Experimental Demonstration of Cognitive Radar for Target Localization under Strong Interference

Xia Li*, Zhen Hu*, Robert C. Qiu*, Michael C. Wicks†

*Cognitive Radio Institute, Department of Electrical and Computer Engineering, Center for Manufacturing Research, Tennessee Technological University, Cookeville, Tennessee 38501, USA
Email: xli43@students.tntech.edu, {zhu, rqiu}@tn-tech.edu
†Sensor Systems Division, University of Dayton Research Institute, Dayton, OH 45469, USA
Email: Michael.Wicks@udri.udayton.edu

Abstract—One of the major objectives of cognitive radar is to form a dynamic closed feedback loop to adapt the spectrum of transmit waveforms to avoid certain interference. In this paper, we build an automatic closed-loop cognitive radar to support experimental study of the radar system in real-world situations. Convex optimization is applied to jointly design sounding waveforms and the matched filters with spectral power suppressed in arbitrary bands and with low correlation sidelobes as well. Target localization is demonstrated under strong interferences. Experimental results are provided to evaluate the performance of the cognitive radar system.

Index Terms—convex optimization, waveform diversity, closed-loop, MIMO distributed radar

I. INTRODUCTION

The road from the current adaptive radar and the radar with a function of waveform design to cognitive radar is the key evolution for the radar system. The adaptive radar focuses more on the adaptation at the receiver. The radar waveform design deals with the probing signal according to some optimization criterion. The dominant feature of cognitive radar is the cognition, which means the radar can actively learn about the environment and the whole radar system forms a dynamic closed feedback loop including the transmitter, environment and receiver [1].

The similar idea of cognitive radar is presented in [2]. The knowledge-aided fully adaptive approach is explored in cognitive radar which can possess unprecedented capabilities for adaptation in the challenging real-world environments. Waveform diversity supports cognitive radar in the radar waveform level. Waveforms can be optimized in real time to achieve the performance gain. In clutter dominant environment, maximizing the target energy and minimizing the clutter energy should be considered simultaneously. Besides, signal processing in the waveform level at the receiver should also be included into the waveform diversity framework. For example, adaptive filter or notch filter can be designed to cancel the clutter effect and interference [3]–[5].

The core parts which differentiate cognitive radar from previous radar systems are cognitive engine and knowledge base. Cognitive engine is the brain of cognitive radar. Intensive computation is involved in cognitive engine which can perform decision making, learning, understanding, conceiving, prediction, and so on. Knowledge base is a special database to store the radar related information. In the knowledge base, radio map for the surveillance area is an important information. Radio map can give the spectrum information in all spatial, temporal, and frequency domains. Besides, the knowledge obtained from previous illuminations should be stored and used for the following radar sounding. Sequential waveform design for target recognition has been discussed in [6].

In this paper, convex optimization will be exploited to support waveform diversity for target localization under strong interference. Optimization stems from human instinct, which tries to minimize or maximize objective function under various constraints. Convex optimization [7] is a sub-field of optimization. The foundation of convex optimization is that if a local minimum exists, the local minimum is a global minimum. As a result, the globally optimal solution will be obtained as long as the engineering problem can be formulated as a convex optimization.

The first version of cognitive radar we are building is an automatic closed-loop ultra-wideband (UWB) multiple input multiple output (MIMO) cognitive radar. The signal bandwidth of the radar system is 500 MHz, which meets the definition of UWB by FCC [8]. Target localization under strong interferences will be demonstrated in cognitive radar. Cognitive radar can automatically sense the spectrum, find the interferences, jointly optimize the sounding waveforms and the matched filters to get the time of arrival (ToA) information, finally locate the target.

The contribution of this paper has three folds:

- The automatic closed-loop cognitive radar is built to support experimental study of the radar system in the real-world situations;
- Convex optimization is applied to joint design of radar transceivers;
- Target localization is demonstrated under strong interferences.

The rest of the paper is organized as follows. Section II presents the system design and implementation of cognitive radar. In section III, waveform diversity using convex opti-
II. SYSTEM DESIGN AND IMPLEMENTATION

A. System Architecture

To demonstrate the concept of the proposed cognitive sensing system, a testbed is implemented [9]. As shown in Fig. 1, the computing engine, a computer running MATLAB, controls the transmitter and the receiver coherently in a loop to constitute a $2 \times 2$ MIMO cognitive radar. The transmitter is a field-programmable gate array (FPGA) based design that can transmit arbitrary sounding waveforms in real time. The receiver is based on a digital phosphor oscilloscope (DPO) to get RF digital samples. Digital signal processing for the receiver is implemented in MATLAB. ToA is obtained by searching the correlation peaks of the correlation between received baseband signal and locally stored finite impulse response (FIR) filter. Based on the ToA, the target position can be estimated by solving a group of non-linear equations. The cognitive engine jointly optimizes the sounding signal waveforms in the transmitter and the FIR filters in the receiver to maximize the auto-correlation peak and minimize the cross-correlation outputs while suppressing interference signal.

As an example, we assume the interference is a narrowband QPSK OFDM signal. The sampling rate of the OFDM interference is the same as the radar sounding signal. Interference tones are placed at several adjacent frequency slots to become a multitone interference (MTI) signal [10]. The interference can occupy arbitrary part of the radar signal frequency.

B. Hardware Implementation

The transmitter is based on an FPGA connected with a computer running MATLAB through USB port. Waveforms generated by MATLAB is transferred to FPGA through this port. The FPGA is connected with two digital to analog converters (DAC) operating at 1 Gsps. Each DAC provides two analog outputs as the inphase and quadrature components of the complex baseband signal.

The block diagram of the circuits inside FPGA is shown in Fig. 2. The baseband radar sounding waveform $p_m \in \mathbb{C}^{L \times 1}$ for the $m$th transmitter antenna is generated by MATLAB. $L$ is the waveform sample points. An interface is designed in MATLAB to send the waveforms in PC memory to USB port. An USB controller block in FPGA is used to receive the waveform data from the USB port. Each pulse can be as long as 160 sample points, making the pulse length of each pulse reach up to 160 ns. A trigger signal is generated to inform the receiver the start of pulse transmission.

Shown in Fig. 3, analog RF components are connected to the output of DAC. The modulator up convert the baseband signal to 4 GHz carrier frequency. The power amplifier can provide up to 30 dBm maximum output power.

In receiver, as shown in Fig. 4, the DPO are connected to 2 receiver antennas. The sampling rate of each channel is
MATLAB Fixed point

R peaks are used to estimate ToA. g received signals and the locally stored FIR filter from 50 GHz to 500 MHz. By conducting correlation between convert the received signal from passband to baseband. The sampling rate of the pulse waveforms is 500 MHz and changed in Fig. 5. The convex optimization engine generate the pulse shaping fixed point conversion USB interface FPGA.

C. Software Implementation

In the transmitter part, the flow of the processing is shown in Fig. 5. The convex optimization engine generate the pulse waveforms based on the spectrum sensing results. The sampling rate of the pulse waveforms is 500 MHz and changed to 1 GHz by an up-sampling module. USB interface driver is used to transfer fixed point waveform data from MATLAB to FPGA.

In the receiver part, the flow of the processing operations is shown in Fig. 6. Digital down converter is used to down convert the received signal from passband to baseband. The down sampling module is used to change the sampling rate from 50 GHz to 500 MHz. By conducting correlation between received signals and the locally stored FIR filter \( g_m \in \mathbb{C}^{L \times 1} \), the correlation peaks are obtained. Positions of the correlation peaks are used to estimate ToA.

The positions of the 2 transmitter antennas are expressed as \( T_m = (x_{tm}, y_{tm}), m = 1, 2 \). The positions of the 2 receiver antennas are similarly expressed as \( R_n = (x_{rn}, y_{rn}), n = 1, 2 \). The target location can be obtained from the solution of unknown variable \( X = (x, y) \) in the non-linear equation:

\[
\tau_{nm} = \frac{1}{c} \left\{ \sqrt{(x_{tm} - x)^2 + (y_{tm} - y)^2} + \sqrt{(x_{rn} - x)^2 + (y_{rn} - y)^2} \right\}
\]

where \( c \) represents the speed of light. \( n = 1, 2 \) and \( m = 1, 2 \). \( \tau_{nm} \) represents over-the-air delay from transmitter antenna \( m \) to receiver antenna \( n \).

To find \( X = (x, y) \), non-linear equations are solved by using Levenberg-Marquardt algorithm [11]. This iterative algorithm needs an initial position value. The initial value for the current illumination is set as the localization result of last illumination. For the first illumination, the initial value is set to the central position of the experiment field.

III. WAVEFORM DIVERSITY USING CONVEX OPTIMIZATION

In our MIMO cognitive radar system, the baseband sounding waveform at the \( n \)th transmitter antenna is \( p_n(t) \) and the corresponding baseband FIR filter at receiver is \( g_m(t) \).

There is an additive interference added to the received signal. The frequency position for the interference signal can be obtained by using spectrum sensing. For one each antenna, 1-bit decision is made for each spectrum bin. The sensing results from different antennas are combined by using the OR logic. In the following analysis, we consider each received signal is affected by the same interference with the frequency position same as the combined spectrum sensing result. The baseband interference signal in the scene is \( i(t) \).

We will not consider the effects of additive white Gaussian noise (AWGN). The target is considered as a static point target.

The baseband signal of the received sounding waveform from the first receiver antenna is

\[
r_1(t) = p_1(t - \tau_{11}) + p_2(t - \tau_{12}) + i(t)
\]

and the baseband signal from the second receiver antenna is

\[
r_2(t) = p_1(t - \tau_{21}) + p_2(t - \tau_{22}) + i(t)
\]

In order to obtain \( \tau_{nm}, r_n(t) \) will pass through the matched filter \( g_m(t) \) which means

\[
y_{nm}(t) = r_n(t) * g_m(t)
\]

\[
y_{nm}(t) = \left( \sum_{m'=1}^{M} p_{m'}(t - \tau_{nm'}) + i(t) \right) * g_m(t)
\]

where * denotes convolution operation. \( \tau_{nm} \) can be estimated by pinpointing the location of the peak of \( p_m(t - \tau_{nm}) * g_m(t) \).

In order to achieve accurate \( \tau_{11}, \tau_{12}, \tau_{21}, \) and \( \tau_{22} \), we need to

- maximize the peak value of \( p_m(t) * g_m(t) \);
- constrain the shape of \( p_m(t) * g_m(t) \) except the peak;
- constrain the shape of \( p_m(t) * g_{m'}(t) \) where \( m \neq m' \);
- suppress \( i(t) * g_m(t) \).

Fig. 4. The block diagram of the receiver RF component

Fig. 5. Transmitter software architecture

Fig. 6. Receiver software architecture

50 GHz. DPO directly samples the received RF signal. DPO starts to record the echoed signal at the same time when the sounding waveform starts to be transmitted. A bandpass filter is used to cancel the interference outside of signal bandwidth. The passband of the filter is 3500 to 4500 MHz, with 30 dB stopband attenuation. Samples are transmitted to MATLAB through gigabyte Ethernet.
The real part of \( \tilde{g}_m \) can be approximately represented as with only consideration of the location of \( p_m \) and \( g_m \) is \( L \). The correlation result, \( p_m \ast g_m \in C^{(2L-1)\times1} \). We always assume the peak of \( p_m \ast g_m \) is \( L \).

Given \( p_m \), assume

\[
p_m = \left[ (p_m)_{L,1} \cdots (p_m)_{1,1} \right]
\]

where \((\cdot)_{i,j}\) denotes the entry in the matrix with the \( i \)th row and the \( j \)th column.

Assume,

\[
\tilde{g}_m = [(\text{real}(g_m))^T (\text{imag}(g_m))^T]^T
\]

where real(\( \cdot \)) returns the real part of complex value and imag(\( \cdot \)) returns the imaginary part of complex value. Meanwhile \( T \) denotes the transpose operator.

Assume,

\[
\tilde{p}_m = [\text{real}(p_m) (-1)\text{imag}(p_m)]
\]

The maximization of the peak value of \( p_m(t) \ast g_m(t) \) can be approximately represented as with only consideration of the real part of \( p_m(t) \ast g_m(t) \),

\[
\text{maximize } \tilde{p}_m \tilde{g}_m
\]

The maximization of \( \tilde{p}_m \tilde{g}_m \) is the same as the maximization of \( (\tilde{p}_m \tilde{g}_m)^2 \) as long as \( \tilde{p}_m \tilde{g}_m \) is equal to or greater than zero. Furthermore,

\[
(\tilde{p}_m \tilde{g}_m)^2 = (\tilde{p}_m \tilde{g}_m)^T (\tilde{p}_m \tilde{g}_m) = (g_m)^T \tilde{p}_m \tilde{g}_m = (g_m)^T \tilde{p}_m \tilde{g}_m = \text{trace}(\tilde{p}_m \tilde{g}_m (g_m)^T) = \text{trace}(\tilde{p}_m \tilde{G}_m)
\]

where trace(\( \cdot \)) denotes matrix trace operator and \( \tilde{p}_m = (p_m)^T \tilde{p}_m \) as well as \( \tilde{G}_m = \tilde{g}_m (g_m)^T \). \( G_m \) should be rank-1 positive semidefinite matrix. However, rank constraint is non-convex constraint, which will be omitted in the following optimization problems. As a result, the optimization problem (8) to get \( \tilde{g}_m \) can be reformulated as,

\[
\text{maximize } \text{trace}(\tilde{p}_m \tilde{G}_m)
\]

Similarly if \( g_m \) is given, the maximization of the peak value of \( p_m(t) \ast g_m(t) \) to get \( p_m \) can be represented as,

\[
\text{maximize } \text{trace}(\tilde{G}_m \tilde{p}_m)
\]

where

\[
\tilde{g}_m = \left[ (g_m)_{L,1} \cdots (g_m)_{1,1} \right],
\]

\[
\tilde{g}_m = [\text{real}(g_m) (-1)\text{imag}(g_m)],
\]

\[
\tilde{p}_m = [(\text{real}(p_m))^T (\text{imag}(p_m))^T]^T,
\]

\[
\tilde{G}_m = (\tilde{g}_m)^T \tilde{g}_m \text{ as well as } \tilde{P}_m = \tilde{p}_m (p_m)^T.
\]

The methodologies of the shape constraints for both \( p_m(t) \ast g_m(t) \) except the peak and \( p_m(t) \ast g_m'(t) \) where \( m \neq m' \) are the same. We take the total energy constraint of \( p_m(t) \ast g_m'(t) \) as an example here.

Given \( p_m \), construct the Toeplitz matrix \( F_m \) as,

\[
(F_m)_{i,j} = \begin{cases} (p_m)_{i-j+1,1}, & 0 \leq i - j \leq L - 1 \\ 0, & \text{else} \end{cases}
\]

Define,

\[
(F_m \ast g_m')_{i,j} = \begin{cases} \textbf{tr}(F_m \ast g_m')_{i,j}, & (i,j) \in \{0,1\} \\ 0, & \text{else} \end{cases}
\]

Thus,

\[
p_m \ast g_m' = F_m \ast g_m' = F_m \ast \tilde{g}_m' \]

and the total energy constraint for cross-correlation is

\[
\|p_m \ast g_m'\|_2^2 = \text{trace}(\tilde{G}_m \ast \tilde{G}_m') = \text{trace}(F_m \ast (\tilde{g}_m')^H \tilde{G}_m') \leq E_{mm'}
\]

where \( H \) denotes transpose conjugate operator.

Similarly, if \( g_m \) is known, the total energy constraint for cross-correlation is

\[
\|p_m' \ast g_m\|_2^2 = \text{trace}(\tilde{G}_m' \ast \tilde{G}_m) \leq E_{mm'}
\]

In order to suppress \( i(t) \ast g_m(t) \), we can notch the spectrum bands occupied by \( i(t) \) in \( p_m(t) \) and \( g_m(t) \). The spectrum position of \( i(t) \) is needed. More information of interference, like amplitude and phase information, is not required. The design of \( p_m(t) \) and \( g_m(t) \) with notched spectrum bands can be incorporated into the convex optimization framework.

Assume \( F \) is the discrete-time Fourier transform operator

\[
\tilde{F} = [F \sqrt{-1}F].
\]

We can get the frequency domain representation of \( p_m \) as

\[
p_m' = \tilde{F}p_m = \tilde{F}p_m
\]

If the \( l \)th row of \( \tilde{F} \) is \( \tilde{f}_l \), each entry in \( p_m' \) can be represented by

\[
(p_m')_{l,1} = \tilde{f}_l p_m, \quad l = 1, 2, \ldots, L
\]

Define,

\[
\tilde{F}_l = \tilde{f}_l^H \tilde{f}_l, \quad l = 1, 2, \ldots, L
\]
Given the support set $\Omega_{\text{notch}}$ of spectrum bands occupied by interference, when $l \in \Omega_{\text{notch}}$:

$$
\left| (p^f_m)_{l,1} \right|^2 = \left| \tilde{f}_l \tilde{p}_m \right|^2 = \tilde{p}_m^T \tilde{f}_l^H \tilde{f}_l \tilde{p}_m = \tilde{p}_m^T \tilde{F}_l \tilde{p}_m = \text{trace}(\tilde{F}_l \tilde{p}_m (\tilde{p}_m)^T)
$$

$$
= \text{trace}(\tilde{F}_l \tilde{p}_m)
\leq \delta_{\text{notch}}
$$

(24)

where $\delta_{\text{notch}}$ is the small positive value. Similarly,

$$
\left| (g^f_m)_{l,1} \right|^2 = \text{trace}(\tilde{F}_l \tilde{G}_m)
\leq \delta_{\text{notch}}, \ l \in \Omega_{\text{notch}}
$$

(25)

Finally, the total energies of $p_m$ and $g_m$ should be bounded,

$$
\| p_m \|^2 = \| \tilde{p}_m \|^2 = (\tilde{p}_m)^T \tilde{p}_m = \text{trace}(\tilde{p}_m (\tilde{p}_m)^T)
= \text{trace}(\tilde{p}_m)
\leq E_p
$$

(26)

where $E_p$ is the energy upper bound for $p_m$. Similarly,

$$
\| g_m \|^2 = \text{trace}(\tilde{G}_m)
\leq E_g
$$

(27)

where $E_g$ is the energy upper bound for $g_m$.

Due to the operation of $p_m \ast g_m$, it is hard to optimize $p_m$ and $g_m$ simultaneously. A semidefinite programming (SDP) based alternative algorithm is derived to get a feasible and optimal solution to $p_m$ and $g_m$.

Given $p_m$, we can get $G_m$ by solving the following SDP,

maximize $\text{trace}(\tilde{P}_m \tilde{G}_m)$
subject to
$\text{trace}(\tilde{F}_l \tilde{G}_m) \leq \delta_{\text{notch}}, \ l \in \Omega_{\text{notch}}$
$\text{trace}((\tilde{P}_m^\text{TToeplitz})^H \tilde{P}_m^\text{TToeplitz} \tilde{G}_m) \leq E_{m' m}, \ m' \neq m$
$\text{trace}(\tilde{G}_m) \leq E_g$

(28)

g_m can be achieved through the optimized $\tilde{G}_m$.

Given $g_m$, we can get $\tilde{P}_m$ by solving the following SDP,

maximize $\text{trace}(\tilde{G}_m \tilde{P}_m)$
subject to
$\text{trace}(\tilde{F}_l \tilde{P}_m) \leq \delta_{\text{notch}}, \ l \in \Omega_{\text{notch}}$
$\text{trace}((\tilde{G}_m^\text{TToeplitz})^H \tilde{G}_m^\text{TToeplitz} \tilde{P}_m) \leq E_{m' m}, \ m' \neq m$
$\text{trace}(\tilde{P}_m) \leq E_p$

(29)

$p_m$ can be achieved through the optimized $\tilde{P}_m$.

In summary, an SDP-based alternative algorithm can be described in algorithm 1.

**Algorithm 1 Waveform Optimization Algorithm**

1: Initialize $p_1$ and $p_2$ using the standard PN sequences.
2: while $p_1$ and $p_2$ do not converge: do
3: Solve the optimization problem (28) to get $g_1$ for fixed $p_1$ and $p_2$.
4: Solve the optimization problem (28) to get $g_2$ for fixed $p_1$ and $p_2$.
5: Solve the optimization problem (29) to get $p_1$ for fixed $g_1$ and $g_2$.
6: Solve the optimization problem (29) to get $p_2$ for fixed $g_1$ and $g_2$
7: end while

![Fig. 7. The position of target and antennas](image-url)

**IV. PERFORMANCE EVALUATION**

In the experiment scenario, as shown in Fig. 7, one target is placed in an area of 9 m². Two target positions are tested, namely the target position 1 and 2.

As the first step, the interference signal is generated by MATLAB and added to the received baseband signal. Next step, AWG will be used to generate the interference signal.

Localization performance is obtained by running the system under different signal to interference ratio (SIR), which is defined as

$$
\text{SIR}_{\text{dB}} = 10 \log_{10} \left( \frac{P_{\text{signal}}}{P_{\text{interference}}} \right)
$$

(30)

Initially, two PN sequence highly orthogonal to each other are selected to be the sounding pulses. The received signal is correlated with a matched filter.

Based on the knowledge of interference position, a new pulse generated by convex optimization is transmitted as the sounding pulse. The total energy constraint for cross-correlation $E_{mm'}$ is set to 1. Waveform total energy constraints, $E_p$ and $E_g$, are set to 1. Spectrum band constraint $\delta_{\text{notch}}$ is set to $10^{-10}$. Fig. 8 shows the spectral power of a waveform generated by convex optimization. One is the spectral power of the digital baseband signal, the other is the spectral power of the RF signal shown in spectrum analyzer.

Performance curve in Fig. 9 is obtained in target position 1. Fig. 10 is obtained in target position 2. Each position is tested in two interference bandwidth, 47.62 MHz and 79.37 MHz,
9.52% and 15.87% respectively of the total sounding signal bandwidth. Root mean square error (RMSE) of localization results is obtained by conducting 1000 trials for one SIR. In different positions, the localization performance is different due to the different channel conditions. In all these conditions, waveform optimization leads to apparently better performance. Under 0 dB SIR, the RMSE can be improved from 5 meters to about 1 meter. When the interference is focused on narrower frequency band, the performance of waveform optimization is better since the optimized waveform has broader bandwidth.

V. Conclusion

This paper has discussed the experimental demonstration of a cognitive radar system for target localization. An automatic realtime closed-loop radar is built to support experimental study of the radar system in real-world situations. In order to avoid reserved frequency bands or narrowband interferences, the transmitted waveforms and the corresponding receiver filters are jointly designed to have low correlation sidelobes and with spectral power suppressed in arbitrary frequency bands. Target localization is demonstrated in an experimental scenario with strong interferences. Experimental performance of the cognitive radar performance is provided with different target locations and different interference bandwidth scenarios. The performance of the system with waveform optimization capability is better than that without waveform optimization. As the next step, AWG will be used to generate the interference signal instead of simulating the signal using MATLAB. Spectrum sensing will be used to determine the frequency position of the interference signal. Waveform optimization will be conducted based on the spectrum sensing result.

ACKNOWLEDGMENT

This work is funded by National Science Foundation through two grants (ECCS-0901420 and ECCS-0821658), and Office of Naval Research through two grants (N00010-10-1-0810 and N00014-11-1-0006).

REFERENCES