Joint Beamforming Design in DFRC Systems for Wideband Sensing and OFDM Communications

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Abstract—Dual-function radar-communication (DFRC) systems, which can efficiently utilize the congested spectrum and costly hardware resources by employing one common waveform for both sensing and communication (S&C), have attracted increasing attention. While the orthogonal frequency division multiplexing (OFDM) technique has been widely adopted to support high-quality communications, it also has great potentials of improving radar sensing performance and providing flexible S&C. In this paper, we propose to jointly design the dual-functional transmit signals occupying several subcarriers to realize multi-user OFDM communications and detect one moving target in the presence of clutter. Meanwhile, the signals in other frequency subcarriers can be optimized in a similar way to perform other tasks. The transmit beamforming and receive filter are jointly optimized to maximize the radar output signal-to-interference-plus-noise ratio (SINR), while satisfying the communication SINR requirement and the power budget. An majorization-minimization (MM) method based algorithm is developed to solve the resulting non-convex optimization problem. Numerical results reveal the significant wideband sensing gain brought by jointly designing the transmit signals in different subcarriers, and demonstrate the advantages of our proposed scheme and the effectiveness of the developed algorithm.

Index Terms—DFRC, OFDM, wide-band sensing, beamforming design.

I. INTRODUCTION

Dual-functional radar-communication (DFRC), which enjoys high spectrum/power/hardware efficiency through sharing one common waveform for both sensing and communication (S&C), is regarded as one key enabling technique for the next-generation wireless networks [1]. Along with the integration and coordination gains, sophisticated designs for the dual-functional waveform are required to handle the conflict requirements of communication and sensing. Towards this goal, the multiple-input multiple-output (MIMO) architecture has been widely employed in DFRC systems, and the transmit beamforming designs to exploit the spatial degrees of freedom (DoFs) for achieving better performance trade-offs have attracted growing attentions in recent years [2]-[5]. While most of existing works focused on narrowband systems, the research on exploiting wideband signals to improve both sensing and communication performance has begun [6]-[11]. A time-frequency waveform design problem was investigated in [6] by optimizing the subcarriers powers in the presence of a low-rate feedback channel for conveying transmit waveform control information. The authors in [7] proposed a subcarrier sharing scheme to efficiently exploit the available bandwidth in the orthogonal frequency division multiplexing (OFDM) DFRC system, in which certain subcarriers are shared for realizing S&C and the other private subcarriers are reserved for exclusive use. To achieve better S&C performance, beamforming designs for the transmit signals in different subcarriers should be considered. The authors in [8] investigated the low peak-to-average power ratio (PAPR) waveform design problem. The hybrid beamforming with finite-resolution phase shifters was studied in [9], [10]. The authors in [11] considered the design for wideband broadcast systems under communication error rate and beampattern constraints. In addition, the deployment of intelligent reflecting surface (IRS) in OFDM-DFRC systems was investigated in [12] to exploit the passive beamforming gain for better S&C performance. Although these works verified the advantages of wideband signals in improving both sensing and communication performance, the adopted radar sensing model is very simplified. Moreover, simply focusing on the joint design of all subcarriers will cause unaffordable computational complexity with limited performance improvement, which is very inefficient in practical OFDM-DFRC systems.

Motivated by above findings, in this paper we propose a more flexible scheme to perform S&C in OFDM-DFRC systems. Specifically, the dual-functional signals occupying several frequency subcarriers are jointly designed for multi-user OFDM communications and detecting one moving target in the presence of widely spread clutter, while the signals in other frequency subcarriers are optimized in a similar way...
to perform other tasks. After establishing practical models for both sensing and communication functionalities, we investigate to maximize the radar output signal-to-interference-plus-noise ratio (SINR) as well as satisfying the communication SINR requirement and the power budget by jointly optimizing the transmit beamforming and receive filter. An efficient algorithm based on the majorization minimization (MM) method is developed to solve the resulting complicated optimization problem. Finally, numerical results reveal the significant sensing gain brought by wideband signals and the advantages of the proposed scheme associated with the developed algorithm.

II. SYSTEM MODEL AND PROBLEM FORMULATION

We consider a wideband DFRC system as depicted in Fig. 1, where a dual-functional BS equipped with two uniform linear arrays (ULAs) of $N_t$ transmit antennas and $N_r$ receive antennas performs S&C using $N'$ subcarriers. In particular, the BS occupies several frequency subcarriers of number $N$ to simultaneously serve $K$ single-antenna users and detect one target located at the azimuth direction $\theta_0$ with speed $v_0$ in the presence of widespread clutter.

It is noted that we only use $N \leq N'$ subcarriers to detect one target and communicate with users, and the residual subcarriers can be flexibly utilized to perform other dual-functional tasks in a similar way. This scheme is feasible since the orthogonality between different subcarriers naturally supports transmitting different data streams to the users without interference. As for the sensing functionality at the radar receiver, the echoes generated from the considered subcarriers can be easily separated by a bandpass filter, while the interference caused by the adjacent frequency subcarriers with Doppler frequency shifts due to the targets/clutter is negligible. For example, the echoes in the subcarrier of $f = 2.4$GHz reflected by a relatively fast target with speed $v = 50$m/s induce the Doppler frequency $f_d = 2v f/c = 800$Hz, while a typical OFDM system usually has $N' \geq 50$ subcarriers [13], each of which has the bandwidth $\Delta f' \geq 50$kHz $\gg f_d$. In addition to enabling more flexibilities, the proposed scheme is also necessary in practical applications, since the computational complexity for jointly designing the transmitted signals in different subcarriers will become prohibitively higher with the increase of $N$, while at the same time the performance improvement in target detection introduced by increased subcarriers is very limited.

A. Transmitted Signal Model

In specific, we denote $s_n[l] \in \mathbb{C}^K$, $n = 1, \ldots, N$, $l = 1, \ldots, L$, as the symbol vector of the $n$-th subcarrier in the $l$-th symbol time-slot for the $K$ users, where $\mathbb{E}\{s_n[l] s_n[l]^H\} = \mathbf{I}_K$, and $L$ denotes the radar pulse/communication frame length. The symbol vector $s_n[l]$ is passed through a well-designed beamforming matrix $\mathbf{W}_n \triangleq [w_{n,1}, \ldots, w_{n,K}] \in \mathbb{C}^{N_r \times K}$, which generates the baseband frequency-domain transmitted signal as

$$\tilde{x}_n[l] = \mathbf{W}_n s_n[l], \quad \forall n. \quad (1)$$

After $N$-point inverse discrete Fourier transform (IDFT), the cyclic prefix (CP) of size $N_{cp}$ is added with duration $T_{cp}$, to avoid inter-symbol interference (ISI). Then, through digital to analog converter (DAC), $\tilde{x}_n[l], \forall n$, are transformed into the baseband analog temporal-domain signal as

$$x(t) = \sum_{n=1}^{N} \mathbf{W}_n s_n[l] e^{j2\pi(n-1)\Delta f t}, \quad (2)$$

where $t \in ((l-1)(T_s + T_{cp}),(l-1)(T_s + T_{cp}) + T_s], \forall l$, $\Delta f$ and $T_s$ are the frequency interval of the OFDM signaling and the OFDM symbol duration, respectively, and we assume $\Delta f = 1/T_s$ to guarantee the orthogonality between different subcarriers. Finally, the signal is up-converted to the radio frequency (RF) domain via $N_t$ RF chains with carrier frequency $f_c$ [9] and then emitted through the $N_t$ antennas.

B. OFDM Communication Model

The wideband channel from the BS to user $k$ is modeled by a $D$-tap ($D \leq N_{cp}$) finite-duration impulse response $\{h_{k,1}, \ldots, h_{k,D}\}$ [14], where $h_{k,d} \in \mathbb{C}^{N_r}$, $\forall d$ are assumed to be perfectly known at the BS. After down-converting, removing the CP, and $N$-point DFT, the received baseband frequency-domain signal for user $k$ in the $n$-th subcarrier is obtained as

$$\tilde{y}_{n,k}[l] = \tilde{h}_{n,k}^H \mathbf{W}_n s_n[l] + \tilde{z}_{n,k}, \quad (3)$$

where $\tilde{h}_{n,k} \in \mathbb{C}^{N_r}$ denotes the frequency-domain communication channel for user $k$, $\tilde{z}_{n,k} \sim \mathcal{CN}(0, \sigma^2)$ is the additive white Gaussian noise (AWGN) at the $k$-th user. The corresponding SINR of the $k$-th user in the $n$-th subcarrier can be written as

$$\text{SINR}_{c,n,k} = \frac{|\tilde{h}_{n,k}^H \mathbf{W}_n s_n|^2}{\sum_{j \neq n} |\tilde{h}_{n,k}^H \mathbf{W}_n s_n|^2 + \sigma^2}, \forall k, \forall n. \quad (4)$$

C. Wideband Radar Model

The received RF echo from the target at the received array can be expressed as
\[
y^\text{RF}(t) = \sum_{n=1}^{N} a_{0,n} b(\theta_0, f_n) a^T(\theta_0, f_n) \mathbf{W}_s \mathbf{s}_n [e^{j 2 \pi (f_n + d_0) t} - t_0],
\]
where \( t = t_0 \in ((l - 1)T + c, (l - 1)T + T_s), \forall l \), \( a_{0,n} \) represents the target reflection coefficient in the \( n \)-th subcarrier with \( \mathbb{E}\{ |a_{0,n}|^2 \} = \sigma_{a,0}^2, f_n \triangleq (n - 1)T + f_c \) denotes the frequency of the \( n \)-th subcarrier, the scalar \( t_0 \) is the two-way propagation delay, \( f_{0,n} = 2v_0f_n/c \) is the target Doppler frequency of the \( n \)-th subcarrier with \( c \) representing the velocity of light, and \( T_{s+c} \triangleq T_s + T_c \) for simplicity. \( a(\theta, f) \) and \( b(\theta, f) \) respectively denote the spatial-frequency steering vector of the transmit and receive signals and are defined as
\[
a(\theta, f) \triangleq [1, e^{-j 2 \pi d_1 \sin \theta / \lambda}, \ldots, e^{-j 2 \pi (N_l - 1) d_1 \sin \theta / \lambda}]^T, \label{eq:a}
\]
\[
b(\theta, f) \triangleq [1, e^{-j 2 \pi d_t \sin \theta / \lambda}, \ldots, e^{-j 2 \pi (N_l - 1) d_t \sin \theta / \lambda}]^T, \label{eq:b}
\]
where \( d_t \) and \( d_c \) denote the transmit and receive antenna spacing, respectively, and \( \lambda \triangleq c/f \) denotes the wavelength.

After down-converting, the baseband temporal-domain echo from the target is written as
\[
y_0(t) = \sum_{n=1}^{N} a_{0,n} b(\theta_0, f_n) a^T(\theta_0, f_n) \mathbf{W}_s \mathbf{s}_n [e^{j 2 \pi f_n (t - t_0)}], \label{eq:y0}
\]
where \( f_{0,n} \triangleq f_n + f_{0,r} - f_c \) denotes the frequency of the \( n \)-th baseband signal of the target echo and we absorb the constant phase terms associated with \( t_0 \) into the target amplitude for simplicity. By sampling \( y_0(t) \) \( N_s \) times during each OFDM symbol, the baseband digital samples in the \( L \)-th time slot can be obtained as
\[
\mathbf{Y}_0[l] \triangleq \left[ \sum_{n=1}^{N} a_{0,n} e^{j 2 \pi f_n (t - l) T_U} \mathbf{p}(f_{0,n}) \mathbf{a}(\theta_0, f_n) a^T(\theta_0, f_n) \mathbf{W}_s \mathbf{s}_n[l] \right], \label{eq:y0l}
\]
D. Problem Formulation

Since the target detection probability is positively related to the radar output SINR, in this paper we aim to maximize \( \text{SINR}_r \), while meeting the communication SINR requirement and the power budget by jointly designing the transmit beamforming \( \mathbf{W}_n, \forall n \), and the receive filter \( \mathbf{w}_r \). The optimization problem is thus formulated as

\[
\begin{align*}
\max_{\mathbf{W}_n, \mathbf{w}_r} & \quad \text{SINR}_r \\
\text{s.t.} & \quad \text{SINR}_{c,n,k} \geq \Gamma_c, \quad \forall k, \forall n, \\
& \quad \sum_{l=1}^{L} \| \mathbf{W}_n s_n[l] \|^2_2 \leq P_{t,n}, \quad \forall n.
\end{align*}
\]

where \( \Gamma_c \) is the communication SINR requirement and \( P_{t,n} \) is the power budget for the transmitted signals in the \( n \)-th subcarrier. Due to the non-convex objective function expressed in (16) and the complicated fractional terms in (17a), (17b), it is very difficult to directly obtain the solution to problem (17).

In the next section, we propose an efficient algorithm based on some sophisticated derivations and the MM method to convert problem (17) into a sequence of favorable sub-problems and iteratively solve them.

III. Joint Transmit Beamforming and Receive Filter Design

A. Problem Reformulation

In order to efficiently solve problem (17), in this subsection we propose to re-formulate it as a more favorable form with less optimization variables. It is clear that there is no constraint on the radar receiver filter \( \mathbf{w}_r \), i.e., with fixed transmit beamforming \( \mathbf{W}_n, \forall n \), the optimization problem for \( \mathbf{w}_r \) can be written as an unconstrained problem:

\[
\begin{align*}
\max_{\mathbf{w}_r} \quad \mathbf{w}_r^H \mathbf{X}_m \mathbf{W}_n \mathbf{X}_r^H \mathbf{w}_r \\
\text{s.t.} & \quad \{ \mathbf{w}_r \} \leq \mathbf{s}_n, \forall n,
\end{align*}
\]

This is a well-known generalized Rayleigh quotient whose optimal solution \( \mathbf{w}_r^* \) can be easily obtained as

\[
\mathbf{w}_r^* = \mathbf{X}_m \mathbf{W}_n \mathbf{X}_r^H \mathbf{w}_r^*.
\]

By substituting \( \mathbf{w}_r^* \) into the original problem (17), the joint transmit beamforming and receive filter design problem is reduced to the beamforming design problem as

\[
\begin{align*}
\min_{\mathbf{W}_n, \mathbf{w}_r} & \quad \mathbf{w}_r^H \mathbf{X}_m \mathbf{W}_n \mathbf{X}_r^H \mathbf{w}_r \\
\text{s.t.} & \quad \{ \mathbf{w}_r \} \leq \mathbf{s}_n, \forall n,
\end{align*}
\]

We observe that the objective function is formulated with respect to the matrix \( \mathbf{X}_m \), which implicitly contains the optimizing variables \( \mathbf{W}_n, \forall n \). To facilitate the following algorithm development, it is necessary to equivalently convert (20a) into an explicit expression with respect to the transmit beamforming \( \mathbf{W}_n, \forall n \).

Using the definitions in (10d) and (10e), the term \( \mathbf{X}_v \) in the objective function (20a) can be re-written as

\[
\mathbf{X}_v = \sum_{n=1}^{N} \mathbf{X}_n v_{0,n}.
\]

Based on the definition of \( \mathbf{X}_n \) in (10a), the term \( \mathbf{X}_n v_{0,n} \) is re-arranged as

\[
\mathbf{X}_n v_{0,n} = \begin{bmatrix} (I_{N_c,N_r} \otimes (s_{n}[1]^H \mathbf{W}_n^H)) v_{0,n,1} \\ (I_{N_c,N_r} \otimes (s_{n}[2]^H \mathbf{W}_n^H)) v_{0,n,2} \\ \vdots \\ (I_{N_c,N_r} \otimes (s_{n}[L]^H \mathbf{W}_n^H)) v_{0,n,L} \end{bmatrix}
\]

where \( v_{0,n,l} \equiv [v_{0,n,1}, \ldots, v_{0,n,L}]^T \) with the \( l \)-th subvector \( v_{0,n,l} \in \mathbb{C}^{N_c \times N_r} \). According to the properties of the Kronecker product, the \( l \)-th term in (22) can be further transformed into

\[
(I_{N_c,N_r} \otimes (s_n[l]^H \mathbf{W}_n^H)) v_{0,n,l} = vec(s_n[l]^H \mathbf{W}_n^H \mathbf{v}_{0,n,l}) = \mathbf{V}_{0,n,l} \mathbf{W}_n s_n[l],
\]

where \( \mathbf{V}_{0,n,l} \in \mathbb{C}^{N_r \times N_r \times N_c} \) is a reshaped version of \( v_{0,n,l} \), i.e.,

\[
v_{0,n,l} = vec(\mathbf{V}_{0,n,l}).
\]

In order to exploit similar derivations for (25) to handle the term \( \mathbf{J}_m \mathbf{X}_m \mathbf{X}_r^H \mathbf{J}_m^H \) in (20a), we first utilize the eigenvalue decomposition to split the semi-definite matrix inner CCMs \( \mathbf{M}_m, \forall m \) as

\[
\mathbf{M}_m = \sum_{r=1}^{\text{rank}(\mathbf{M}_m)} \gamma_{m,r} \mathbf{u}_{m,r} \mathbf{u}_{m,r}^H,
\]

where \( \gamma_{m,r} \) denotes the non-zero eigenvalue of \( \mathbf{M}_m \) with the corresponding eigenvector \( \mathbf{u}_{m,r} \), and we define \( \mathbf{u}_{m,r} = \sqrt{\gamma_{m,r}} \mathbf{u}_{m,r} \) for simplicity. Then, following the transformation procedures in (21)-(25), the term \( \mathbf{J}_m \mathbf{X}_m \mathbf{X}_r^H \mathbf{J}_m^H \) can be equivalently expressed with respect to \( \mathbf{w} \) as

\[
\mathbf{J}_m \mathbf{X}_m \mathbf{X}_r^H \mathbf{J}_m^H = \sum_{r=1}^{\text{rank}(\mathbf{M}_m)} \mathbf{J}_m \mathbf{X}_m \mathbf{u}_{m,r} \mathbf{u}_{m,r}^H \mathbf{X}_r^H \mathbf{J}_m^H
\]
where we define $T_{m,r} \triangleq J_{m,r}(S^T \otimes I_{N_s})_r, \ldots, U_{m,r,N}$, $U_{m,r,n} \triangleq \text{BldDiag}(U_{m,r,1}, \ldots, U_{m,r,n,L})$, $U_{m,r,n,l} \in C_{N_sN_r}$, and $u_{m,r,n}$ is the $n$-th sub-vector of $u_{m,r} \in C_{N_sN_rN_t}$. Therefore, substituting the results in (25) and (27b) into (20a), the beamforming design problem can be explicitly and equivalently re-formulated as
\[
\min_w - w^H T_0 A^{-1}(w) T_0 w \quad \text{s.t. } \text{SINR}_{c,n,k} \geq \Gamma_c, \quad \forall k, \forall n,
\]
\[
\sum_{l=1}^L \|W_n s_n[l]\|^2 \leq P_{t,n}, \quad \forall n,
\]
where $A(w) \triangleq \sum_{m=-M}^M \sum_{r=1}^{\text{rank}(M_m)} T_{m,r} w w^H T_{m,r}^H + \sigma^2 I$.

**B. MM-based Transformation**

The beamforming design problem (28) is still difficult to solve due to the complicated non-convex objective function (28a). In this subsection, we propose to utilize the MM method to construct a sequence of more tractable problems to be optimized until convergence. Specifically, in the $t$-th iteration, a convex surrogate function is constructed to approximate the objective function (28a) and serve as an upper-bound that should be minimized in the next iteration. According to *Lemma 1* in [17], an upper-bound surrogate function for (28a) can be obtained by
\[
(28a) \leq \text{Tr}\{A^{-1}(w_t) T_0 w_t w_t^H T_0^H A^{-1}(w_t) A(w)\} - 2\text{Tr}\{w_t^H T_0 A^{-1}(w_t) T_0 w\} + c_1
\]
\[
= w^H U_t w - \mathfrak{R}\{b_t^H w\} + c_2,
\]
where $c_1$ and $c_2$ are constant terms that are irrelevant to variable $w$, $w_t$ denotes the obtained solution in the $t$-th iteration, and
\[
U_t \triangleq \sum_{m=-M}^M \sum_{r=1}^{\text{rank}(M_m)} T_{m,r}^H A^{-1}(w_t) T_0 w_t w_t^H T_0 A^{-1}(w_t) T_{m,r},
\]
\[
b_t \triangleq 2 T_0 A^{-1}(w_t) T_0 w_t.
\]
With the above derivations, the optimization problem in each iteration is formulated as
\[
\min_w w^H U_t w - \mathfrak{R}\{b_t^H w\} \quad \text{s.t. } \text{SINR}_{c,n,k} \geq \Gamma_c, \quad \forall k, \forall n,
\]
\[
\sum_{l=1}^L \|W_n s_n[l]\|^2 \leq P_{t,n}, \quad \forall n.
\]

**Algorithm 1 Joint Transmit Beamforming and Receive Filter Design Algorithm**

**Input:** $T_0, T_{m,r}, \forall m, \forall r, P_{t,n}, \tilde{h}_{n,k}, \forall k, \forall n, \sigma, \Gamma_c, \epsilon.$

**Output:** $w^*, w_t^*$.

1. Initialize $t := 0$, initialize $w_0$ by solving (32).
2. repeat
3. Update $U_t$ and $b_t$ by (30);
4. Update $w_{t+1}$ by solving (31);
5. $t := t + 1$;
6. until $\|w_t - w_{t-1}\|_2/\|w_{t-1}\|_2 \leq \epsilon$.
7. $w^* := w_t$.
8. Calculate $w_t^*$ by (19).

**C. Summary**

Now the proposed joint transmit beamforming and receive filter design algorithm is straightforward and summarized in Algorithm 1, where $\epsilon$ is a parameter to judge the convergence. In summary, the transmit beamforming $w$ is iteratively updated by solving (31) until convergence and then the receive filter $w_t^*$ is calculated by (19). In addition, to provide more flexibility for maximizing radar SINR and ensure a good convergence, we initialize $W_n$, $\forall n$ by maximizing the minimum communication SINR while satisfying the power budget as
\[
\max_{W_n} \min_{k} \text{SINR}_{c,n,k} \quad \text{s.t. } \sum_{l=1}^L \|W_n[s_n[l]]\|_2^2 \leq P_{t,n}/L,
\]
which is a well-known SINR balancing problem and can be easily solved [18].

**IV. SIMULATION RESULTS**

In this section, we provide simulation results to illustrate the performance of our proposed algorithm in a wideband DFRC system. We assume that the BS equipped with $N_t = N_r = 4$ antennas with spacing $d_t = 2c/f_c$, $d_r = 0.5c/f_c$, serves the $K = 3$ users and detects one target located at the azimuth $\theta_0 = 0^\circ$ with speed $v_0 = 20$m/s and reflection power $\sigma_0 = -10$dB, $\forall n$. The clutter is reflected from 5 range cells with $N_c = 30$ patches, which are uniformly distributed in the azimuth range $(0^\circ, 360^\circ)$ and the velocity range $(0, 50)$ m/s with power $\sigma_{c,n} = -10$dB, $\forall n$. The AWGN power at the users and the radar receive is $\sigma^2 = -20$dB and $\sigma_0^2 = -10$dB, respectively. The carrier frequency is $f_c = 2.4$GHz, the subcarrier spacing is $\Delta f = 0.2$MHz, the OFDM data symbol duration is $T_s = 5\mu s$, the CP duration $T_{cp} = 2\mu s$, and the frame length is $L = 8$. In addition, the parameter for judging convergence is set as $\epsilon = 1e-4$.

To illustrate the sensing gain brought by the multiple subcarriers in wideband systems, we first plot radar output SINR versus the number of subcarriers in Fig. 2, where the power budget for each subcarrier is set as $P_{s,n} = P_t/N$ with $P_t$ representing the total power for the system, and the number of sampling is $N_s = 5$. It is obvious that the radar output SINR improves with the increase of the number of subcarriers or the
V. Conclusions

In this paper, we investigated the joint beamforming design for a wideband DFRC system, in which several subcarriers are exploited to simultaneously detect one target in the presence of clutter and serve multiple users. The transmit beamforming and receive filter were jointly optimized to maximize the radar SINR while satisfying the communication SINR requirement and the power budget. Numerical results demonstrated the remarkable performance improvement brought by wideband signals and revealed the rationality of the proposed scheme.

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