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Vector Modulator-Based Analog Beamforming Using the Least Euclidean Distance Criterion

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ABSTRACT In traditional multiple-input multiple-output (MIMO) receivers, radio frequency (RF) front-ends are exposed to interference as no analog spatial filtering is employed before the digital beamforming stage. Therefore the RF front-end is power-hungry, and analog to digital converters require a high dynamic range. In this paper, we consider an analog beamforming system in case of narrowband signals to cancel interference early in the analog domain, thus reducing the required ADC resolution. In contrast to existing analog beamformers with only phase shifts, our proposed design employs vector modulators where the coefficients can be selected from a set of weights with variable phases and amplitudes. We also propose an efficient and fast Euclidean distance algorithm to determine the analog beamformer coefficients while being suitable for realistic scenarios. Finally, an expression is introduced to estimate the interference rejection achieved by employing the proposed algorithm and a vector modulator in the RF domain. The introduced algorithm leads to considerable improvement in computational complexity by slightly sacrificing interference rejection.

INDEX TERMS Hybrid beamforming, interference mitigation, least Euclidean distance, RF phase shifter and vector modulator.

I. INTRODUCTION

Multiple-input multiple-output (MIMO) communication systems utilize multiple antennas at both transmitter and receiver to obtain selection diversity and improve spectral efficiency [1]. However, the existence of co-channel interference can limit the achievable diversity gains in MIMO communication systems. Although digital beamforming could mitigate the interference by applying spatial filtering in the digital domain, the radio frequency (RF) front-ends and analog to digital converters (ADCs) are still exposed to strong interference. Hence, the ADCs are compelled to allocate a considerable amount of their dynamic range to digitize undesirable signals [2].

Hybrid beamforming systems have been introduced to address the problem of high dynamic range requirements [2]–[4]. Introducing analog beamformers in the RF domain reduces interference, which leads to a less required dynamic range of the ADCs. In [5], analog beamforming was implemented through phase shifts at each antenna with a limited set of phase shift settings. A precomputed set of beamforming vectors was introduced in [6] to steer the beam in a specific direction to capture the one with the highest energy. The mentioned papers focused on generating a single data stream. Further improvement in data rate could be achieved by considering multiple data streams. A quantized matching pursuit algorithm was introduced in [7] to design an analog beamformer by minimizing the mean square error between a pilot signal and the system output at baseband. [8] employed Kronecker decomposition to decompose the analog beamforming vectors into some unit-modulus vectors to null the interference and combine user signals coherently. A zero-forcing-based precoder was introduced in [9] to capture the largest power gain while adding low complexity to the design. Other hybrid beamforming systems based on the minimum mean square error (MMSE) criterion can be found in [10]–[12]. All mentioned references have focused on controlling phases in designing analog beamformers without considering amplitude variations. Besides, they considered a low resolution for the analog coefficients in their design.

As opposed to analog beamformers with only phase shifts presented in many papers at the system level, circuit-based

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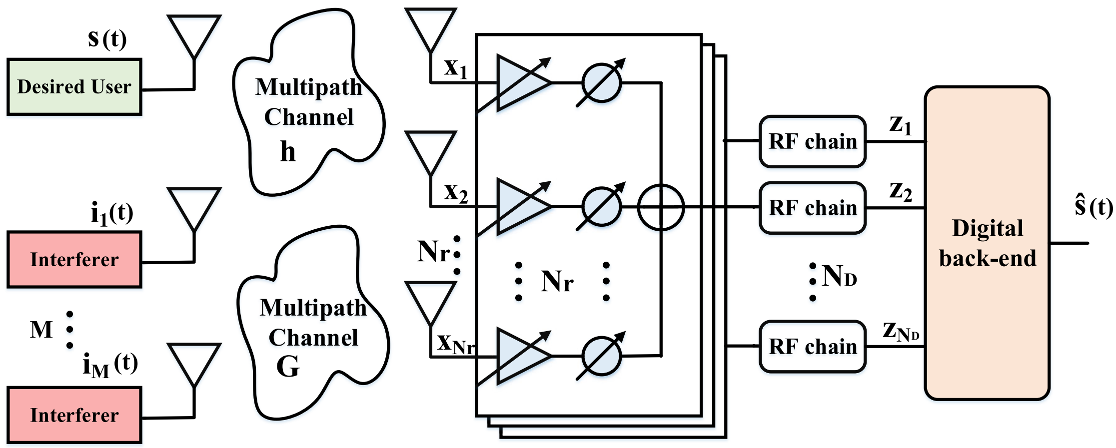


FIGURE 1. A single-user model in MIMO communication. The receiver contains an antenna array of N_r antennas which are exposed to the signal of the desired user and to interference generated by M users in adjacent cells.

papers describe the design of analog beamformers by setting phase and amplitude at individual antennas signals by a vector modulator [13]–[17]. Although it makes the design more complex, it is considered as an advancement over phase-shift analog beamforming by providing more interference rejection in the RF domain. On the other hand, even though vector modulators provide more resolution for the analog coefficients, literature usually concentrates on the vector modulators' design rather than addressing the question of how the coefficients should be determined. [18]–[20] addressed the way of determining the analog beamformer coefficients. However, these papers have considered only simple line-of-sight (LOS) scenarios, which seems to be over-simplistic in a realistic wireless scenario.

This paper addresses the existing literature gap, developing a fast and effective algorithm to cancel interference by controlling both phase and amplitude while being suitable for LOS and non-line-of-sight (NLOS) scenarios. The proposed algorithm works based on minimizing the Euclidean distance between the complex weight factors of a hybrid beamformer and a fully-digital beamformer. We investigate the effectiveness of utilizing vector modulators rather than phase shifters in terms of interference rejection. To do so, we exploit the vector modulator presented in [16], and develop a new analog beamformer equipped with weights that change phase and amplitude. The obtained results are compared with the optimal results presented in [7]. At the cost of a few dB less interference suppression, we drastically reduce the complexity of the algorithm and produce an appropriate solution for realistic wireless scenarios. Finally, we introduce an expression to quantify the amount of interference mitigation using the proposed algorithm.

This paper is organized as follows. Section II gives a brief overview of the system model. The methods for determining the weights for an analog beamformer are explained in section III. In the fourth section, we present an expression to approximate the interference rejection. Section V covers the results, and finally, the conclusion is provided in section VI.

The following notations have been used in this paper. $\mathbf{A}$ is a matrix; $\mathbf{a}$ represents a vector and a stands for a scalar. $(.)^T$, $(.)^H$ and $(.)^{-1}$ represent the transpose, hermitian and inverse operators, respectively. $\|A\|$ is the norm of A, and finally, the expectation is denoted by E.

II. SYSTEM MODEL

A. RECEIVED DATA MODEL

Consider a single-user model in an uplink MIMO communication system, as shown in Fig. 1. The receiver is equipped with an antenna array of N_r antennas and receives data from a single-antenna desired user. The receiver is also exposed to severe interference contributed by $M(M < N_r)$ users (interferer 1 to M) in adjacent cells. The baseband equivalent representation of the received signal can be modeled in discrete time by:

$$\mathbf{x}[k] = \mathbf{h}s[k] + \mathbf{G}\mathbf{i}[k] + \mathbf{n}[k] \quad (1)$$

where the vector $\mathbf{h} = [h_1 \ h_2 \ldots h_{N_r}]^T$ represents the data channel while $\mathbf{G} = [\mathbf{g}_1 \mathbf{g}_2 \ldots \mathbf{g}_M]^T$ is an $N_r \times M$ matrix of the interference channel response where $\mathbf{g}_i = [g_{i1}, g_{i2} \ldots g_{iN_r}]^T$ describes the channel between the i^{th} interferer and the receiver. $s[k]$ and $\mathbf{i}[k]$ show the desired user signal and interference signals, respectively. Finally, $\mathbf{n}[k]$ is an $N_r \times 1$ vector and represents additive white Gaussian noise (AWGN) with zero mean and unknown covariance matrix.

B. FULLY-DIGITAL BEAMFORMING

By considering the minimum mean-square error (MMSE) criterion and linear beamforming, the fully-digital beamformer (3) is the solution of (2) [1]:

$$\mathbf{w_{opt}} = \arg\min_{\mathbf{w_{opt}}} \ E\|s[k] - \mathbf{w_{opt}}^H \mathbf{x}[k]\|^2 \quad (2)$$

$$\mathbf{w_{opt}} = \mathbf{R_x}^{-1} \mathbf{r_{xs}} \quad (3)$$

where $\mathbf{R_x} = E\{\mathbf{x}[k]\mathbf{x}^H[k]\}$ and $\mathbf{r_{xs}} = E\{\mathbf{x}[k]s^H[k]\}$ are the covariance matrix and the cross-correlation vector, respectively. The estimation of these values needs accessibility to all antenna signals and the availability of a reference signal (training sequence) for the desired user. This paper assumes that the channel is perfectly estimated, and the covariance matrix and the cross-correlation vector are well estimated during the training phase. Furthermore, the effects of quantization by the ADCs are ignored. The fully-digital beamformer, also interchangeably called the ideal beamformer, is considered as a reference design in this paper.

C. ANALOG BEAMFORMING

As previously stated, we aim to reject interference before it reaches the ADCs, and it is directly translated into reducing the dynamic range and, therefore, power consumption. To do so, we put an analog beamformer after the low-noise amplifiers in the RF domain to cancel interference. The antenna signals are mapped to N_D output chain(s) by linearly combining the antenna signals, and are stacked into an $N_D \times 1$ vector. The discretized output of the analog beamformer after downconversion or baseband signal can be represented as follows:

$$\mathbf{z}[k] = \mathbf{W}^H \mathbf{x}[k] \tag{4}$$

where $\mathbf{W}$ is an $N_r \times N_D$ matrix and N_D denotes the number of analog beamformer outputs. Each entry of $\mathbf{W}$ represents the phase shift and/or amplitude variation applied by analog beamforming. Then by using the digital beamformer $\boldsymbol{\vartheta} = [\vartheta_1, \vartheta_2, \ldots, \vartheta_{N_D}]^T$ the final output is:

$$\mathbf{y}[k] = \boldsymbol{\vartheta}^H \mathbf{z}[k]. \tag{5}$$

By using the MMSE estimator, the final digital beamformer is obtained as follows:

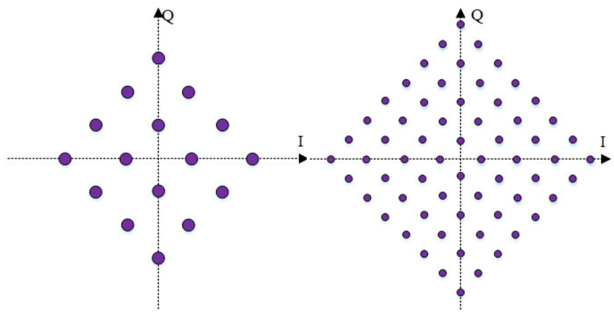
$$\begin{aligned} \boldsymbol{\vartheta}_0 &= \arg\min_{\boldsymbol{\vartheta}} \; E||s[k] - \boldsymbol{\vartheta}^H \mathbf{z}[k]||^2 \\ \boldsymbol{\vartheta}_0 &= \mathbf{R}_{\mathbf{z}}^{-1} \mathbf{r}_{\mathbf{zs}} \end{aligned} \tag{6}$$

where $\mathbf{R}_{\mathbf{z}} = E\{\mathbf{z}[k]\mathbf{z}^H[k]\}$ and $\mathbf{r}_{\mathbf{zs}} = E\{\mathbf{z}[k]s^H[k]\}$. Therefore, for a given $\boldsymbol{W}$ matrix, calculating the digital beamformer is straightforward, and hence, the key step is designing the analog beamformer, which is explained in the next section.

As we mentioned briefly in the introduction, only a limited number of analog beamformer weights are available. The weights are selected from a limited set of complex values, which is called the alphabet. An alphabet could consist of either only phase shifts or variable phases and amplitudes. While using only phase shifts is easier to implement, changing amplitude provides more degrees of freedom in suppressing interference. A vector modulator gives us the ability to control the phase and amplitude. We use the vector modulator presented in [16] to build different alphabets. Fig. 2 shows a 4×4 and an 8×8 vector modulator.

III. ANALOG BEAMFORMING DESIGN

In this section, we explain the calculation of the analog beamforming settings. The proposed algorithm based on the least Euclidean distance is described. This section also briefly explains the algorithm in [7] as we use it as a benchmark in the simulations.



(a) 4×4 vector modulator (b) 8×8 vector modulator

FIGURE 2. Vector modulator with two different sizes.

A. MMSE-BASED EXHAUSTIVE SEARCH ANALOG BEAMFORMING

The introduced algorithm in [7] aims to design a hybrid beamformer that shows the same performance as the fully-digital beamformer. In other words, the product of the analog beamformer $\boldsymbol{W}$ and the digital beamformer $\boldsymbol{\vartheta}$ equals the fully-digital beamformer $\mathbf{w}_{\mathbf{opt}}$:

$$\mathbf{W}\boldsymbol{\vartheta} = \mathbf{R}_{\mathbf{x}}{}^{-1}\mathbf{r}_{\mathbf{xs}} \Leftrightarrow \underline{\mathrm{W}}\boldsymbol{\vartheta} = \underline{\mathrm{r}}_{\mathbf{xs}} \tag{7}$$

where, $\underline{\mathrm{W}} = \mathbf{R}_{\mathbf{x}}{}^{1/2}\mathbf{W}$ and $\underline{\mathrm{r}}_{\mathbf{xs}} = \mathbf{R}_{\mathbf{x}}{}^{-1/2}\mathbf{r}_{\mathbf{xs}}$. Based on (7), $\underline{\mathrm{r}}_{\mathbf{xs}}$ needs to be in the subspace spanned by the columns of $\underline{\mathrm{W}}$, which would require total freedom in choosing the weights for the analog beamformer. As the analog beamformer only allows a limited choice of weights, the columns of $\underline{\mathrm{W}}$ are selected in a way that they span a subspace that is close to $\underline{\mathrm{r}}_{\mathbf{xs}}$. As [7] explains, each column of $\underline{\mathrm{W}}$ is calculated by:

$$\underline{\mathrm{w}}_i = \underset{\underline{\mathrm{w}}_i \in \underline{\mathrm{D}}}{\mathrm{argmax}} \frac{|\underline{\mathrm{w}}_i^H \, \underline{\mathrm{r}}_{\mathbf{xs}}|}{\|\underline{\mathrm{w}}_i\|} \tag{8}$$

where $\underline{\mathrm{D}} = \mathbf{R}_{\mathbf{x}}{}^{1/2}D$ and D represents the set of all possible vectors that each column of $\mathbf{W}$ can take. Based on equation (8), this algorithm searches exhaustively through the dictionary to find each column of $\underline{\mathrm{W}}$. Therefore, being time consuming is the disadvantage of using this algorithm.

B. ANALOG BEAMFORMING DESIGN USING LEAST EUCLIDEAN DISTANCE

In the previous sub-section, we explained the analog beamformer calculation based on the MMSE exhaustive search method. Although it is an optimal solution in suppressing interference, being time-consuming still is a bottleneck, as it looks for the best result out of a linear combination of all antennas. We target minimizing the Euclidean distance between the complex weight factors of the analog beamformer and the fully-digital beamformer, i.e., choosing the coefficient for each antenna individually. We also consider a vector instead of a matrix for the analog beamformer due to the existence of a single desired user. By considering $\mathbf{w}_{\mathbf{opt}}$

as the fully-digital beamformer and $\mathbf{w}$ as the selected analog beamforming vector, the problem is addressed by

$$w_i = \underset{w_i \in A}{\operatorname{argmin}} \left\| w_{opt_i} - w_i \right\| \tag{9}$$

where, $i = 1, 2, .., N_r$ and A (the alphabet), denotes the set of coefficients that each single antenna can take.

IV. ANALYSIS

A. COMPLEXITY ANALYSIS

This sub-section compares the computational complexity of the proposed Euclidean distance and exhaustive search [7] algorithms in terms of the number of iterations.

According to [7], the analog beamformer coefficients are set by searching exhaustively through the dictionary to find the maximum correlations between the $\underline{w}_i$ and $\underline{r}_{\mathbf{xs}}$. Nevertheless, the proposed Euclidean distance algorithm searches for an optimal local solution for each antenna individually. If N denotes the number of coefficients in each analog beamformer tap (alphabet size), the size of the dictionary D is N^{N_r}. To find $\mathbf{w}$ based on the least Euclidean distance, we need to search among N coefficients for N_r times while it takes N^{N_r} steps to search exhaustively through the dictionary. Therefore, the complexities of the Euclidean distance and exhaustive search algorithms are in the order of $\mathcal{O}(N \times N_r)$ and $\mathcal{O}(N^{N_r})$, respectively. For instance, Fig. 2a shows an alphabet with 16 different analog settings. In this case, the number of searches in the exhaustive search algorithm is $16^4 = 65536$, while only $16 \times 4 = 64$ computations are needed in the proposed Euclidean distance algorithm.

B. INTERFERENCE REJECTION

This section analyzes the effect of errors in the analog beamforming weights on the interference rejection performance of an array system. We derive an expression for characterizing the interference suppression in the presence of phase and amplitude errors in the antenna weights. The derivation has been done based on the proposed Euclidean distance algorithm.

Assume that $\mathbf{w_{opt}}$ presents the ideal choice for the analog beamforming weights, and $\mathbf{w}$ shows its deviated version that takes into account errors in phase and amplitude. By representing the weights in Cartesian coordinates and taking into account the errors in the in-phase (I) and quadrature (Q) components, we can define the complex vectors $\mathbf{w_{opt}}$ and $\mathbf{w}$ as:

$$\begin{aligned} \mathbf{w_{opt}} &= [w_1, w_2, .., w_{N_r}]^T \\ \mathbf{w} &= [(w_1 + e_{w_1}), (w_2 + e_{w_2}), \ldots, w_{N_r} + e_{w_{N_r}})]^T \end{aligned} \tag{10}$$

where $e_{w_k} = e_{w_{k_I}} + je_{w_{k_Q}}$ $1 \le k \le N_r$, $e_{w_{k_I}}$ and $e_{w_{k_Q}}$ are i.i.d real random variables. Let $\mathbf{H_I} = [\mathbf{h_1}, \mathbf{h_2}, .., \mathbf{h_M}]$ be the matrix where the columns describe the channel of the interfering signals. It is important to mention that rejecting interference is possible when $M < N_r$. As $\mathbf{w_{opt}}$ is the ideal beamforming vector, it is orthogonal to the subspace created by a column span of $\mathbf{H_I}$. Adding errors in phase and amplitude leads to deviation from $\mathbf{w_{opt}}$ which results in a reduction of the desired signal power and an increase in the interference strength. A slight deviation implies roughly no power reduction of the desired signal but, low-power interference is introduced.

Below, we will show that the average interference rejection over different channel realization ($\overline{IR}$) is estimated as:

$$\overline{IR} \approx E\left[\frac{N_r^2}{\left\|\mathbf{w_{opt}} - \mathbf{w}\right\|^2}\right] = \frac{3(R-1)^2}{2} \tag{11}$$

where R is the number of points of the squared vector modulator on the horizontal line through the origin. Interference Rejection (IR) is defined as the ratio of the signal to interference power after applying the analog beamforming weights (SIR_{out}) to the signal to interference power before using the analog beamformer (SIR_{in}). SIR_{in} is also the ratio of the desired user power to the sum of all interference power, received at antenna one. It is the same for all antennas.

Derivation: we consider $N_D = 1$ and one interfering signal (channel $\mathbf{h}$ corresponding to the interfering signal) with unit power. The optimal analog beamformer is normalized to be completely inside the squared alphabet. Therefore, for each antenna, the amplitude of the selected point in the alphabet can be larger or smaller than the amplitude of the normalized optimal beamformer. After applying the analog beamformer, $\mathbf{w}$, $\overline{IR}$ is defined as:

$$\begin{aligned} \overline{IR} &= E\left[\frac{SIR_{out}}{SIR_{in}}\right] = E\left[\frac{S_{out}^2}{S_{in}^2}\frac{I_{in}^2}{I_{out}^2}\right] \\ &= E\left[\frac{G_{s_{LE}}^2 S_{in}^2}{S_{in}^2}\frac{I_{in}^2}{G_I^2 I_{in}^2}\right] \approx N_r^2 E\left[\frac{1}{G_I^2}\right] \end{aligned} \tag{12}$$

where $G_{s_{LE}}$ and G_I are the array gains for the signal and interference, respectively. $G_{s_{LE}} = G_{s_{opt}} + \Delta G_s$, where $G_{s_{opt}}$ represents the array gain for the signal while using the ideal analog beamformer. $G_{s_{LE}}$ is approximately N_r as we can ignore the slight power variation of the signal after applying $\mathbf{w}$. G_I can also be represented as $G_I = G_{I_{opt}} + \Delta G_I$, where $G_{I_{opt}}$ is the interference array gain after applying the optimal beamformer, which is zero. Subject to the minor deviation between $\mathbf{w_{opt}}$ and $\mathbf{w}$, $(\mathbf{w_{opt}} - \mathbf{w})$ and $\mathbf{h}$ are almost co-directional and subsequently, the projection of $(\mathbf{w_{opt}} - \mathbf{w})$ onto $\mathbf{h}$ is approximately $(\mathbf{w_{opt}} - \mathbf{w})$. Therefore, the interference leakage power can be represented by $\left\|\mathbf{w_{opt}} - \mathbf{w}\right\|^2$.

We assume that the equidistant square vector modulator is circumscribed by a circle with radius 1 which is illustrated in Fig. 3. Hence, the distance between two adjacent constellation points is calculated as $d = \frac{2}{R-1}$ where R is the number of points of the squared vector modulator on the horizontal line through the origin. We mentioned earlier, $e_{w_{k_I}}$ and $e_{w_{k_Q}}$ $1 \le k \le N_r$ are i.i.d real random variables. Let's consider a uniform distribution for $e_{w_{k_I}}$ and $e_{w_{k_Q}}$. Based on the vector modulator, the absolute value of the errors in both the in-phase and quadrature parts vary from 0 to $\frac{d}{2}$.

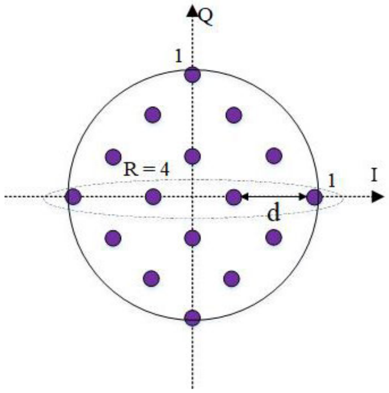


FIGURE 3. Distance between two adjacent settings in the in-phase and quadrature components.

Therefore, $E\left[e^2_{w_{k_I}}\right]$ and $E\left[e^2_{w_{k_Q}}\right]$ are calculated as follows:

$$E\left[e^2_{w_{k_I}}\right] = \int_0^{\frac{1}{R-1}} (R-1)x^2 dx = \frac{1}{3(R-1)^2} \tag{13}$$

$$E\left[e^2_{w_{k_Q}}\right] = \int_0^{\frac{1}{R-1}} (R-1)x^2 dx = \frac{1}{3(R-1)^2}, \tag{14}$$

and the average interference rejection over different channel realization ($\overline{IR}$) is:

$$\overline{IR} \approx \frac{N_r^2}{E\left[\left\|\mathbf{w}_{\mathbf{opt}} - \mathbf{w}\right\|^2\right]} = \frac{N_r^2}{N_r^2(E\left[e^2_{w_{k_I}}\right] + E\left[e^2_{w_{k_Q}}\right])}$$
$$= \frac{1}{2E\left[e^2_{w_{k_I}}\right]} = \frac{3(R-1)^2}{2}. \tag{15}$$

V. SIMULATION RESULTS

This section provides the simulation results to show the effect of employing a vector modulator on the interference suppression compared to phase shifters, to validate the proposed expression, and to compare the computational complexity of all algorithms discussed in the paper. The worst-case scenario, i.e., considering the maximum number of interferers, has been chosen to fully demonstrate the effectiveness of the proposed algorithm. The simulation parameters are set as follows unless specified otherwise. $N_R = 4$ receive antennas, three interferers ($M = 3$) with $SIR_{in} = -5$ dB, one desired user, $N_D = 1$ RF chain, and the uncoded quadrature phase-shift keying (QPSK) modulation for both desired user and interference signals. Finally, the results are obtained by averaging 1000 Monte Carlo runs, with independent Rayleigh channel realizations.

A. SINR CRITERION

Fig. 4 depicts the signal to interference and noise ratio (SINR) at the output of the analog beamformer as a function of the input signal to noise ratio (SNR). Input SNR is defined as the ratio of the desired user to noise power, received at antenna 1. The SINR is increasing for all three cases for increasing SNR as noise is dominant. However, at higher SNR, interference is becoming dominant, and hence, curves are saturating (except for the ideal beamformer) as they reach their limits in interference rejection. As the figure shows, employing vector modulators in the analog beamformer improves SINR by a factor higher than 3 dB (at SNRs higher than 15 dB) in comparison to the alphabet with only phase shifts. This improvement can be justified by mentioning that amplitude variations provide more degrees of freedom in interference rejection.

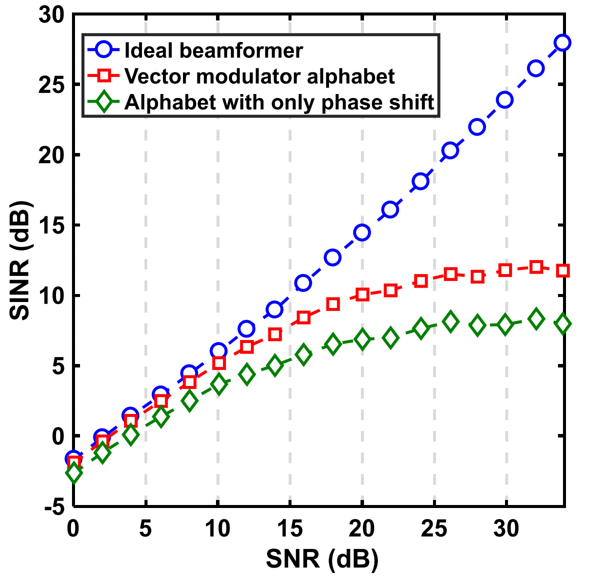


FIGURE 4. SINR comparison between the ideal beamformer, the vector modulator, and the phase shifter used in an analog beamformer where the alphabet size is 16.

B. COMPLEXITY COMPARISON

To show the advantage of using the least Euclidean distance algorithm over the exhaustive search in terms of computational complexity, we plot the number of searches versus different alphabet sizes in Fig. 5. The exhaustive search algorithm searches through the dictionary to find the optimum coefficients. As the dictionary grows exponentially with an increase in alphabet size, the search number follows the same trend. However, this trend is linear by using the proposed Euclidean distance algorithm.

C. INTERFERENCE REJECTION VERSUS DIFFERENT ALPHABET SIZE

This subsection verifies the proposed expression in (15) and shows the impact of different alphabet sizes on the interference rejection achieved by the proposed and the exhaustive search algorithms. To do so, we generated vector modulator alphabets with different sizes varying from 3×3 (alphabet size 9) to 12×12 (alphabet size 144) and simulated the interference rejection under the predefined setup parameters. Fig. 6 shows the interference rejection as a function of the alphabet size where the input SNR is set to 30 dB to rule out the effect of noise on the results. We display

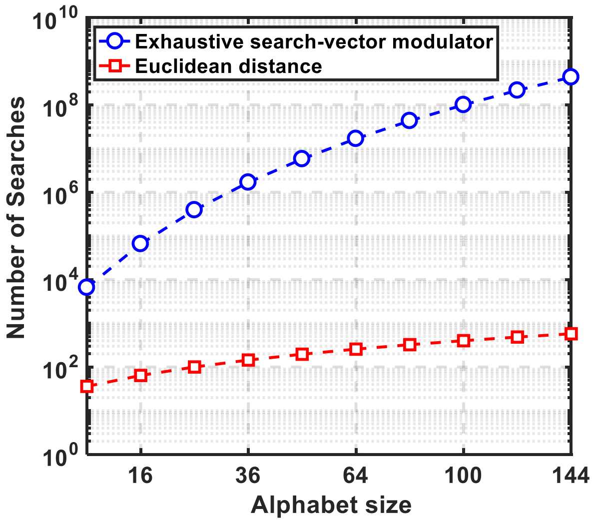


FIGURE 5. Complexity versus alphabet size for the Euclidean distance and exhaustive search algorithms.

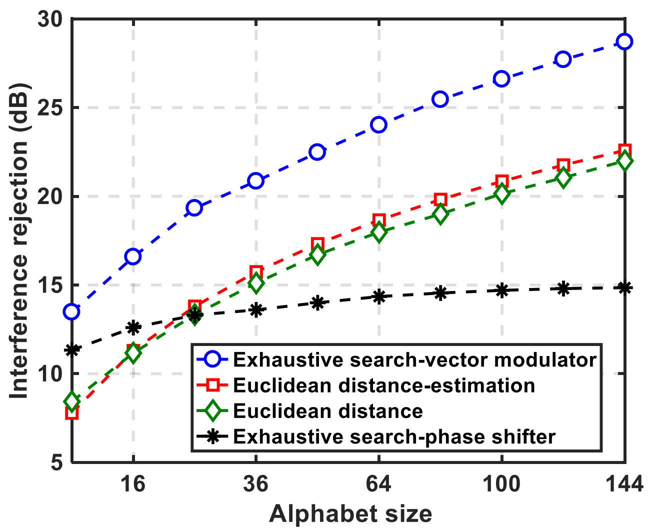


FIGURE 6. Interference rejection versus different sizes of vector modulators.

the estimated interference rejection based on (15) and the simulation results. As it can be clearly seen in Fig. 6, there is a gap between the performance of the exhaustive search phase-shifter beamformer and the other beamformers, which are all using a vector modulator, for alphabet sizes larger than 25. This performance gap is widening by an increase in alphabet size. Fig. 6, therefore, confirms the advantage of using a vector modulator over a phase shifter. Besides, the results of the proposed expression derived in section V match the simulation results achieved by using the proposed Euclidean distance algorithm. Except for the 3×3 alphabet, the estimated values for IR are slightly higher than the simulated values due to ignoring the power reduction of the desired signal in the proposed expression.

VI. CONCLUSION

This study investigated an efficient and fast method to determine beamforming weights that work properly in realistic scenarios. Although the exhaustive search is quite efficient in interference suppression, it is a process with high computational complexity that makes it impractical to use for large alphabets. Hence, we proposed a much lower complexity method, which is called the Euclidean distance algorithm. The proposed algorithm decreases the process complexity at the cost of a few dB less interference rejection than exhaustive search. For instance, for four antennas and an alphabet size of 16, the number of calculations is reduced from 65536 to 64 computations at the price of 5 dB lower interference mitigation.

The second goal of this research was to show the effectiveness of utilizing vector modulators rather than phase shifters in terms of interference rejection. We designed a new hybrid beamforming system equipped with a vector modulator in the analog beamformer, followed by a digital beamformer at the baseband. Thanks to variable phase shifts and amplitude gains, the interference mitigation performance is better than beamformers with only phase shifts.

As a final contribution, an expression was proposed to quantify the interference rejection obtained in the RF domain. The expression is derived based on the proposed Euclidean distance algorithm. The approximated values fitted the simulation results with less than 1 dB error.

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