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***Abstract*—In this paper, we discuss the possibility to suppress interference for wideband multiple-input-multiple-output (MIMO) radar, using only the temporal properties of the signals. The idea is to use tunable filters each connected to a wideband waveform generator, and to derive the optimal power spectral density (PSD) of the resulting signals in a known environment. The metric used to evaluate the enhancement in the system performance is the signal-to-noise and interference ratio (SNIR), from which the optimal transmit and receive filter properties are derived. We discuss two optimization approaches: one alternating and one joint algorithm. Each method is separated into two cases: for a total power constraint and for an individual power constraint on the transmit filters, respectively. Numerical validation illustrates the possibility to suppress the interference in the temporal domain, without actually forming a spatial null in the direction of the interference.**

I. Introduction

Multiple-input-multiple-output (MIMO) radar has in the recent years drawn considerable interest. A comparison between the related MIMO communication area and MIMO radar is presented in [1]. Further, an investigation of how MIMO antenna systems can potentially improve the radar performance is discussed from a general point of view in [2].

In this paper, we seek to find the optimal signal design to suppress interference. In the literature there are two main design approaches, where the first focuses on the spatial properties of the transmit signal and the second concerns the temporal properties of the transmitter–receiver chain. For the first approach, the problem is usually expressed as an optimization of the spatial correlations of the waveforms. Research so far concerns mainly narrowband radar, see e.g. [3] [4], where the optimization procedure may involve finding the covariance matrix of the waveforms that coincides with a desirable transmit beampattern. For the wideband case the problem is reformulated as matching the cross spectral density matrix to a desired spatial beampattern [5]. In [6] the signals are instead described by the Fourier transform of the beampattern.

The focus of our work is on the second case, where the temporal properties of the signals are used to suppress interference. This is typically achieved by using tunable filters at the transmitter and receiver sides. In the area of MIMO communication, a multitude of studies concerning this design process have been performed, e.g. the optimal design of space-time precoders and decoders is described in [7], and in [8] methods to design complex relay networking beamforming weights are addressed. A similar problem has been considered for radar [9], where an alternating method to design the transmit and receive filters for an extended target in clutter is discussed.

In this paper, we propose two algorithms for optimizing the transmit and receive filters for a known scenario. The environment may include the target(s) of interest, interference and clutter. The focus is on signal-dependent interference (smart jamming). Performance is measured at the receiver output as a function of the signal-to-noise-and-interference ratio (SNIR). Note that the investigation is a first step to reach a fully adaptive system, where the waveforms are continuously optimized to a time-variant scenario. Moreover, a combination between a spatial and a temporal optimization may further enhance the system performance.

Notation: Time domain samples are denoted by lowercase letters, vectors by boldface lowercase letters and matrices with boldface uppercase letters. Underlined samples, vectors and matrices are used to emphasize a signal transformed into the frequency domain. The transpose of a vector or a matrix is denoted $(\cdot)^T$, the complex conjugate as $(\cdot)^C$ and the complex conjugate transpose as $(\cdot)^H$.

II. Problem Formulation

Consider the radar system shown in Fig. 1, with K transmit antenna subarrays, each equipped with K_s elements and L receive antenna subarrays, each equipped with L_s elements. Note that no spatially adaptive filters are used at the receiver side. This ensures that no cancellation directions are formed in the receive pattern. Let $v_k(t) = c_k(t)e^{j2\pi f_c t}$ denote the

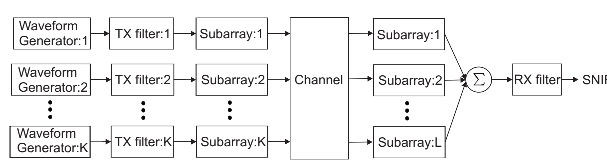

Fig. 1. Schematic view of the system model.

continuous time domain signal sequence, where $c_k(t)$ is an arbitrary signal generated from waveform generator k with bandwidth B, $k = 1 \ldots K$ and f_c being the carrier frequency. Sampling $c_k(t)$ at $t = nT_s$, where $T_s = \frac{1}{B}$ gives the discrete baseband version of the signal, $c_k[n]$, $n = 0 \ldots N-1$, which is assumed to be a white noise sequence. Each waveform generator is connected to a transmit filter, here modeled as a finite impulse response (FIR) filter with an order of P, yielding

$$s_k[n] = \sum_{p=0}^{P} b_{k,p} c_k[n-p], \tag{1}$$

where $b_{k,p}$ are the complex weights associated with the FIR filters. The signal is expressed as its equivalent discrete Fourier transform (DFT)

$$\underline{s}_k[m] = \sum_{n=0}^{N-1} \sum_{p=0}^{P} b_{k,p} c_k[n-p] e^{-j2\pi \frac{nm}{N}} = \mathbf{f}^T[m]\mathbf{C}_k \mathbf{b}_k. \tag{2}$$

Here, $m = 0 \dots N-1$, $\mathbf{f}[m] = [1 \dots e^{-j2\pi \frac{(N-1)m}{N}}]^T$, $\mathbf{b}_k = [b_{k,0} \dots b_{k,P}]^T$ and $\mathbf{C}_k = \begin{bmatrix} c_k[0] & \dots & c_k[-P] \\ \vdots & \ddots & \vdots \\ c_k[N-1] & \dots & c_k[N-1-P] \end{bmatrix}$ is a cyclically shifted signal matrix.

Further, if we assume true time delay (TTD) technology in the transmitters and receivers, the signals will not be distorted by the wide signal bandwidth. At an angle α in a 2D space (ignoring the effect of the time delay), the signal is

$$\underline{r}_\alpha[m] = \sum_{k=1}^{K} \underline{\mathbf{a}}_T^T(\alpha, \frac{m}{NT_s}) \underline{\mathbf{w}}_{T,k}^C(\theta_T, \frac{m}{NT_s}) \underline{s}_k[m]. \tag{3}$$

In (3), $\underline{\mathbf{a}}_T^T(\alpha, \frac{m}{NT_s})$ describes the physical antenna, where α is the angle of departure (AOD) at frequency index $\frac{m}{NT_s}$ and $\underline{\mathbf{w}}_{T,k}^C(\theta_T, \frac{m}{NT_s})$ is the beamforming vector for the k^{th} subarray, θ_T being the steering angle. The sum of the received signals is expressed as

$$\underline{y}[m] = \sum_{l=1}^{L} \left(\sum_{i=1}^{I} \gamma_i[m] \underline{\mathbf{w}}_{R,l}^H(\theta_R, \frac{m}{NT_s}) \underline{\mathbf{a}}_R(\beta_i, \frac{m}{NT_s}) \underline{r}_{\alpha_i}[m] + \underline{n}_l[m]) \right), \tag{4}$$

where θ_R is the steering angle for the l^{th} receiver antenna subarray, β_i is the angle of arrival (AOA) and $\gamma_i[m]$ the corresponding complex radar cross section (RCS), including path loss. The receiver noise at the l^{th} subarray is denoted $\underline{n}_l[m]$. The received signal, $\underline{\mathbf{y}}$, is passed through a temporal receive filter $\underline{\mathbf{h}}$ giving the output signal $\mathbf{u}$, which in the time domain is

$$u[n] = y[n] * h[n] = \sum_{i=0}^{N-1} h^c[n-i] y[i]. \tag{5}$$

The peak of $u[n]$ is obtained when $n = N-1$, i.e. $u[N-1] = \sum_{i=0}^{N-1} h^c[N-1-i]y[i] = \tilde{\mathbf{h}}^H \mathbf{y}$, where $\tilde{\mathbf{h}} = [h[N-1] \dots h[0]]^T$ and $\mathbf{y} = [y[0] \dots y[N-1]]^T$. Define a matrix $\mathbf{F}$ that consists of the DFT coefficients and note that $\mathbf{F}^H\mathbf{F} = N\mathbf{I}$, which results in

$$u[N-1] = \tilde{\mathbf{h}}^H \frac{\mathbf{F}^H\mathbf{F}}{N}\mathbf{y} = \frac{1}{N}\tilde{\mathbf{h}}^H\mathbf{F}^H\mathcal{F}(\mathbf{y}) = \frac{\tilde{\mathbf{h}}^H\mathbf{F}^H\underline{\mathbf{y}}}{N}, \tag{6}$$

where $\underline{\mathbf{y}}$ is the DFT of $\mathbf{y}$ and is given in (4). We seek to maximize the SNIR at the receiver output, defined as

$$\text{SNIR} = \frac{E[|\tilde{\mathbf{h}}^H\mathbf{F}^H\underline{\mathbf{y}}_s|^2]}{E[|\tilde{\mathbf{h}}^H\mathbf{F}^H\underline{\mathbf{y}}_i|^2] + E[|\tilde{\mathbf{h}}^H\mathbf{F}^H\underline{\mathbf{n}}|^2]}. \tag{7}$$

Here, $\underline{\mathbf{y}}_s$ and $\underline{\mathbf{y}}_i$ are the received signal components from the targets and the interference, respectively, and $\underline{\mathbf{n}}$ is the total amount of receiver noise.

III. Optimization Procedure

In this section, the transmit and receive filters that yield optimal system performance are derived. We assume the target and interference signals to be deterministic (i.e. fixed), whereas the noise is random. This gives the following maximization problem

$$\max_{\mathbf{b},\tilde{\mathbf{h}}} \quad \text{SNIR}(\mathbf{b},\tilde{\mathbf{h}}) = \frac{|\tilde{\mathbf{h}}^H\underline{\mathbf{X}}_s\mathbf{b}|^2}{|\tilde{\mathbf{h}}^H\underline{\mathbf{X}}_i\mathbf{b}|^2 + E[|\tilde{\mathbf{h}}^H\underline{\mathbf{n}}_F|^2]} \quad \text{subject to} \quad ||\mathbf{b}_k||^2 \le P_{T,k}^{\max}. \tag{8}$$

In (8), $\mathbf{b} = [\mathbf{b}_1^T \dots \mathbf{b}_K^T]^T$, $\underline{\mathbf{n}}_F = \mathbf{F}^H\underline{\mathbf{n}}$, $\mathbf{F}^H\underline{\mathbf{y}}_s = \underline{\mathbf{X}}_s\mathbf{b}$, $\mathbf{F}^H\underline{\mathbf{y}}_i = \underline{\mathbf{X}}_i\mathbf{b}$ and $P_{T,k}^{\max}$ is the power constraint associated with the transmit filters. Further, $\underline{\mathbf{X}}_s = [\underline{\mathbf{x}}_s[0] \dots \underline{\mathbf{x}}_s[N-1]]^T$ and $\underline{\mathbf{x}}_s[m] = [g_{s,1}(m)\underline{c}_1[m] \dots g_{s,1}(m)\underline{c}_1[m-P]\; g_{s,2}(m)\underline{c}_2[m] \dots g_{s,K}(m)\underline{c}_k[m-P]]^T$, where $g_{s,k}(m) = \gamma_s[m]\underline{\mathbf{w}}_R^H(\theta_R, \frac{m}{NT_s})\underline{\mathbf{a}}_R(\beta_s, \frac{m}{NT_s})\underline{\mathbf{a}}_{T,k}^T(\alpha_s, \frac{m}{NT_s})\underline{\mathbf{w}}_{T,k}^C(\theta_T, \frac{m}{NT_s})$. The matrix $\underline{\mathbf{X}}_i$ is derived in a similar way.

Continuing, we discuss two optimization methods: one alternating approach (see Section III-A) and one joint transmit-receive approach (see Section III-B). This gives the possibility to compare the performance of both methods. The optimal transmit filters are derived for two cases: a total power constraint and for an individual power constraint on each subarray. To the respectively best knowledge of the authors the latter case has not been studied for MIMO radar. Nonetheless, it is of great importance as large antenna arrays typically are divided into several subarrays with a power amplifier at each element. Therefore, a power constraint associated with each subarray is equivalent to (as the same waveform is used for each element in the subarray) a power constraint one each element.

A. Alternating Transmit–Receive Filter Optimization

Herein, the transmit and receive filter coefficients are optimized separately, i.e. to solve the maximization problem (8), we first find the optimal $\tilde{\mathbf{h}}$ while keeping $\mathbf{b}$ fixed, and second we find the optimal $\mathbf{b}$ for a fixed $\tilde{\mathbf{h}}$. The problem is evaluated for two power constraint formulations (see Section III-A1 and III-A2). To initialize, the weight coefficients are set to be the optimal weights for the case where no interference is present (the matched filter).

1) Optimization Using a Total Power Constraint: The alternating method is similar to the one described in [9] and [8]. However, our formulation is a combination of both methods and evaluated for a different application. First, (8) is maximized with respect to $\tilde{\mathbf{h}}$

$$\max_{\tilde{\mathbf{h}}} \quad \text{SNIR}(\tilde{\mathbf{h}}) = \frac{\tilde{\mathbf{h}}^H\underline{\mathbf{X}}_s\mathbf{b}\mathbf{b}^H\underline{\mathbf{X}}_s^H\tilde{\mathbf{h}}}{\tilde{\mathbf{h}}^H(\underline{\mathbf{X}}_i\mathbf{b}\mathbf{b}^H\underline{\mathbf{X}}_i^H + \underline{\mathbf{R}}_n)\tilde{\mathbf{h}}}. \tag{9}$$

Here, $\underline{\mathbf{R}}_n = E[\underline{\mathbf{n}}_F \underline{\mathbf{n}}_F^H]$ is the covariance matrix of the receiver noise. By using the Cauchy-Schwarz inequality with vectors $(\underline{\mathbf{R}}_{j,n}^{1/2}\tilde{\mathbf{h}})$ and $(\underline{\mathbf{R}}_{j,n}^{-1/2}\underline{\mathbf{X}}_s\mathbf{b})$, $(\underline{\mathbf{R}}_{j,n} = \underline{\mathbf{X}}_i\mathbf{b}\mathbf{b}^H\underline{\mathbf{X}}_i^H + \underline{\mathbf{R}}_n)$, the upper bound is obtained when $\tilde{\mathbf{h}}_{\text{opt}} = \underline{\mathbf{R}}_{j,n}^{-1}\underline{\mathbf{X}}_s\mathbf{b}$. In the second step, (8) is maximized with respect to $\mathbf{b}$. The optimization problem is reformulated as (see [8] for details)

$$\max_{\hat{\mathbf{b}}} \quad \text{SNIR}(\hat{\mathbf{b}}) = \frac{P_T^{\max}\hat{\mathbf{b}}^H\underline{\mathbf{T}}\hat{\mathbf{b}}}{\hat{\mathbf{b}}^H(P_T^{\max}\underline{\mathbf{Q}} + \sigma^2_{n,\tilde{\mathbf{h}}}\mathbf{I})\hat{\mathbf{b}}} \qquad (10)$$
$$\text{subject to} \quad ||\hat{\mathbf{b}}||^2 = 1,$$

where $\hat{\mathbf{b}} = \mathbf{b}/\sqrt{P_T^{\max}}$, $\underline{\mathbf{T}} = \underline{\mathbf{X}}_s^H\tilde{\mathbf{h}}\tilde{\mathbf{h}}^H\underline{\mathbf{X}}_s$, $\underline{\mathbf{Q}} = \underline{\mathbf{X}}_i^H\tilde{\mathbf{h}}\tilde{\mathbf{h}}^H\underline{\mathbf{X}}_i$ and $\sigma^2_{n,\tilde{\mathbf{h}}} = \tilde{\mathbf{h}}^H\underline{\mathbf{R}}_n\tilde{\mathbf{h}}$. The solution is $\hat{\mathbf{b}}_{\text{opt}} = P_T^{\max}(P_T^{\max}\underline{\mathbf{Q}} + \sigma^2_{z,\tilde{\mathbf{h}}}\mathbf{I})^{-1}\underline{\mathbf{X}}_s^H\tilde{\mathbf{h}}$, normalized to satisfy the unit norm constraint and finally $\mathbf{b}_{\text{opt}} = \sqrt{P_T^{\max}}\hat{\mathbf{b}}_{\text{opt}}$.

2) Optimization Using a Power Constraint for each Subarray: In this section, we describe the alternating optimization when each subarray has an individual power constraint. The receive filter is independent of the power constraint. Thus, it is solved as in the previous section. However, the second step is

$$\max_{\mathbf{b}} \quad \text{SNIR}(\mathbf{b}) = \frac{|\mathbf{b}^H\underline{\mathbf{X}}_s^H\tilde{\mathbf{h}}|^2}{|\mathbf{b}^H\underline{\mathbf{X}}_i^H\tilde{\mathbf{h}}|^2 + \tilde{\mathbf{h}}^H\underline{\mathbf{R}}_n\tilde{\mathbf{h}}} \qquad (11)$$
$$\text{subject to} \quad ||\mathbf{b}_k||^2 \le P_{T,k}^{\max},$$

where $P_{T,k}^{\max}$ is the maximum allowable transmit power of the k^{th} subarray. Exploiting the properties of the trace operator denoted as $\text{Tr}(\cdot)$, and by introducing a slack variable t gives the reformulation by means of semidefinite relaxation [8]

$$\max_{\mathbf{B},t} \quad t \qquad (12)$$
$$\text{subject to} \quad \text{Tr}(\mathbf{B}(\underline{\mathbf{T}} - t\underline{\mathbf{Q}})) \ge \sigma^2_{n,\tilde{\mathbf{h}}}t \qquad (13)$$
$$\mathbf{B}_{kk} \le P_{T,k}^{\max}, \mathbf{B} \succeq 0.$$

Here, $\mathbf{B} = \mathbf{b}\mathbf{b}^H$. The optimization problem is quasi convex and can be solved using a bisection technique [10]. Thus, a convex feasibility problem is solved in each step for a fixed value of $t \in [t_l \quad t_u]$, t_l and t_u being the lower and upper bound for the feasibility problem. The variable, t, is increased when there exist a feasible solution and decreased when the solution is infeasible. Note that in semidefinite relaxation the rank constraint on the matrix $\mathbf{B}$ is ignored. However, this is not always fulfilled. In the sequel we assume that $\mathbf{B}$ has rank–1 if 99% of the energy is contained in the largest eigenvalue. Thus, if $\text{rank}(\mathbf{B}) \neq 1$, a randomization technique [11] is used to find a valid solution.

B. Joint Transmit–Receive Filter Optimization

Herein we discuss a joint optimization approach, where the receive filter is directly given by the transmit filters. Thus, we optimize the transmit filters while keeping the receive filter matched to the transmitted signals in the mainbeam direction. For fixed transmit filters the upper bound of the SNIR is obtained when $\tilde{\mathbf{h}}_{\text{opt}} = \underline{\mathbf{R}}_{j,n}^{-1}\underline{\mathbf{X}}_s\mathbf{b}$. Consequently, we write the maximization problem as

$$\max_{\mathbf{b}} \quad \text{SNIR}(\mathbf{b}) = \mathbf{b}^H\underline{\mathbf{X}}_s^H\mathbf{R}_{j,n}^{-1}\underline{\mathbf{X}}_s\mathbf{b}$$
$$\text{subject to} \quad ||\mathbf{b}_k||^2 \le P_{T,k}^{\max}. \qquad (14)$$

Using the matrix inversion lemma yields the more computationally efficient form

$$\text{SNIR}(\mathbf{b}) = \mathbf{b}^H\hat{\underline{\mathbf{X}}}_s^H\hat{\underline{\mathbf{X}}}_s\mathbf{b} - \frac{\mathbf{b}^H\hat{\underline{\mathbf{X}}}_s^H\hat{\underline{\mathbf{X}}}_i\mathbf{b}\mathbf{b}^H\hat{\underline{\mathbf{X}}}_i^H\hat{\underline{\mathbf{X}}}_s\mathbf{b}}{1 + \mathbf{b}^H\hat{\underline{\mathbf{X}}}_i^H\hat{\underline{\mathbf{X}}}_i\mathbf{b}}, \qquad (15)$$

In (15), $\hat{\underline{\mathbf{X}}}_s = \mathbf{R}_{j,n}^{-1/2}\underline{\mathbf{X}}_s$ and $\hat{\underline{\mathbf{X}}}_i = \mathbf{R}_{j,n}^{-1/2}\underline{\mathbf{X}}_i$ which results in the following optimization problem

$$\max_{\mathbf{b}} \quad \text{SNIR}(\mathbf{b})$$
$$\text{subject to} \quad ||\mathbf{b}_k||^2 \le P_{T,k}^{\max}, \qquad (16)$$

where $P_{T,k}^{\max}$ is a total or divided power constraint for the weight coefficients. The optimization problem consist of a non–convex nonlinear objective function with constraints that are nonlinear, i.e. for the total and divided power constraint. A full grid search is of exponential complexity. We have therefore used a local optimization technique, initialized in a similar way as the alternating optimization. However, there is no guarantee that the solution will converge to a global optimum. Nonetheless, through experimental validation it is seen that the cost function has several local maxima each with slightly different SNIR, implying that the algorithm is insensitive to the initialization vector. Thus, even by reaching a local maximum we achieve the maximization of the SNIR.

IV. Numerical results

In this section, we evaluate the performance of the proposed algorithms. The following settings are used: a monostatic radar, ($\alpha = \beta$), equipped with a uniform linear array (ULA) with $K = 15$ and $L = 15$ elements in the transmitter and receiver, respectively, divided equally into 3 subarrays. The carrier frequency is $f_c = 9$ GHz with a bandwidth of $B = 1$ GHz. This corresponds to a relative bandwidth of 11%. The inter-spacing between the elements is $d = \frac{c}{2(f_c+B/2)}$. The time domain samples $c_k[n]$, $n = 1 \dots 256$, are generated as complex white Gaussian noise with variance 1. We investigate the case where a target with a radar cross section (RCS), pathloss and effective receive subarray area of $\gamma_s[m] = \gamma_s = 20\log_{10}(\frac{P_s}{P^{\text{tot}}})$ dB, ($P_s = 10^{-2}$ and P^{tot} is the total maximum transmit power) is situated at an angle $\alpha_s = 30°$. The interference consists of a repetition jammer, which retransmits the signal with an amplified power, here $P_j = 10^2$ at $\alpha_j = -30°$. The antenna array is steered towards the target of interest, i.e. $\theta_T = \theta_R = \alpha_s$. The receiver noise is generated as complex white Gaussian noise with a 10 times higher power than the total maximum transmit power. The results are averaged over 200 Monte Carlo trials.

The SNIR is evaluated for a total power of $P_T^{\max} = 1$ W, and for a divided power of $P_{T,k}^{\max} = \frac{1}{3}$ W. In Fig. 2, the comparison of the SNIR for different number of FIR filter orders is illustrated. As seen, the joint and the alternating optimization methods follow each other and are bounded by the matched filter bound, i.e. when no interference exist. As expected, the SNIR is reduced when the power cannot freely be distributed between the transmit filters.

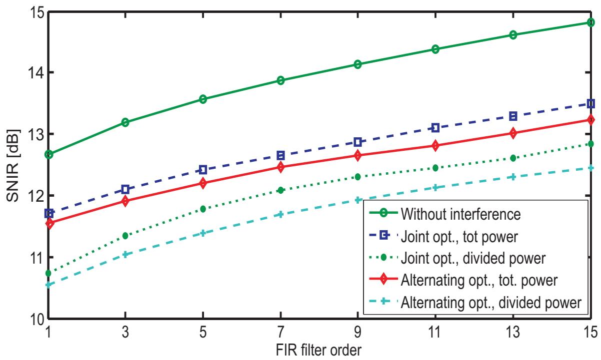

Fig. 2. Comparison of the maximum SNIR for both optimization method evaluated for different FIR filter orders.

Continuing, we expect that the alternating and the joint optimization converge to the same value. Hence, by increasing the number of iterations and forcing the alternation to continue, the difference reduces from 3% after 10 iterations to 0.03% after 20.000 iterations for the total power constraint. In comparison, for the divided power constraint it is not obvious that the algorithms converge to the same SNIR due to the semidefinte relaxation, where the solution may not satisfy the rank–1 condition.

To evaluate the solution of the joint optimization; 100 randomly initialized weight coefficients are used as start values for the optimization, and each trial achieves the same SNIR with a maximum difference in the fourth decimal.

An initial study of the expected loss in performance when the prior knowledge of the angle towards the target and jammer differ from the actual position is shown in Table I. The results are evaluated for a mismatch error of $\sigma_e^2 = [0.5, 1, 2, 5]$ degrees, $P = 3$; and averaged over 200 Monte Carlo runs for each SNIR maximization trial.

TABLE I
PERFORMANCE DEGRADATION DUE TO POINTING ERROR

	$\sigma_e^2 = 0.5$	$\sigma_e^2 = 1$	$\sigma_e^2 = 2$	$\sigma_e^2 = 5$
E[SNIR], α_s error	11.3	11.2	11.1	10.5
E[SNIR], α_j error	11.3	11.1	10.4	7.6

In Fig. 3 the normalized wideband antenna array beampattern is illustrated. We define it as the sum over the frequencies at an angle α, i.e. $\text{AP}[\alpha] = \sum_{m=0}^{N-1} |\underline{r}_\alpha[m]|^2$, where α is divided into $\alpha = [-90 \ldots 90)$ grid points corresponding to the angles of interest. As illustrated, there is no cancellation in the direction towards the jammer. Thus, the jammer is suppressed using the temporal properties of the waveforms. However, the power directed against the target is increased, though the power enhancement is not large enough to cancel the effect of the jammer.

V. CONCLUDING REMARKS

In this paper, we have discussed the possibility to suppress interference using the temporal properties of the transmitted signals. By using tunable filters, two algorithms that maximize the SNIR for a wideband MIMO radar system are formulated. Numerical validation show that it is possible to suppress the interference in the temporal domain instead of in the spatial domain. The algorithms are formulated for two different power constraints on the transmit filters: a total and an individual power constraint. Worth noting is that this investigation is a first step to reach a fully adaptive system, where the waveforms are continuously optimized to a time-variant scenario. A drawback of the proposed approaches is that prior knowledge of the environment has to be known.

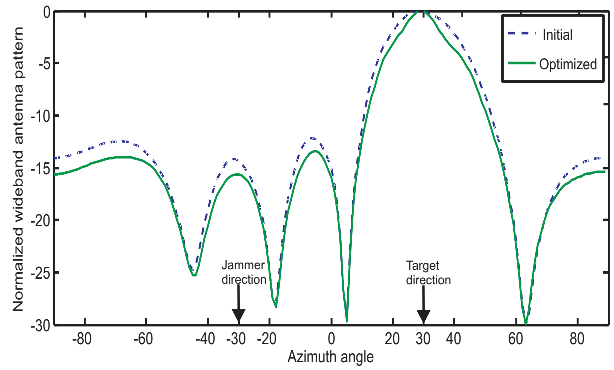

Fig. 3. Wideband antenna array beampattern before and after optimization, for individual power constraint on each subarray.

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