectral efficiency of the DFCW improves as the time-bandwidth product increases, because the spectral density approaches a rectangular shape. Here we consider the sequence length of each waveform (N) and Number of antennas ($T_x$).

The Discrete Frequency Coding Space Time (DFCSTW) Waveform is defined as

$$S_p(t, \Phi) = \sum_{n=0}^{N-1} e^{j2\pi f_n^p (t-nT_p)} e^{j2\pi x_p \sin \Phi_i \frac{f_r}{c}} \begin{cases} e^{-j2\pi kn^3}, & 0 \leq t \leq T_p \\ 0, & \text{elsewhere} \end{cases}, p = 1, 2\ldots T_x$$ (1)

where $s$ is the frequency slope, $s=B/T$ and $p=1, 2, \ldots, T_x$. $T$ is the subpulse time duration. $N$ is the number of subpulse that is continuous with the coefficient sequence $\{n_1, n_2, \ldots, n_N\}$ with unique permutation of sequence $\{0,1,2,\ldots,N\}$. $f_n^p = n \Delta f$ is the coding frequency of subpulse $n$ of waveform $p$ in the waveform. $\Delta f$ is the frequency step. Where $x_p, i=1,2,\ldots,T_x$ denote the position of $T_x$ transmitting antennas located at an angle $\theta_i$ when viewed from an origin. The choice of BT, $T, \Delta f$ and $B/\Delta f$ values are crucial for the waveform design. Different lengths of firing sequence (N) have different values for each of the above mentioned parameters [11].
3. SIGNAL MODEL

The received signals for MIMO radar can be formulated as

\[ r_i = S_i * T_i + S_i * C_l + V_i, \quad i = 1, 2, ..., R_x \]  (2)

Where \( S \) is the transmitted code matrix. \( T_i = [T_{i1}, \ldots, T_{iTx}]^T \), \( i = 1, 2, \ldots, R_x \) are the complex values accounting for both the target backscattering. \( V = [V_{i1}, \ldots, V_{iTx}]^T, \quad i = 1, 2, \ldots, R_x \) are noise component. \( r_i = [r_{i1}, \ldots, r_{iN}]^T, \quad i = 1, 2, \ldots, R_x \) are the echo signals of the \( i \)th receiver antenna contaminated by the clutter. The clutter vectors \( n_i \) are assumed as compound Gaussian random vector i.e., [22]

\[ C_l_i = \sqrt{\alpha_i \beta_i}, \quad i = 1, \ldots, R_x \]  (3)

The texture \( \alpha_i \) is non-negative random variable which models the variation in power that arises from the spatial variation in the backscattering of the clutter and the speckle components \( \beta_i \) are correlated complex circular Gaussian vectors and independent to one other. This \( \alpha_i \) is independent Zero-mean complex circular Gaussian vector with covariance matrix.

\[ R_i = E[n_i n_i^H] = \alpha_i r_o \]  (4)

Whereas \( r_o = [C_l C_l^H] \) is the covariance structure and \( H \) is complex conjugate. The Compound Gaussian clutter is samples from K-distribution with pdf.

\[ f(z) = \frac{2v}{\tau(v)} \left( \frac{2v}{\mu} z \right)^{vz} k_v - 1 \left( \frac{2v}{\mu} z \right) \]  (5)

The texture component \( \sqrt{\alpha_i} \) is gamma distribution with pdf

\[ f(\tau_i) = \frac{1}{\sqrt{\nu}} \left( \frac{\nu}{\mu} \right)^{\nu - 1} e^{-\nu/\mu}\sqrt{\tau_i} u(\tau_i) \]  (6)

where \( \Gamma(\cdot) \) is the Eulerian Gamma function, \( \nu > 0 \) is the parameter ruling the shape of the distribution, \( u(\cdot) \) denotes the unit-step function, and \( K_v (\cdot) \) is the modified second kind Bessel function with order \( \nu \), which rules the clutter spikiness. The smaller the value of \( \nu \), the higher is the tails of the distribution. The distribution will become Gaussian for \( \nu \rightarrow \infty \).

The clutter has exponential correction structure of covariance matrix \( r_o \), the (i,j) element of which is \( \rho_i^{\mu} \), where \( \rho \) is one-lag correlation coefficient. The Power Spectral Density of clutter is generally located in low frequency region & Clutter spread is controlled by \( \nu \). The small the values of \( \nu \) the spikier is the clutter.

3.1. GLRT Detector

Suppose \( k \) (\( k \geq N \)) secondary data vector, sharing the same covariance structure of the primary data is available, \( r_i \) and \( r_{ik}, \quad i = 1, 2, \ldots, R_x, \quad k = 1, 2, \ldots, k \) are the received signal from the primary and
secondary data. Then, the detecting of a target with MIMO radar can be formulated in terms of the following binary hypotheses test.

\[
H_0: \begin{cases} 
    r_i = S_i^*C_i + V_i, i = 1,2,\ldots,R_x \\
    r_{ik} = S_{ik}^*C_{ik} + V_{ik}, i = 1,2,\ldots,R_x, k = 1,2,\ldots,k 
\end{cases}
\]

\[
H_1: \begin{cases} 
    r_i = S_i^*T_i + S_i^*C_i + V_i, i = 1,2,\ldots,R_x \\
    r_{ik} = S_{ik}^*T_{ik} + S_{ik}^*C_{ik} + V_{ik}, i = 1,2,\ldots,R_x, k = 1,2,\ldots,k 
\end{cases}
\]  

(7)

The GLRT detector [22] based on the primary data can be obtained by replacing the unknown parameters with their maximum likelihood estimates in the likelihood ratio. The GLRT detector [22] of the complex amplitude \( TH \)

\[
\prod_{i=1}^{r} \frac{r_i^H r_o^{-1} r_i}{r_i^H (r_o^{-1} - r_o^{-1} S(S^H r_o^{-1} S)^{-1} S^H r_o^{-1} r_o) r_o} < TH \\
H_o
\]  

(8)

where the TH is variable detection threshold. In a practical adaptive radar system, the covariance matrix of the clutter is estimated from a set of secondary data, which must be representative of the samples in the Cell Under Test (CUT). To make the detectors ensure the CFAR property w.r.t texture statistics, a normalized sample covariance matrix is adopted, based on the secondary data collected by the receiver antennas.

\[
\hat{R}_{oi} = \frac{N}{K} \sum_{k=1}^{K} \frac{n_{ik} n_{ik}^H}{n_{ik} n_{ik}^H} 
\]  

(9)

Substituting eq. (9) in eq. (8), we get the adaptive detector.

\[
\prod_{i=1}^{r} \frac{r_i^H r_o^{-1} r_i}{r_i^H (r_o^{-1} - r_o^{-1} S(S^H r_o^{-1} S)^{-1} S^H r_o^{-1} r_o) r_o} < TH \\
H_o
\]  

(10)

For a given value of \( N \), as \( k \) varies the proposed adaptive detectors end up coincident with real scenario. However, for finite values of \( K \), the performance of the estimate and eventually of the adaptive detector itself depends upon the actual values of \( N \). Thus it is necessary to quantify the loss of the proposed decision strategy with respect to its non adaptive counterpart under situations of exact covariance matrix.

In order to compare the performance of the GLRT detector with Gaussian clutter GLRT detector (GC-GLRT),

\[
\sum_{i=1}^{r} r_i^H r_o^{-1} S(S^H r_o^{-1} S)^{-1} S^H r_o^{-1} r_i < TH 
\]  

(11)

In order to limit the computational burden, we assume \( P_{fa} \) as \( 10^{-4} \) and also to save the simulation time. The transmit code matrix \( S \) is the orthogonal DFCSTW and the Signal-to-Clutter Ratio (SCR) is defined as

\[
SCR = \frac{\alpha^2}{NT_x} \text{tr}[S^H r_o^{-1} S] 
\]  

(12)
4. DESIGN RESULTS

Consider an orthogonal DFCSTW code set for 4×4 MIMO radar with code length of N=8. The simulation is carried out in MATLAB. The frequency code sets are optimized using ACC_PSO [24]. These sequences are considered in eq. (1) to generate the orthogonal DFCSTW set with good correlation properties. The above generated Code sequence matrix is used in the signal model mentioned in the section II to generate the DFCSTW where θ_i is the angle of the Tx transmitting antennas are generated randomly and placed linearly. Thus generated waveform is implemented in the signal model mentioned in the section III.

The P_ds of GLRT and of GC-GLRT are plotted versus SCR with P_{fa}=10^{-4}, N=8, N_T =4, N_R =4, ρ=0.9, K=64, v=0.5 in Fig.1. The performance of GLRT is better than GC-GLRT. Fig 2 shows the pds versus SCR for orthogonal DFCSTW code set and Space Time Code (STC) waveforms. It can be observed that the performance of orthogonal DFCSTW code is better than STC. The value of Pds is 0.6 at -25 dB in [21] and the simulated result is 0.68 at the same dB value for DFCSTW code. This shows that the orthogonal DFCSTW code sets perform better in the spikier clutter than STC.

Fig 3 shows the pds versus SCR for orthogonal DFCSTW code set for different values of v. As the values of v decreases the clutter is spikier and the detector works better of the value of v=0.3 than v=0.8. Thus the DFCSTW code set has better detection performance in spikier environment. In Table 1, lists the values of Pds versus SCR for -10 dB. It can be observed that the performance of this paper is much better than the existing statistics [20-22]. The curves show that the performance of GLRT with orthogonal DFCSTW code set is better in spikier clutter.

In Fig 4 the pds versus SCR for orthogonal DFCSTW code set for different values of transmitter. As the values of transmitter number increases the detector works better of the value of Tx=2 to Tx=4. Thus the DFCSTW code set has better detection performance in spikier environment. In Fig 5 the pds versus SCR for orthogonal DFCSTW code set for different values of receiver. As the values of receiver number varies, it can observed that the detector works better for Rx=4.

Figure 1. Pd versus SCR plots of GLRT (solid curves) and GCGLRT (dashed curves) receivers with Orthogonal DFCSTW Code set for Pfa=10^{-4}, N = 8, Tx =4, Rx =4, ρ = 0.9, K = 64, v=0.5 parameter.
Table 1. Values of Pd Vs SCR

<table>
<thead>
<tr>
<th>Literature</th>
<th>SCR (in dB)</th>
<th>Pd</th>
</tr>
</thead>
<tbody>
<tr>
<td>[20]</td>
<td>-10</td>
<td>0.02</td>
</tr>
<tr>
<td>[21]</td>
<td>-10</td>
<td>0.95</td>
</tr>
<tr>
<td>[22]</td>
<td>-10</td>
<td>0.8</td>
</tr>
<tr>
<td>For the simulated result</td>
<td>-10</td>
<td>0.998</td>
</tr>
</tbody>
</table>

Figure 2. Pd versus SCR plots of GLRT (solid curves) and GCGLRT (dashed curves) receivers with Orthogonal DFCSTW Code set and STC for Pfa=10^{-4}, N = 8, Tx =4, Rx =4, ρ = 0.9, K = 64, v=0.5 parameter.

Figure 3. Pd versus SCR plots of GLRT (solid curves) and GCGLRT (dashed curves) receivers with Orthogonal DFCSTW Code set for Pfa=10^{-4}, N = 8, Tx =4, Rx =4, ρ = 0.9, K = 64, v=0.3, 0.5 and 0.8 as variable parameter.
5. CONCLUSIONS

In this paper, orthogonal DFCSTW is modelled to transmit in a spikier clutter environment. The results show that the performance of $P_d$ is better for DFCSTW codes than STC waveform. The results also show that the DFCSTW code has the better performance in spikier clutter.

REFERENCES


AUTHORS

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