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Positional Modulation Design Based on Multiple Phased Antenna Arrays

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ABSTRACT Traditional directional modulation (DM) designs are normally based on a single phased array, with the assumption that eavesdroppers and desired users are in different directions. However, it is not always the case, as it is possible that they are in the same direction. As a result, signals received by eavesdroppers will be approximately the same as desired users. To solve the problem, a multiple antenna array model is introduced, and the principle of the design is that the eavesdropper located in the same direction as the desired user for one antenna array may not be in the same direction for another antenna array. Based on such a model, in this paper, a power minimisation constraint for the design is proposed, and an artificial noise component is included, resulting in a scrambled phase response for eavesdroppers. Beam pattern, phase pattern and BER are used as metrics to show the effectiveness of the designs based on multiple fixed antenna arrays and location optimised antenna arrays, respectively.

INDEX TERMS Directional modulation, multiple phased antenna arrays, positional modulation

I. INTRODUCTION

Directional modulation (DM) as a physical layer security technique was introduced in [1] for the first time to keep known constellation mappings in a desired direction or directions, while scrambling them for the remaining ones, by combining the direct radiation beam and reflected beams in the far-field. In [2], a reconfigurable array was designed by switching elements for each symbol to achieve DM. A method named dual beam DM was introduced in [3], where the I and Q signals are transmitted by different antennas. In [4], [5], phased arrays were employed to show that DM can be implemented by phase shifting the transmitted antenna signals properly. To further increase the channel capacity, two design methods were proposed. One is the multi-carrier based phased antenna array design for DM in [6], where an inverse Discrete Fourier Transform (IDFT) structure was exploited. The other is to use crossed-dipole antenna array as transmitter in [7], where DM and signal's orthogonal polarisation were first combined together in the proposed design. In [8], a design to enhance the transmission security by using direction dependent antenna modulation based on a directional antenna array with wide element spacing was presented. The directional error rate characteristics of the system can be significantly improved by replacing the isotropic elements by directional array elements. In [9], the BER was employed for DM transmitter synthesis by linking the BER performance to the settings of phase shifters. Static and dynamic interference were combined and added into the DM design in [10], with improved security achieved based on the same level of transmission power. A more systematic pattern synthesis approach was presented in [11], [12], where information pattern and interference patterns are created together to achieve DM, followed by an artificial-noise-aided zero-forcing synthesis approach in [13], and a multi-relay design in [14]. An eight-element time-modulated antenna array with constant instantaneous directivity in desired directions was proposed in [15]. The main idea of the design is that the array transmits signals without time modulation in the desired direction, while transmitting time-modulated signals in other directions. Due to the time modulation, the radiation pattern of the array changes with time.

Most of the designs are based on the assumption that eavesdroppers and desired users are in different directions; however, it is not always the case in practice. If eavesdroppers and desired users are in the same direction, then obviously traditional DM design to make null steering or scrambled constellations to directions of eavesdroppers is not going to work [12], [16]–[19]. To solve the problem, in [20] a multi-path model was introduced and employed, where signals via both line of sight (LOS) and reflected paths are combined at

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the receiver side. In this paper, we provide another design structure, where a multiple antenna array model is introduced. The idea of the design for solving the eavesdropper problem is that eavesdroppers in the same direction as desired users for one phased antenna array may not be in the same direction for another antenna array. Then signals from multiple phased antenna arrays can be combined at the receiver side, resulting in distinguishable signals for desired user locations and eavesdroppers. Based on this idea, we propose a power minimisation constraint for the design, with an additional artificial noise component included. Moreover, the sparse antenna array design problem in the context of positional modulation is also considered, with corresponding formulations provided.

The remaining part of this paper is structured as follows. A review of multiple phased arrays is given in Sec. II. Positional modulation (PM) design based on multiple antenna arrays is presented in Sec. III. Design examples are provided in Sec. IV, with conclusions drawn in Sec. V.

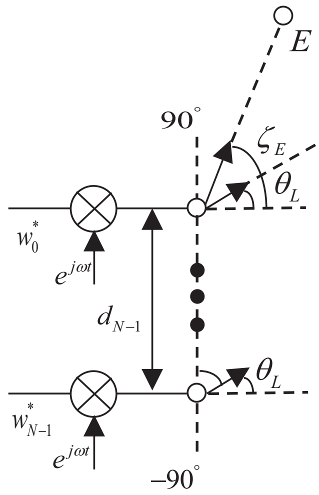

FIGURE 1: Directional modulation based on a single phased array.

II. REVIEW OF MULTIPLE PHASED ARRAYS

The objective of DM design is to find the set of weight coefficients giving the desired constellation values to the desired direction or directions, while making the magnitude of signal as low as possible and phase scrambled for the rest of the transmit angles. For most of the current designs, the eavesdropper E is assumed not to be in the same direction as the desired receiver L, as shown in Fig. 1. Here the transmission angle $\theta_L \in [-90°, 90°]$ for the desired location and $\zeta_E \in [-90°, 90°]$ for the eavesdropper. The spacing between the n-th to the zeroth antenna is represented by d_n for $n = 1, \ldots, N-1$. The weight coefficient for the n-th antenna is denoted by w_n ($n = 0, \ldots, N-1$).

The steering vector of the array for desired locations is a function of angular frequency ω and transmission angle θ_L, given by

$$\mathbf{s}_L(\omega, \theta_L) = [1, e^{j\omega d_1 \sin\theta_L/c}, \ldots, e^{j\omega d_{N-1}\sin\theta_L/c}]^T, \quad (1)$$

where $\{\cdot\}^T$ is the transpose operation, and c is the speed of propagation. Similarly, the steering vector for eavesdroppers as a function of ω and ζ_E, can be represented by

$$\mathbf{s}_E(\omega, \zeta_E) = [1, e^{j\omega d_1 \sin\zeta_E/c}, \ldots, e^{j\omega d_{N-1}\sin\zeta_E/c}]^T. \quad (2)$$

Then, the beam response of the array for the desired location is given by

$$p_L(\theta_L) = \mathbf{w}^H \mathbf{s}_L(\omega, \theta_L), \quad (3)$$

and the response for the eavesdropper is

$$p_E(\zeta_E) = \mathbf{w}^H \mathbf{s}_E(\omega, \zeta_E), \quad (4)$$

where $\{\cdot\}^H$ represents the Hermitian transpose, and $\mathbf{w}$ is the weight vector including all corresponding coefficients

$$\mathbf{w} = [w_0, w_1, \ldots, w_{N-1}]^T. \quad (5)$$

However, eavesdroppers and desired user locations could be in the same direction, as shown in Fig. 2, and in this case their beam responses will be the same, as $\mathbf{s}_L(\omega, \theta_L) = \mathbf{s}_E(\omega, \zeta_E)$. To solve the problem, a multiple phased array system can be employed, as shown in Fig. 2.

Here without loss of generality, we assume there are in total K uniform linear arrays (ULA), each with N antennas. The desired position is represented by L, the eavesdropper is represented by E, and an obstacle, which can reflect the transmitted signals and cause interference to the system, is denoted by B. The distance from the k-th antenna array ($k = 0, \ldots, K-1$) to the desired receiver, eavesdropper and obstacle are represented by d_{kL}, d_{kE} and d_{kB}, respectively. Moreover, the weight coefficient for the n-th antenna of the k-th antenna array is denoted by $w_{k,n}$, $k = 0, \ldots, K-1$ and $n = 0, \ldots, N-1$.

To make phase shift in sidelobe regions more scrambled to improve transmission security, we add a vector, representing artificial noise $\mathbf{a}_k = [a_{k,0}, \ldots, a_{k,N-1}]^T$ for $k = 0, \ldots, K-1$, which is orthogonal to the steering vector for the desired position, i.e.,

$$\mathbf{a}_k^T \mathbf{s}_L(\omega, \theta_{kL}) = 0. \quad (6)$$

Given this constraint, any change of $\mathbf{a}_k$ will not affect the signal received at the desired user location. The effective coefficient $x_{k,n}$ for the n-th antenna of the k-th antenna array is given by

$$x_{k,n} = w_{k,n}^* + a_{k,n}. \quad (7)$$

Then all $x_{k,n}$ for $n = 0, \ldots, N-1$ for the k-th antenna array can be gathered as a vector, represented by

$$\mathbf{x}_k = [x_{k,0}, \ldots, x_{k,N-1}]^T. \quad (8)$$

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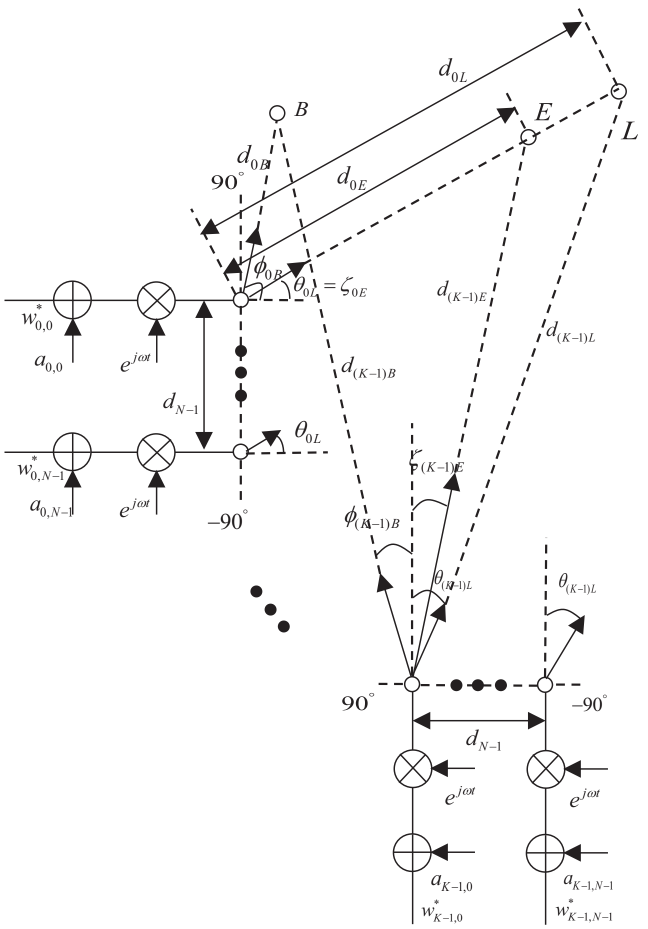

FIGURE 2: Directional modulation based on multiple phased arrays.

Therefore, the signal $y_L(\theta_L)$ received at the desired user is

$$\begin{aligned} y_L(\theta_L) &= \sum_{k=0}^{K-1} \mathbf{x}_k^T \mathbf{s}_L(\omega, \theta_{kL}) \\ &= \sum_{k=0}^{K-1} (\mathbf{w}_k^* + \mathbf{a}_k)^T \mathbf{s}_L(\omega, \theta_{kL}) \\ &= \sum_{k=0}^{K-1} \mathbf{w}_k^H \mathbf{s}_L(\omega, \theta_{kL}). \end{aligned} \tag{9}$$

Here $\mathbf{w}_k$ is the vector for all coefficients of the k-th antenna array

$$\mathbf{w}_k = [w_{k,0}, \ldots, w_{k,N-1}]^T, \tag{10}$$

and the steering vector $\mathbf{s}_L(\omega, \theta_{kL})$ is represented by

$$\mathbf{s}_L(\omega, \theta_{kL}) = [1, e^{j\omega d_1 \sin\theta_{kL}/c}, \ldots, e^{j\omega d_{N-1} \sin\theta_{kL}/c}]^T, \tag{11}$$

for $k = 0, \ldots, K-1$.

For signal $y_E(\zeta_E)$ received by the eavesdropper, due to the randomly generated artificial noise, the phase of $y_E(\zeta_E)$ is affected by artificial noise, and $y_E(\zeta_E)$ is not the same as $y_L(\theta_L)$.

$$\begin{aligned} y_E(\zeta_E) &= \sum_{k=0}^{K-1} \mathbf{x}_k^T \mathbf{s}_E(\omega, \zeta_{kE}) \\ &= \sum_{k=0}^{K-1} (\mathbf{w}_k^* + \mathbf{a}_k)^T \mathbf{s}_E(\omega, \zeta_{kE}) \\ &= \sum_{k=0}^{K-1} (\mathbf{w}_k^H \mathbf{s}_E(\omega, \zeta_{kE}) + \mathbf{a}_k^T \mathbf{s}_E(\omega, \zeta_{kE})), \end{aligned} \tag{12}$$

where $\mathbf{s}_E(\omega, \zeta_{kE})$ is given by

$$\mathbf{s}_E(\omega, \zeta_{kE}) = [1, e^{j\omega d_1 \sin\zeta_{kE}/c}, \ldots, e^{j\omega d_{N-1} \sin\zeta_{kE}/c}]^T. \tag{13}$$

Similarly, the signal received at obstacle B can be represented by

$$\begin{aligned} y_B(\phi_B) &= \sum_{k=0}^{K-1} \mathbf{x}_k^T \mathbf{s}_B(\omega, \phi_{kB}) \\ &= \sum_{k=0}^{K-1} (\mathbf{w}_k^H \mathbf{s}_B(\omega, \phi_{kB}) + \mathbf{a}_k^T \mathbf{s}_B(\omega, \phi_{kB})), \\ \mathbf{s}_B(\omega, \phi_{kB}) &= [1, e^{j\omega d_1 \sin\phi_{kB}/c}, \ldots, e^{j\omega d_{N-1} \sin\phi_{kB}/c}]^T. \end{aligned} \tag{14}$$

However, the above analysis is based on the assumption that there is no path loss and phase shift in the design. In practice, power attenuation and phase shift caused by different transmission paths need to be considered [16]. Here the attenuation ratio from the k-th antenna array for the path length d_{kL} to the desired location L, the path length d_{kE} to eavesdropper E, and the path length d_{kB} to obstacle B are given by [16]

$$\begin{aligned} \nu_{k,L} &= d/d_{kL}, \\ \xi_{k,E} &= d/d_{kE}, \\ \epsilon_{k,B} &= d/d_{kB}, \end{aligned} \tag{15}$$

where d is assumed to be the distance with unity power. Moreover, the phase shifts for the paths d_{kL}, d_{kE} and d_{kB} are given by

$$\begin{aligned} \psi_{k,L} &= 2\pi \times rem(d_{kL}, \lambda), \\ \rho_{k,E} &= 2\pi \times rem(d_{kE}, \lambda), \\ \beta_{k,E} &= 2\pi \times rem(d_{kB}, \lambda), \end{aligned} \tag{16}$$

where $rem(A, \lambda)$ represents the remainder of A divided by λ.

Then, in the multiple antenna array transmission model, with power attenuation and phase shift caused by different transmission paths, the signal received at desired user location can be represented by

$$y_L(\theta_L) = \sum_{k=0}^{K-1} \nu_{k,L} e^{j\psi_{k,L}} (\mathbf{w}_k^H \mathbf{s}_L(\omega, \theta_{kL})). \tag{17}$$

Similarly, the signal received at the eavesdropper and obstacle are given by

$$y_E(\zeta_E) = \sum_{k=0}^{K-1} \xi_{k,E} e^{j\rho_{k,E}} (\mathbf{x}_k^T \mathbf{s}_E(\omega, \zeta_{kE})), \tag{18}$$

$$y_B(\phi_B) = \sum_{k=0}^{K-1} \epsilon_{k,B} e^{j\beta_{k,B}} (\mathbf{x}_k^T \mathbf{s}_B(\omega, \phi_{kB})). \tag{19}$$

III. PROPOSED DM DESIGN BASED ON MULTIPLE PHASED ARRAYS

A. DESIGN BASED ON GIVEN ANTENNA ARRAYS

Here we assume R desired locations L_r for $r = 0, \ldots, R-1$ and Z eavesdroppers E_z for $z = 0, \ldots, Z-1$ in the design. Then for M-ary signaling, there are M sets of $p_{m,L_r}(\theta_{L_r})$ representing desired array responses for desired receiver location L_r and the m-th constellation point ($m = 0, \ldots, M-1$). The weight vector and artificial noise vector for the k-th antenna array and m-th constellation point are given by

$$\begin{aligned} \mathbf{w}_{m,k} &= [w_{m,k,0}, \ldots, w_{m,k,N-1}]^T, \\ \mathbf{a}_{m,k} &= [a_{m,k,0}, \ldots, a_{m,k,N-1}]^T. \end{aligned} \tag{20}$$

Moreover, we have $\mathbf{s}_{L_r}(\omega, \theta_{kL_r})$, $\mathbf{s}_{E_z}(\omega, \zeta_{kE_z})$ and $\mathbf{s}_{B_g}(\omega, \phi_{kB_g})$ to represent steering vectors for desired location L_r, eavesdropper E_z and obstacle B_g ($r = 0, \ldots, R-1$, $z = 0, \ldots, Z-1$ and $g = 0, \ldots, G-1$) from the k-th antenna array, respectively. For attenuation ratio and phase shift, we have ν_{k,L_r}, ξ_{k,E_z}, ϵ_{k,B_g}, ψ_{k,L_r}, ρ_{k,E_z} and β_{k,B_g}, represented by

$$\begin{aligned} \nu_{k,L_r} &= d/d_{kL_r}, \\ \xi_{k,E_z} &= d/d_{kE_z}, \\ \epsilon_{k,B_g} &= d/d_{kB_g}, \\ \psi_{k,L_r} &= 2\pi \times rem(d_{kL_r}, \lambda), \\ \rho_{k,E_z} &= 2\pi \times rem(d_{kE_z}, \lambda), \\ \beta_{k,B_g} &= 2\pi \times rem(d_{kB_g}, \lambda), \end{aligned} \tag{21}$$

denoting the attenuation ratio for desired location L_r, eavesdropper E_z and obstacle B_g, and phase shift for L_r, E_z and B_g from the k-th antenna array, respectively.

Then, based on the assumption that the receiver antenna is omni-directional (signals from all directions can be received), with the above parameters for the m-th constellation point, the coefficients can be formulated as

$$\begin{aligned} \min_{\mathbf{w}_{m,k}} \ & \sum_{z=0}^{Z-1} || \sum_{k=0}^{K-1} \xi_{k,E_z} e^{j\rho_{k,E_z}} (\mathbf{x}_{m,k}^T \mathbf{s}_{E_z}(\omega, \zeta_{kE_z})) ||_2 \\ \text{s.t.} \ & \sum_{k=0}^{K-1} \nu_{k,L_r} e^{j\psi_{k,L_r}} (\mathbf{w}_{m,k}^H \mathbf{s}_{L_r}(\omega, \theta_{kL_r})) = p_{m,Lr}(\theta_{L_r}) \\ & \mathbf{a}_{m,k}^T \mathbf{s}_{L_r}(\omega, \theta_{kL_r}) = 0 \\ & for \quad r = 0, \ldots, R-1 \\ & \qquad k = 0, \ldots, K-1, \end{aligned} \tag{22}$$

where $\mathbf{a}_{m,k}$ is the coefficient vector component orthogonal to the steering vector of the desired user location for the m-th constellation point. Here the cost function is to minimise the sum of signal power received at all eavesdroppers. Moreover, each time we solve the above problem given one constellation point, we randomly generate the orthogonal noise component $\mathbf{a}_{m,k}$ and for M-ary signaling, we have M sets of randomly generated $\mathbf{a}_{m,k}$. Given the randomness of the artificial noise vector, the phase response at eavesdroppers is scrambled. To increase the randomness of the phase response at eavesdroppers, for each constellation point, we could randomly generate more than one set of artificial noise vector $\mathbf{a}_{m,k}$ and find the corresponding $\mathbf{w}_{m,k}$, so that for the same symbol, we have multiple coefficient vector $\mathbf{x}_{m,k}$ to use. Therefore, phase constraint for signal received by eavesdroppers is not needed in the design, which is different from the previous design [20]. The first equality constraint is to keep the signal received at the desired location L_r the same as desired constellation points.

For the obstacle B_g ($g = 0, \ldots, G-1$) in the transmission range, we can set a null response to them, as signal to the obstacle may be reflected to the desired position, which causes interference to the received signal. Then for the m-th constellation point, the corresponding constraint for the obstacle B_g can be given by

$$\begin{aligned} & \sum_{k=0}^{K-1} \epsilon_{k,B_g} e^{j\beta_{k,B_g}} (\mathbf{x}_{m,k}^T \mathbf{s}_{B_g}(\omega, \phi_{kB_g})) = 0 \\ & for \quad g = 0, \ldots, G-1. \end{aligned} \tag{23}$$

Adding the constraint (23) to (22), to achieve a design with the minimum power and scrambled phase response for eavesdroppers, and the maximum power and the given phase shift for desired locations, without the effect of multipath caused during transmission, we have

$$\begin{aligned} \min_{\mathbf{w}_{m,k}} \ & \sum_{z=0}^{Z-1} || \sum_{k=0}^{K-1} \xi_{k,E_z} e^{j\rho_{k,E_z}} (\mathbf{x}_{m,k}^T \mathbf{s}_{E_z}(\omega, \zeta_{kE_z})) ||_2 \\ \text{s.t.} \ & \sum_{k=0}^{K-1} \nu_{k,L_r} e^{j\psi_{k,L_r}} (\mathbf{w}_{m,k}^H \mathbf{s}_{L_r}(\omega, \theta_{kL_r})) = p_{m,Lr}(\theta_{L_r}) \\ & \sum_{k=0}^{K-1} \epsilon_{k,B_g} e^{j\beta_{k,B_g}} (\mathbf{x}_{m,k}^T \mathbf{s}_{B_g}(\omega, \phi_{kB_g})) = 0 \\ & \mathbf{a}_{m,k}^T \mathbf{s}_{L_r}(\omega, \theta_{kL_r}) = 0 \\ & for \quad r = 0, \ldots, R-1 \\ & \qquad g = 0, \ldots, G-1 \\ & \qquad k = 0, \ldots, K-1. \end{aligned} \tag{24}$$

The above problem can be solved using cvx, a package for specifying and solving convex problems in Matlab [21], [22]. Note that all constellation points for a fixed θ_{kL_r} share the same steering vector and so does the steering vector for a fixed ζ_{kE_z}.

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B. DESIGN BASED ON LOCATION OPTIMISED ANTENNA ARRAYS

In this section, sparse antenna array design [23], [24] is considered, using the compressive sensing (CS) based method [25]–[28].

The principle of the sparse array design is to find the minimum number of antennas from a large number of potential antennas, with a similar or better performance than ULA. The criterion of removing an antenna location is that the corresponding weight coefficient is lower than or equal to a given threshold. Then for the m-th constellation point, with an omni-directional antenna receiver the optimised antenna locations can be found by

$$\begin{aligned}
\min_{\mathbf{w}_{m,k}} \quad & \sum_{k=0}^{K-1} ||\mathbf{w}_{m,k}||_1 \\
\text{s.t.} \quad & \sum_{z=0}^{Z-1} || \sum_{k=0}^{K-1} \xi_{k,E_z} e^{j\rho_{k,E_z}} (\mathbf{x}_{m,k}^T \mathbf{s}_{E_z}(\omega, \zeta_{kE_z}))||_2 \leq \alpha \\
& \sum_{k=0}^{K-1} \nu_{k,L_r} e^{j\psi_{k,L_r}} (\mathbf{w}_{m,k}^H \mathbf{s}_{L_r}(\omega, \theta_{kL_r})) = p_{m,Lr}(\theta_{L_r}) \\
& \sum_{k=0}^{K-1} \epsilon_{k,B_g} e^{j\beta_{k,B_g}} (\mathbf{x}_{m,k}^T \mathbf{s}_{B_g}(\omega, \phi_{kB_g})) = 0 \\
& \mathbf{a}_{m,k}^T \mathbf{s}_{L_r}(\omega, \theta_{kL_r}) = 0 \\
& for \quad r = 0, \ldots, R-1 \\
& \qquad g = 0, \ldots, G-1 \\
& \qquad k = 0, \ldots, K-1,
\end{aligned} \tag{25}$$

where $||\cdot||_1$ represents the l_1 norm, and α denotes a small value for the power of eavesdroppers. However, the above (25) cannot guarantee the same optimised antenna locations for all M constellation points. To find a common set of antenna locations, we introduce group sparsity [5], [6], and then replace the cost function in (25) by

$$\min_{\mathbf{w}_{m,k}} \quad \sum_{k=0}^{K-1} ||\hat{\mathbf{w}}||_1, \tag{26}$$

with the same constraints as in (25), where

$$\tilde{\mathbf{w}}_{k,n} = [w_{m,k,n}, \ldots, w_{M-1,k,n}], \tag{27}$$

$$\hat{\mathbf{w}}_k = [||\tilde{\mathbf{w}}_{k,0}||_2, ||\tilde{\mathbf{w}}_{k,1}||_2, \ldots, ||\tilde{\mathbf{w}}_{k,N-1}||_2]^T, \tag{28}$$

$$\hat{\mathbf{w}} = [\hat{\mathbf{w}}_0, \ldots, \hat{\mathbf{w}}_{K-1}]^T. \tag{29}$$

Here $\tilde{\mathbf{w}}_{k,n}$ represents the coefficients of the n-th antenna for the k-th antenna array and m-th symbol, $\hat{\mathbf{w}}_k$ is to calculate the minimum number of antennas for the k-th antenna array, and $\hat{\mathbf{w}}$ is for the minimum number of antennas for all K antenna arrays. Note: to remove the n-th antenna of the k-th array, ideally all coefficients at this antenna location need to be zero-valued.

As the reweighted l_1 norm minimisation has a closer approximation to the l_0 norm [29]–[31], we can modify (26) into the reweighted form in a similar way as in [5], where at the u-th iteration the PM design with an omni-directional antenna receiver can be formulated by

$$\begin{aligned}
\min_{\mathbf{w}_{m,k}} \quad & \sum_{k=0}^{K-1} \sum_{n=0}^{N-1} \delta_{k,n}^u ||\tilde{\mathbf{w}}_{k,n}^u||_2 \\
\text{s.t.} \quad & \sum_{z=0}^{Z-1} || \sum_{k=0}^{K-1} \xi_{k,E_z} e^{j\rho_{k,E_z}} ((\mathbf{x}_{m,k}^u)^T \mathbf{s}_{E_z}(\omega, \zeta_{kE_z}))||_2 \leq \alpha \\
& \sum_{k=0}^{K-1} \nu_{k,L_r} e^{j\psi_{k,L_r}} ((\mathbf{w}_{m,k}^u)^H \mathbf{s}_{L_r}(\omega, \theta_{kL_r})) = p_{m,Lr}(\theta_{L_r}) \\
& \sum_{k=0}^{K-1} \epsilon_{k,B_g} e^{j\beta_{k,B_g}} ((\mathbf{x}_{m,k}^u)^T \mathbf{s}_{B_g}(\omega, \phi_{kB_g})) = 0 \\
& \mathbf{a}_{m,k}^T \mathbf{s}_{L_r}(\omega, \theta_{kL_r}) = 0 \\
& for \quad r = 0, \ldots, R-1 \\
& \qquad g = 0, \ldots, G-1 \\
& \qquad k = 0, \ldots, K-1.
\end{aligned} \tag{30}$$

Here the superscript u indicates the u-th iteration, and $\delta_{k,n}$ is the reweighting term for the n-th row of coefficients at the k-th antenna array, given by

$$\delta_{k,n}^u = (||\tilde{\mathbf{w}}_{k,n}^{u-1}||_2 + \gamma)^{-1}, \tag{31}$$

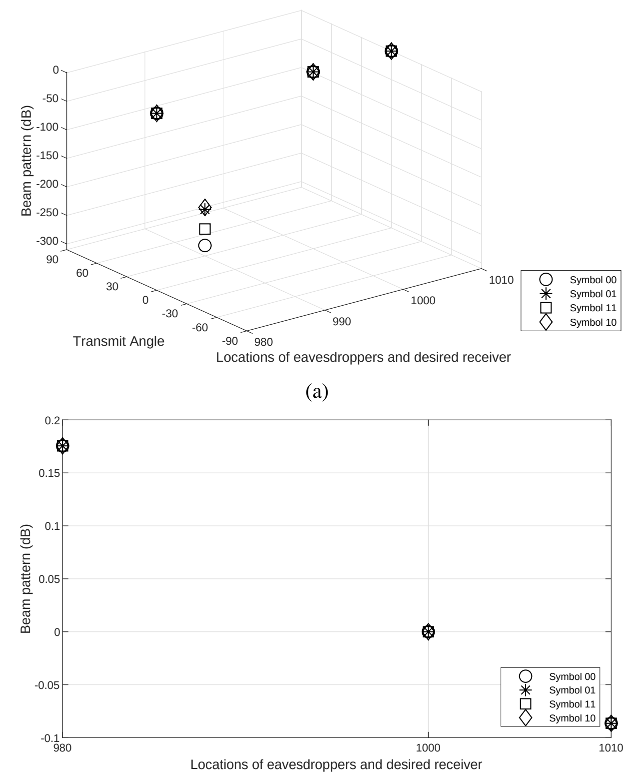
where $\gamma > 0$ is required to provide numerical stability and the iteration process is described as in [5].

IV. DESIGN EXAMPLES

Without loss of generality, we assume $K = 2$ antenna arrays with $N = 10$ antennas for each array. In the design, one desired location, two eavesdroppers and one obstacle are considered. In detail, we assume for the zeroth antenna array, the desired location L_0 and eavesdroppers E_0 and E_1 are all located in the same direction, where $\theta_{0L_0} = \zeta_{0E_0} = \zeta_{0E_1} = 0°$, with the distance $d = d_{0L_0} = 1000\lambda$, $d_{0E_0} = 980\lambda$ and $d_{0E_1} = 1010\lambda$. The obstacle B_0 is in the transmit angle $\phi_{0B_0} = 30°$, with the distance $d_{0B_0} = 990\lambda$. For the first antenna array the desired location L_0 is in the transmit angle $\theta_{1L_0} = 0°$ with the distance $d_{1L_0} = 1000\lambda$. Based on the geometry of the above design example, we can deduce ζ_{1E_0}, ζ_{1E_1}, ϕ_{1B_0}, d_{1E_0}, d_{1E_1} and d_{1B_0}

$$\begin{aligned}
\zeta_{1E_0} &\approx 1.1458°, \\
\zeta_{1E_1} &\approx -0.5729°, \\
\phi_{1B_0} &\approx 5.45°, \\
d_{1E_0} &\approx 1000.2, \\
d_{1E_1} &\approx 1000, \\
d_{1B_0} &\approx 1501.8.
\end{aligned} \tag{32}$$

Moreover, the desired response for desired location is a value of one (magnitude, the gain is 0dB) with 90° phase shift (QPSK), i.e. symbols '00', '01', '11', '10' correspond to 45°, 135°, −135° and −45°, respectively. For sparse antenna array design $\gamma = 0.001$, representing that antennas associated



(a)

(b)

FIGURE 3: Resultant beam patterns in (a) 3D and (b) 2D based on the single ULA design.

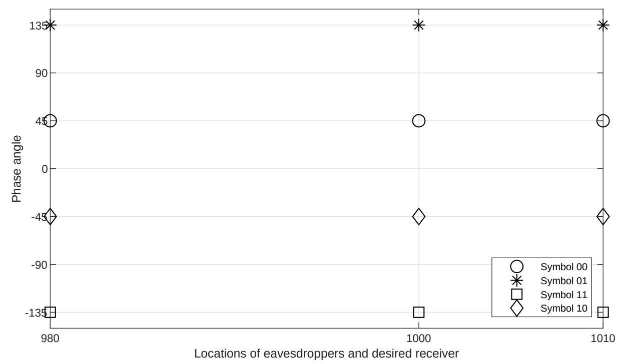


FIGURE 4: Phase pattern for the eavesdroppers and desired receiver based on the single ULA design.

with a weight value smaller than or equal to 0.001 will be removed.

For comparison, we provide three-dimensional (3D) and two-dimensional (2D) beam patterns, and phase response for DM design, based on a single phased antenna array and multiple phased antenna arrays to represent the effectiveness of our proposed method. Moreover, in the design, the bit error rate (BER) result for these locations is presented and calculated based on in which quadrant the received point lies

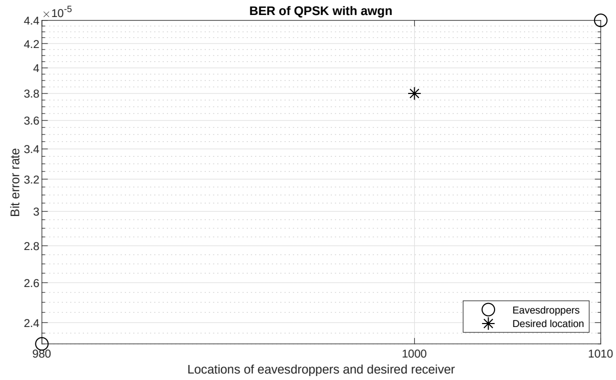


FIGURE 5: BER for the eavesdroppers and desired receiver based on the single ULA design.

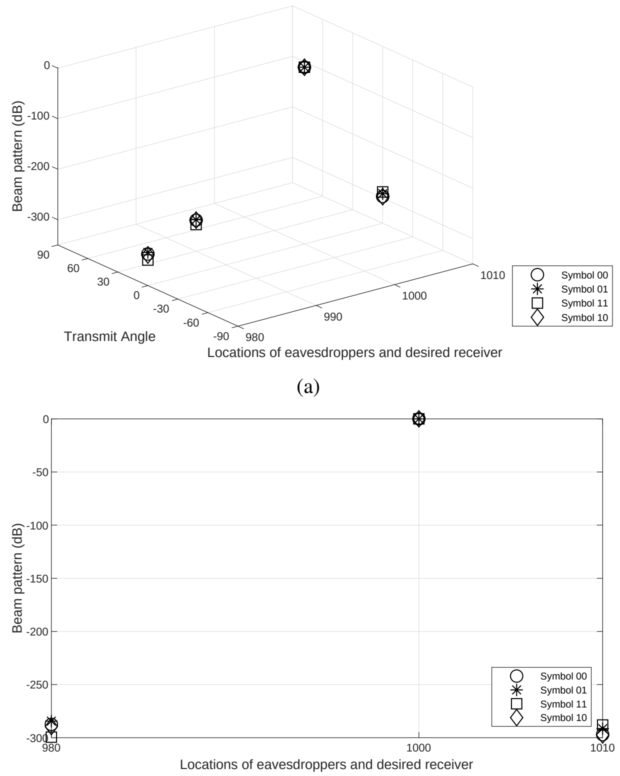
in the complex plane

$$BER = \frac{Error\ bits}{Total\ number\ of\ bits}. \tag{33}$$

Here the signal to noise ratio (SNR) is set at 12 dB at the desired location, and then with the unit average power of all randomly generated 10^6 transmitted bits, the noise variance σ^2 is 0.0631. Moreover, we assume that the additive white Gaussian noise (AWGN) level is the same for all eavesdroppers, and a random noise with this power level is then generated.

For the single antenna array design, the 3D beam response is shown in Fig. 3(a), where we can see that the response for the obstacle B_0 is the lowest, down to -300 dB at the transmit angle $30°$. The responses for the eavesdroppers and desired location are approximately the same. To further show the difference between the eavesdroppers and desired location, we zoom in Fig. 3(a) and see it from the 'transmit angle' axis, which leads to a 2D figure, as shown in Fig. 3(b). It can be seen that the responses for the eavesdropper E_0 with the shortest distance d_{0E_0} is the highest, while the response for the eavesdropper E_1 is the lowest, due to the longest distance d_{0E_1}. In other words, the shorter the distance, the higher the power of beam response. The phase response for these eavesdroppers and desired location is shown in Fig. 4, where the phase for all three locations are the same as the given constellation points (QPSK). Fig. 5 shows the BER for these locations, where BERs are all down to 10^{-5}, meaning that signals received by eavesdroppers can be demodulated correctly. Therefore, it can be concluded that with the single antenna array, if eavesdroppers and desired location are in the same direction, then DM cannot distinguish them, as the phase of signals received by these locations are the same, and the only difference is the power, which depends on the distance of the corresponding transmission path.

For the multiple antenna array design with an omni-directional antenna receiver in (24), the resultant 3D and 2D beam responses for the eavesdroppers, obstacle and desired location are shown in Fig. 6(a) and 6(b), where we can see that the responses for the obstacle B_0 and eavesdroppers



(a)

(b)

FIGURE 6: Resultant beam patterns in (a) 3D and (b) 2D based on the multiple ULA design (24).

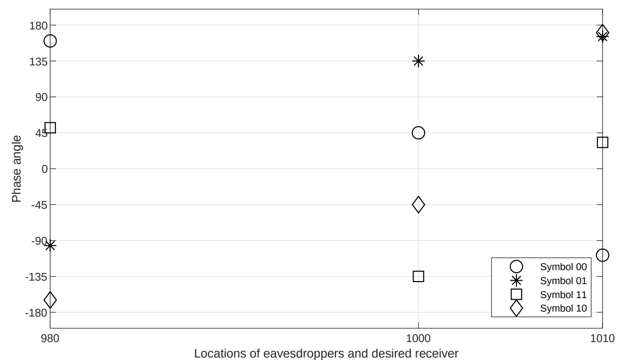


FIGURE 7: Phase pattern for the eavesdroppers and desired receiver based on the multiple ULA designs (24).

E_0 and E_1 are all down to -300 dB, while the response for the desired location is 0 dB, overcoming the problem associated with the single phased array case shown in Fig. 3. Moreover, Fig. 7 shows that the phase for all symbols at the eavesdroppers is distorted, while at the desired location L_0 the phase pattern is exactly the same as QPSK. The corresponding BER is shown in Fig. 8, where only at the desired location, BER is low to 10^{-5}, while at eavesdroppers it is around 0.5, further demonstrating the effectiveness of our proposed method. The beam pattern, phase pattern and BER for the corresponding sparse array design in (30) are similar

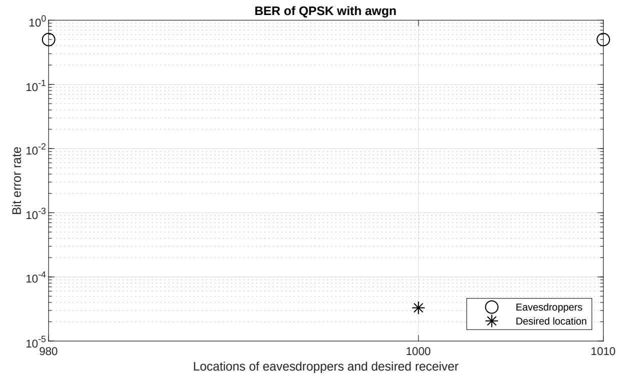


FIGURE 8: BERs patterns for the eavesdroppers and desired receiver based on the multiple ULA design (24).

to the design for the ULA in (24).

V. CONCLUSIONS

In this paper, a multiple antenna array transmission model has been studied for positional modulation, where signals from multiple phased antenna arrays are combined at the receiver side. The idea of the design is that the eavesdropper in the same direction as the desired user location for one phased array may not be in the same direction for another phased array. To achieve PM in the design, a minimum power constraint is proposed, and the phase pattern response at the eavesdroppers does not need to be controlled explicitly, due to the introduced artificial noise component. To avoid interference from obstacles distorting signals received at the desired location, the corresponding null steering constraint is also introduced. Moreover, we have also considered the sparse antenna array design problem in the context of positional modulation, with corresponding formulations provided. By the proposed designs, the minimum power and scrambled phase of signals for eavesdroppers, and the maximum power and given phase shift for desired locations have been achieved, while at the same time avoiding the multipath effect caused by obstacles.

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