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Robust Synthesis Scheme for Secure Multi-Beam Directional Modulation in Broadcasting Systems

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ABSTRACT Recently, directional modulation has become an active research area in wireless communications due to its security. Unlike existing research work, we consider a multi-beam directional modulation (MBDM) scenario with imperfect desired direction knowledge. In such a setting, a robust synthesis scheme is proposed for MBDM in broadcasting systems. In order to implement the secure transmission of a confidential message, the beamforming vector of the confidential message is designed to preserve its power as possible in the desired directions by minimizing its leakage to the eavesdropper directions while the projection matrix of artificial noise (AN) is to minimize the effect on the desired directions and force AN to the eavesdropper directions by maximizing the average receive signal-to-artificial-noise ratio at desired receivers. Simulation results show that compared with conventional methods, the proposed robust scheme achieves much better bit error rate performance along desired directions for a given signal-to-noise ratio (SNR). From the secrecy-rate aspect, the proposed scheme performs better than conventional methods for almost all SNR regions. In particular, in the medium and high SNR regions, the rate improvement of the proposed scheme over conventional methods is significant.

INDEX TERMS Multi-beam, directional modulation, robust, artificial noise, projection matrix, leakage, secrecy rate, bit error rate.

I. INTRODUCTION

Compared to wired communications, wireless communications are confronted with more serious security issues due to their broadcasting nature. Thus, in recent years, secure physical-layer wireless transmission has become an extremely important research area in wireless networks [1]–[15]. The seminal works in [1] and [2] reveal that secrecy capacity is achievable when the eavesdropper has a weaker channel than the desired receiver. In addition to conventional key-based cryptographic technologies employed at the higher protocol layers for secure data transmission, wireless transmission with physical-layer secrecy provides another security [3]–[5], which makes wireless communication ultra secure.

Recently, directional modulation (DM) has emerged as one of key physical layer security techniques in wireless communication [6]–[13]. In [6] and [7], the authors propose a synthesis method for DM, which relies on the intergrating reflectors and switches in the near-field of an on-chip dipole antenna. In a far-field scenario, a similar modulation method is proposed in [8] and [9]. In the above methods, the DM synthesis is achievable by applying signals directly onto the radio frequency (RF) frontend. Due to the need for high-speed RF switches or RF phase shifters, the development and applications of the RF-based synthesis techniques for DM are limited. In the recent years, a novel digital basedband synthesis method is proposed by the authors in [10] and [11], which realizes the DM baseband synthesis by designing

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excitation signal and artificial noise (AN) vectors. The above methods are restricted to the single-direction DM systems. In [12], the authors propose an orthogonal projection (OP) method to design the weights for the desired direction and artificial noise in multi-beam DM systems. Nevertheless, the method was somewhat complex because we must obtain the power pattern along the desired direction first and generate the orthogonal basis corresponding to the nulls of the power pattern of transmit antenna array. For their calculations, perfect direction knowledge must be available at the transmitter. However, in a practical wireless communication system, the transmitter obtains only imperfect direction knowledge, and in [13] the authors propose a new robust DM synthesis method for single-desired directional modulation by considering the direction measurement errors. We focus on a multi-beam directional modulation scenario, in which the useful/confidential message is broadcasted towards multiple desired receivers with different directions.

In this paper, we investigate a robust multi-beam direction modulation (MBDM) synthesis in broadcasting system. Firstly, in the scenario with perfect desired direction angle information, we present a hybrid beamforming method, which designs the AN projection matrix by maximizing the signal-to-artificial-noise (AN) ratio (Max-SANR) at the desired receivers, and the beamforming vector of the useful signal by minimizing its leakage at the transmitter to eavesdroppers, called maximizing signal-to-leakage-noise ratio (Max-SLNR). Subsequently, in the presence of the desired direction angle measurement errors, we redesign the corresponding robust Max-SANR and Max-SLNR by using the idea of conditional expectation, respectively. Their detailed expressions are derived under two different scenarios: imperfect and unknown eavesdropper directions. Finally, by simulation, the performance of the proposed methods are verified in a multi-direction modulation broadcasting scenario.

The rest of the paper is organized as follows: Section II describes the system model. In Section III, we propose a high-performance synthesis method with the perfect knowledge of desired directions under two cases: perfect and unknown directions of eavesdroppers. In Section IV, when the imperfect desired direction knowledge is assumed, we propose a robust synthesis method of Max-SANR and Max-SLNR by using conditional expectation. Simulation results are shown in section V. Finally, we draw our conclusions in Section VI.

Notations: throughout the paper, matrices, vectors, and scalars are denoted by letters of bold upper case, bold lower case, and lower case, respectively. Signs $(\cdot)^T$, $(\cdot)^{-1}$, $(\cdot)^{\dagger}$, $(\cdot)^H$, and $\mathrm{tr}(\cdot)$ denote transpose, inverse, Moore-Penrose pseudo-inverse, conjugate transpose and trace, respectively. Operation $\|\cdot\|$ denotes the norm of a complex number. The notation $\mathbb{E}\{\cdot\}$ refers to the expectation operation. Matrices $\mathbf{I}_N$ and $\mathbf{0}_{N\times M}$ denote the $N \times N$ identity matrix and $N \times M$ matrix of all zeros, respectively. Set S_n denotes the integer set $\{1, 2, \cdots, n\}$.

II. SYSTEM MODEL AND PROBLEM STATEMENT

Let us consider a basic multi-beam directional modulation (MBDM) broadcasting system, which consists of a transmitter with one N-element linear antenna array, M desired receivers, and K eavesdroppers. In such a system, $\Theta_d = \left\{\theta_d^1, \cdots, \theta_d^M\right\}$ denotes the set of desired directions, and $\Theta_e = \left\{\theta_e^1, \cdots, \theta_e^K\right\}$ denotes the set of eavesdropper directions, where θ_d^m and θ_e^n are the direction of desired receiver m and eavesdropper n, respectively. At the transmitter, the transmit signal is made up of two parts: a confidential message and an artificial noise signal. The former is forwarded to the desired receivers by using the useful beamforming vector and the latter is used to interfere with the eavesdroppers by the projection matrix.

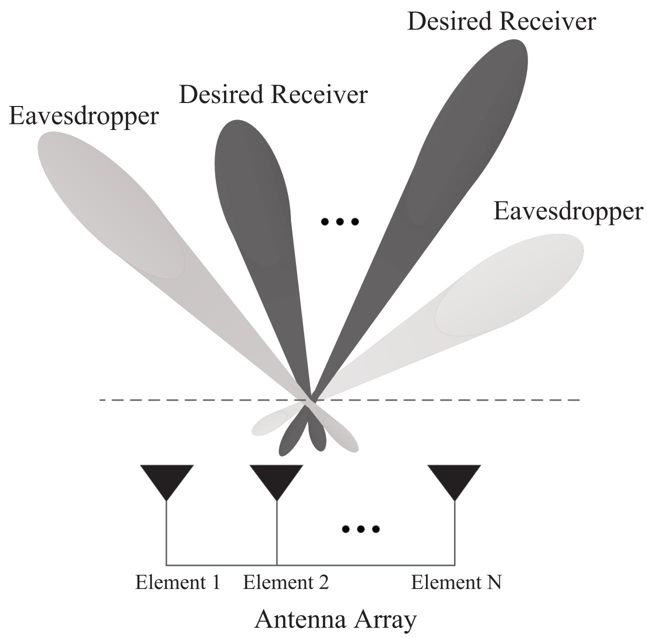

FIGURE 1. Schematic diagram of multi-direction modulation broadcasting system.

In general, the DM technique in Fig. 1 can be only used in line-of-sight (LoS) channel. How to extend the DM to the fading channel scenario is a challenging problem. In the paper, we will still focus on the synthesis method of MBDM in a LoS channel. The N-tuple transmitted signal vector is expressed as

$$\mathbf{s} = \beta_1\sqrt{P_s}\mathbf{v}_d x_d + \alpha\beta_2\sqrt{P_s}\mathbf{P}_{AN}\mathbf{z} \tag{1}$$

where x_d is the transmit confidential message symbol, P_s is the total transmit power constraint, β_1 and β_2 are the factors for controlling power with $\beta_1^2+\beta_2^2 = 1$, $\mathbf{z}$ is an N-tuple vector with the probability density function (PDF) being $\mathcal{CN}(0, \mathbf{I}_N)$, called artificial noise (AN), and α denotes the normalization factor with $\alpha^2\mathbb{E}\left\{\mathrm{tr}\left[\mathbf{P}_{AN}\mathbf{z}\mathbf{z}^H\mathbf{P}_{AN}^H\right]\right\} = 1$. In the above equation, $\mathbf{v}_d$ is the transmit beamforming vector of useful data, and $\mathbf{P}_{AN}$ is the projection matrix for controlling the direction of AN. Below, our major goal is to design or optimize $\mathbf{v}_d$ and $\mathbf{P}_{AN}$ such that both high secure rate and low bit error rate can be achieved in the presence of perfect/imperfect desired direction information. In (1), due to the use of $\mathbf{v}_d$ and $\mathbf{P}_{AN}$, the useful information x_d is broadcasted towards all desired receivers without being intercepted by eavesdroppers, as indicated in Fig. 1.

Providing that $\mathbf{v}_d$ and $\mathbf{P}_{AN}$ are well designed, after the transmit signal vector in (1) passes through the antenna array and the LoS channel, the received signal at any direction θ is

$$\begin{aligned} y(\theta) &= \mathbf{h}^H(\theta)\mathbf{s} + w_\theta \\ &= \underbrace{\beta_1\sqrt{P_s}\mathbf{h}^H(\theta)\mathbf{v}_d x_d}_{\text{Useful signal}} + \underbrace{\alpha\,\beta_2\sqrt{P_s}\mathbf{h}^H(\theta)\mathbf{P}_{AN}\mathbf{z}}_{\text{Artificial noise}} + \underbrace{w_\theta}_{\text{Noise}}, \end{aligned} \tag{2}$$

where the steering vector $\mathbf{h}^H(\theta)$ is defined by

$$\mathbf{h}^H(\theta) \triangleq \frac{1}{\sqrt{N}}\left[e^{j2\pi\psi_\theta(1)} \cdots e^{j2\pi\psi_\theta(n)} \cdots e^{j2\pi\psi_\theta(N)}\right] \tag{3}$$

with the phase function $\psi_\theta(n)$ given by

$$\psi_\theta(n) \triangleq \frac{(n-(N+1)/2)L_t\cos\theta}{\lambda},\quad n = 1, 2, \ldots, N \tag{4}$$

where θ is the direction angle, L_t denotes the transmit antenna separation, and λ is the wavelength of the transmit signal carrier.

The received signals along the mth desired and the kth eavesdropper directions are

$$y(\theta_d^m) = \mathbf{h}^H(\theta_d^m)\mathbf{s} + w_d^m, \tag{5}$$

and

$$y(\theta_e^k) = \mathbf{h}^H(\theta_e^k)\mathbf{s} + w_e^k, \tag{6}$$

respectively, where w_d^m and w_e^k denote the receive additive white Gaussian noises with PDFs being $w_d^m \sim \mathcal{CN}(0, \sigma_d^2)$ and $w_e^k \sim \mathcal{CN}(0, \sigma_e^2)$ at desired receiver m and eavesdropper receiver k, respectively,

To design $\mathbf{v}_d$ and $\mathbf{P}_{AN}$, we have to collect all the received signals along all desired and eavesdropper directions. Let us stack all desired direction receive signals in (5) together as a desired received vector $\mathbf{y}(\Theta_d)$ and all eavesdropper receive signals in (6) together as an eavesdropper received vector $\mathbf{y}(\Theta_e)$ as follows

$$\begin{aligned} \mathbf{y}(\Theta_d) &= \mathbf{H}^H(\Theta_d)\mathbf{s} + \mathbf{w}_d \\ &= \beta_1\sqrt{P_s}\mathbf{H}^H(\Theta_d)\mathbf{v}_d x_d + \alpha\beta_2\sqrt{P_s}\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN}\mathbf{z} + \mathbf{w}_d, \end{aligned} \tag{7}$$

and

$$\begin{aligned} \mathbf{y}(\Theta_e) &= \mathbf{H}^H(\Theta_e)\mathbf{s} + \mathbf{w}_e \\ &= \beta_1\sqrt{P_s}\mathbf{H}^H(\Theta_e)\mathbf{v}_d x_d + \alpha\beta_2\sqrt{P_s}\mathbf{H}^H(\Theta_e)\mathbf{P}_{AN}\mathbf{z} + \mathbf{w}_e, \end{aligned} \tag{8}$$

respectively, where $\mathbf{y}(\Theta_d) = [y(\theta_d^1),\ y(\theta_d^2),\ \cdots,\ y(\theta_d^M)]^T$ is the vector of the received signals along all desired directions, $\mathbf{y}(\Theta_e) = [y(\theta_e^1),\ y(\theta_e^2),\ \cdots,\ y(\theta_e^K)]^T$ is the vector of the received signals along all eavesdropper directions, and the conjugate transposes of the desired and eavesdropper channel matrices $\mathbf{H}^H(\Theta_d)$ and $\mathbf{H}^H(\Theta_e)$ are

$$\mathbf{H}(\Theta_d) = \left[\mathbf{h}(\theta_d^1) \quad \mathbf{h}(\theta_d^2) \cdots \mathbf{h}(\theta_d^M)\right], \tag{9}$$

and

$$\mathbf{H}(\Theta_e) = \left[\mathbf{h}(\theta_e^1) \quad \mathbf{h}(\theta_e^2) \cdots \mathbf{h}(\theta_e^K)\right], \tag{10}$$

respectively. In Eqs. (7) and (8), $\mathbf{w}_d$ and $\mathbf{w}_e$ are the received $M \times 1$ and $K \times 1$ complex Gaussian vectors in the desired and eavesdropper directions, respectively. Here, $w_d^m \sim \mathcal{CN}(\mathbf{0}_{M\times 1}, \sigma_d^2\mathbf{I}_M)$ and $w_e^k \sim \mathcal{CN}(\mathbf{0}_{K\times 1}, \sigma_e^2\mathbf{I}_K)$.

According to the concept of secrecy rate in [1]–[3], we define the secrecy rate of MBDM systems as follows

$$R_s = \max_{m\in S_M}\left\{0,\ \min_{k\in S_K}(R(\theta_d^m) - R(\theta_e^k))\right\}, \tag{11}$$

where $R(\theta_d^m)$ is the achievable rate of the link from the transmitter to the desired receiver m and $R(\theta_e^k)$ is the achievable rate of the link from the transmitter to the eavesdropper k. Based on (7) and (8), $R(\theta_d^m)$ and $R(\theta_e^k)$ are

$$\begin{aligned} R(\theta_d^m) &\triangleq I(y(\theta_d^m); [x_d, \theta_d^m]) \\ &= \log_2\left(1 + \frac{\beta_1^2 P_s\mathbf{h}^H(\theta_d^m)\mathbf{v}_d\mathbf{v}_d^H\mathbf{h}(\theta_d^m)}{\sigma_d^2 + \alpha^2\beta_2^2 P_s\mathbf{h}^H(\theta_d^m)\mathbf{P}_{AN}\mathbf{P}_{AN}^H\mathbf{h}(\theta_d^m)}\right), \end{aligned} \tag{12}$$

and

$$\begin{aligned} R(\theta_e^k) &\triangleq I(y(\theta_e^k); [x_d, \theta_e^k]) \\ &= \log_2\left(1 + \frac{\beta_1^2 P_s\mathbf{h}^H(\theta_e^k)\mathbf{v}_d\mathbf{v}_d^H\mathbf{h}(\theta_e^k)}{\sigma_e^2 + \alpha^2\alpha^2\beta_2^2 P_s\mathbf{h}^H(\theta_e^k)\mathbf{P}_{AN}\mathbf{P}_{AN}^H\mathbf{h}(\theta_e^k)}\right), \end{aligned} \tag{13}$$

where the function $I(y; [x, \theta])$ denotes the mutual information along direction θ between the input x and the output y.

III. PROPOSED BEAMFORMING PROFILE WITH PERFECT KNOWLEDGE OF DESIRED DIRECTIONS

In the case of perfect knowledge of desired directions, we consider two scenarios: unknown and perfect eavesdropper direction knowledge. Under the two situations, we design the corresponding synthesis methods of MBDM by using the Max-SLNR and Max-SANR criterion.

A. PERFECT DIRECTIONS OF EAVESDROPPERS

Since we know directions of both desired receivers and eavesdroppers, it is easy to design $\mathbf{v}_d$ and $\mathbf{P}_{AN}$. By means of Eqs. (7) and (8), and the basic concept of leakage in [16] and [17], the useful beamforming vector $\mathbf{v}_d$ is given by the following optimization

$$\begin{aligned} &\max_{\mathbf{v}_d}\ \text{SLNR}(\mathbf{v}_d) \\ &\text{subject to } \mathbf{v}_d^H\mathbf{v}_d = 1, \end{aligned} \tag{14}$$

where

$$\mathrm{SLNR}(\mathbf{v}_d) = \frac{\beta_1^2 P_s \mathrm{tr}\left\{\mathbf{v}_d^H \mathbf{H}(\Theta_d)\mathbf{H}^H(\Theta_d)\mathbf{v}_d\right\}}{\mathrm{tr}\left\{\mathbf{v}_d^H(\beta_1^2 P_s \mathbf{H}(\Theta_e)\mathbf{H}^H(\Theta_e) + \sigma_e^2 \mathbf{I}_N)\mathbf{v}_d\right\}}, \tag{15}$$

where

$$\mathbf{G}(\Theta_d) = \mathbf{H}(\Theta_d)\mathbf{H}^H(\Theta_d), \tag{16}$$

and

$$\mathbf{G}(\Theta_e) = \mathbf{H}(\Theta_e)\mathbf{H}^H(\Theta_e), \tag{17}$$

which are called the Grammian matrices for the desired and eavesdropper directions, respectively. Using the generalized Rayleigh-Ritz theorem [18], the excitation vector $\mathbf{v}_d$ for maximizing the SLNR can be obtained from the eigenvector corresponding to the largest eigen-value of the matrix [16]–[20]

$$\left[\mathbf{H}(\Theta_e)\mathbf{H}^H(\Theta_e) + \frac{\sigma_e^2}{\beta_1^2 P_s}\mathbf{I}_N\right]^{-1} \mathbf{H}(\Theta_d)\mathbf{H}^H(\Theta_d). \tag{18}$$

Since we get the useful beamforming vector $\mathbf{v}_d$, we compute the AN projecting matrix by maximizing the average receive SANR at the desired receivers

$$\max_{\mathbf{P}_{AN}} \quad \mathrm{SANR}(\mathbf{P}_{AN},\ \mathbf{v}_d), \tag{19}$$

where

$$\mathrm{SANR}(\mathbf{P}_{AN},\ \mathbf{v}_d) \triangleq \frac{\beta_1^2 P_s \mathrm{tr}\left\{\mathbf{H}^H(\Theta_d)\mathbf{v}_d\mathbf{v}_d^H\mathbf{H}(\Theta_d)\right\}}{\alpha^2\left(1-\beta_1^2\right) P_s \mathrm{tr}\left\{\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN}\mathbf{P}_{AN}^H\mathbf{H}(\Theta_d)\right\} + \sigma_d^2}. \tag{20}$$

The optimization problem in (19) can be converted to the following equivalent problem

$$\min_{\mathbf{P}_{AN}} \quad \frac{\alpha^2\beta_2^2 P_s \mathrm{tr}\left\{\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN}\mathbf{P}_{AN}^H\mathbf{H}(\Theta_d)\right\}}{\beta_1^2 P_s \mathrm{tr}\left\{\mathbf{H}^H(\Theta_d)\mathbf{v}_d\mathbf{v}_d^H\mathbf{H}(\Theta_d)\right\}}. \tag{21}$$

Given $\mathbf{v}_d$, the denominator of the objective function in (21) is constant, then minimizing the objective function in (21) yields a zero numerator

$$\mathrm{tr}\left\{\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN}\mathbf{P}_{AN}^H\mathbf{H}(\Theta_d)\right\} = 0 \tag{22}$$

due to the nonnegative property of its numerator. The above identity is equivalent to

$$\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN} = \mathbf{0}_{M\times N}, \tag{23}$$

which means that all column vectors of projection matrix $\mathbf{P}_{AN}$ span the null space of the desired steering matrix $\mathbf{H}^H(\Theta_d)$. Based on this and assuming $N \geq M$, we construct the following orthogonal projector

$$\mathbf{P}_{AN} = \mathbf{I}_N - \mathbf{H}(\Theta_d)\left[\mathbf{H}^H(\Theta_d)\mathbf{H}(\Theta_d)\right]^{-1}\mathbf{H}^H(\Theta_d), \tag{24}$$

with $\mathbf{P}_{AN}^2 = \mathbf{P}_{AN}$, and

$$\begin{aligned}\mathbf{H}^H(\Theta_d)\mathbf{P}_{AN} &= \mathbf{H}^H(\Theta_d)\left\{\mathbf{I}_N - \mathbf{H}(\Theta_d)\left[\mathbf{H}^H(\Theta_d)\mathbf{H}(\Theta_d)\right]^{-1}\mathbf{H}^H(\Theta_d)\right\} \\ &= \mathbf{0}_{M\times N}.\end{aligned} \tag{25}$$

B. UNKNOWN DIRECTIONS OF EAVESDROPPERS

Conversely, when the directions of eavesdroppers are unavailable at the transmitter, we treat all directions outside the main lobes of all desired directions as the direction interval of eavesdroppers defined as

$$\Omega_e = [0, \pi] \setminus \bigcup_{n=1}^{M} \Omega_d^n, \tag{26}$$

where

$$\Omega_d^n = \left[\theta_d^n - \frac{\theta_{BW}}{2},\ \theta_d^n + \frac{\theta_{BW}}{2}\right] \tag{27}$$

with $\theta_{BW} = \frac{2\lambda}{NL_t}$ being the first null beamwidth (FNBW) [21]. Following this idea, by using Eqs. (2) and (7), we design $\mathbf{v}_d$ by maximizing the SLNR as the following optimization

$$\begin{aligned}&\max_{\mathbf{v}_d} \ \mathrm{SLNR}(\mathbf{v}_d) \\ &\text{subject to } \mathbf{v}_d^H\mathbf{v}_d = 1,\end{aligned} \tag{28}$$

where

$$\mathrm{SLNR}(\mathbf{v}_d) \triangleq \frac{\beta_1^2 P_s \mathrm{tr}\left\{\mathbf{v}_d^H\mathbf{G}(\Theta_d)\mathbf{v}_d\right\}}{\beta_1^2 P_s \mathrm{tr}\left\{\mathbf{v}_d^H\left[\oint_{\Omega_e}\mathbf{h}(\theta)\mathbf{h}^H(\theta)d\theta\right]\mathbf{v}_d\right\} + \sigma_w^2}, \tag{29}$$

which yields the optimal optimization variable $\mathbf{v}_d$ being the eigen-vector corresponding to the largest eigenvalue of the matrix

$$\left[\oint_{\Omega_e}\mathbf{h}(\theta)\mathbf{h}^H(\theta)d\theta + \frac{\sigma_w^2}{\beta_1^2 P_s}\mathbf{I}_N\right]^{-1}\mathbf{G}(\Theta_d). \tag{30}$$

Similar to the previous subsection, we can obtain the same projection matrix $\mathbf{P}_{AN}$ as shown in (24), which is independent of the directions of eavesdroppers and which only depends on the desired directions.

IV. PROPOSED ROBUST BEAMFORMING PROFILE WITH IMPERFECT KNOWLEDGE OF DESIRED DIRECTIONS

In a practical MBDM system, it is impossible for an MBDM transmitter to obtain the perfect direction angle of the desired and eavesdropper directions. The direction angles are available by some typical direction estimation methods like MUSIC, Capon [22], etc. In the presence of channel noise and interference, there always exists the measurement errors in the estimated directions. These measurement errors will have a direct impact on the design of the beamforming vector and the projection matrix. For the convenience of the derivation below, these measurement errors are modelled as independently identically distributed (i.i.d) random variables. The measured desired and eavesdropper direction vectors are expressed as

$$\underbrace{\left[\hat{\theta}_d^1,\ \hat{\theta}_d^2,\ \cdots,\ \hat{\theta}_d^M\right]^T}_{\hat{\Theta}_d} = \underbrace{\left[\theta_d^1,\ \theta_d^2,\ \cdots,\ \theta_d^M\right]^T}_{\Theta_d} + \underbrace{\left[\Delta\theta_d^1,\ \Delta\theta_d^2,\ \cdots,\ \Delta\theta_d^M\right]^T}_{\Delta\Theta_d}, \tag{31}$$

and

$$\underbrace{\left[\hat{\theta}_e^1,\ \hat{\theta}_e^2,\ \cdots,\ \hat{\theta}_e^K\right]^T}_{\hat{\Theta}_e} = \underbrace{\left[\theta_e^1,\ \theta_e^2,\ \cdots,\ \theta_e^K\right]^T}_{\Theta_e} + \underbrace{\left[\Delta\theta_e^1,\ \Delta\theta_e^2,\ \cdots,\ \Delta\theta_e^K\right]^T}_{\Delta\Theta_e}, \tag{32}$$

respectively, where the PDFs of $\Delta\theta_d^i$ and $\Delta\theta_e^i$ are approximated as the truncated Gaussian distributions

$$f_d(x) = \begin{cases} \frac{1}{K_d\sqrt{2\pi}\sigma_{\theta_d}} e^{-\frac{(x-\mu_{\theta_d})^2}{2\sigma_{\theta_d}^2}} & -\Delta x_d < x < \Delta x_d \\ 0 & x \in [-\pi, -\Delta x_d] \cup [\Delta x_d, \pi], \end{cases} \tag{33}$$

and

$$f_e(x) = \begin{cases} \frac{1}{K_e\sqrt{2\pi}\sigma_{\theta_e}} e^{-\frac{(x-\mu_{\theta_e})^2}{2\sigma_{\theta_e}^2}} & -\Delta x_e < x < \Delta x_e \\ 0 & x \in [-\pi, -\Delta x_e] \cup [\Delta x_e, \pi], \end{cases} \tag{34}$$

where K_d and K_e are the normalization factors defined as

$$K_d = \int_{-\Delta x_d}^{\Delta x_d} \frac{1}{\sqrt{2\pi}\sigma_{\theta_d}} e^{-\frac{x^2}{2\sigma_{\theta_d}^2}} d(x), \tag{35}$$

and

$$K_e = \int_{-\Delta x_e}^{\Delta x_e} \frac{1}{\sqrt{2\pi}\sigma_{\theta_e}} e^{-\frac{x^2}{2\sigma_{\theta_e}^2}} d(x). \tag{36}$$

In the above, the measurement errors $\Delta\theta_d^i$ and $\Delta\theta_e^i$ have the variances $\sigma_{\theta_d}^2$ and $\sigma_{\theta_e}^2$, respectively. The corresponding upper limits Δx_d and Δx_e of the above integral are $\Delta\theta_{d,\max}$ and $\Delta\theta_{e,\max}$, respectively.

A. IMPERFECT DIRECTIONS OF EAVESDROPPERS

Given all measured $\hat{\theta}_d^i$ and $\hat{\theta}_e^i$, by utilizing the prior knowledge of $\Delta\Theta_d$ and $\Delta\Theta_e$, we define the new conditional SLNR (CSLNR) as follows

$$\text{CSLNR}_1(\mathbf{v}_d) \triangleq \frac{\beta_1^2 P_s \text{tr}\left\{\mathbf{v}_d^H \mathbf{R}_d \mathbf{v}_d\right\}}{\text{tr}\left\{\mathbf{v}_d^H\left[\beta_1^2 P_s \mathbf{R}_e + \sigma_e^2 \mathbf{I}_N\right]\mathbf{v}_d\right\}}, \tag{37}$$

where

$$\mathbf{R}_d \triangleq \mathbb{E}\left\{\mathbf{H}(\Theta_d)\mathbf{H}^H(\Theta_d)|\hat{\Theta}_d\right\} = \mathbb{E}\left\{\mathbf{G}(\Theta_d)|\hat{\Theta}_d\right\}, \tag{38}$$

and

$$\mathbf{R}_e \triangleq \mathbb{E}\left\{\mathbf{H}(\Theta_e)\mathbf{H}^H(\Theta_e)|\hat{\Theta}_e\right\} = \mathbb{E}\left\{\mathbf{G}(\Theta_e)|\hat{\Theta}_e\right\} \tag{39}$$

where the detailed expressions of $\mathbf{R}_d$ and $\mathbf{R}_e$ are derived in Appendix A. Maximizing (37) forms the following optimization

$$\max_{\mathbf{v}_d}\ \text{CSLNR}_1(\mathbf{v}_d) \quad \text{subject to } \mathbf{v}_d^H\mathbf{v}_d = 1, \tag{40}$$

which gives the optimal beamforming vector $\hat{\mathbf{v}}_d$ being the eigenvector corresponding to the largest eigen-value of the matrix

$$\left[\mathbf{R}_e + \frac{\sigma_e^2}{\beta_1^2 P_s}\mathbf{I}_N\right]^{-1}\mathbf{R}_d. \tag{41}$$

In the same manner, given $\hat{\Theta}_d$ and the PDF of $\Delta\Theta_d$, we can define the conditional expectation of projection matrix $\mathbf{P}_{AN}$

$$\mathbb{E}\left\{\mathbf{P}_{AN}|\hat{\Theta}_d\right\} = \mathbf{I}_N - \underbrace{\mathbb{E}\left\{\mathbf{H}(\Theta_d)\left[\mathbf{H}^H(\Theta_d)\mathbf{H}(\Theta_d)\right]^{-1}\mathbf{H}^H(\Theta_d)|\hat{\Theta}_d\right\}}_{\mathbf{P}_d}. \tag{42}$$

Obviously, it is not easy to compute the above conditional expectation $\mathbf{P}_d$ in the above expression. For simplification, $\mathbf{P}_d$ is replaced by the following easy-to-calculate formula

$$\mathbf{P}_d = \mathbb{E}\left\{\mathbf{H}(\Theta_d)\mathbf{B}\mathbf{H}^H(\Theta_d)|\hat{\Theta}_d\right\}, \tag{43}$$

where $\mathbf{B} = [\mathbf{A}]^{-1}$ with

$$\mathbf{A} = \mathbb{E}\left\{\mathbf{H}^H(\Theta_d)\mathbf{H}(\Theta_d)|\hat{\Theta}_d\right\}. \tag{44}$$

The derivation processes of the above matrices $\mathbf{A}$ and $\mathbf{P}_d$ are similar to that of $\mathbf{R}_d$ in Appendix A. Here, we omit their derivation process.

B. UNKNOWN DIRECTIONS OF EAVESDROPPERS

When the directions of eavesdroppers are unavailable, eavesdroppers may be located in any direction of the remaining angle region excluding all main beams of all desired directions. Thus, the direction angle region for the eavesdroppers is the integral interval defined as

$$\hat{\Omega}_e \triangleq [0, \pi] \setminus \bigcup_{m=1}^{M} \hat{\Omega}_d^m, \tag{45}$$

where

$$\hat{\Omega}_d^m = \left[\hat{\theta}_d^m - \frac{\theta_{BW}}{2}, \ \hat{\theta}_d^m + \frac{\theta_{BW}}{2}\right]. \tag{46}$$

Based on the above definition and leakage idea [16], [17], in the scenario without the eavesdroppers direction knowledge, the conditional SLNR is given by

$$\text{CSLNR}_2(\mathbf{v}_d) = \frac{\beta_1^2 P_s \text{tr}\left\{\mathbf{v}_d^H \mathbf{R}_d \mathbf{v}_d\right\}}{\text{tr}\left\{\mathbf{v}_d^H \left[\beta_1^2 P_s \bar{\mathbf{R}}_e + \sigma_d^2 \mathbf{I}_N\right] \mathbf{v}_d\right\}}, \tag{47}$$

where

$$\bar{\mathbf{R}}_e = \oint_{\hat{\Omega}_e} \mathbf{h}(\theta)\mathbf{h}^H(\theta) d\theta, \tag{48}$$

which can be easily evaluated and computed by Matlab. Maximizing (47) forms the following optimization

$$\max_{\mathbf{v}_d} \ \text{CSLNR}_2(\mathbf{v}_d) \quad \text{subject to } \mathbf{v}_d^H \mathbf{v}_d = 1, \tag{49}$$

which yields the optimal $\hat{\mathbf{v}}_d$ as the eigenvector corresponding to the largest eigen-value of the matrix

$$\left[\oint_{\hat{\Omega}_e} \mathbf{h}(\theta)\mathbf{h}^H(\theta) d\theta + \frac{\sigma_d^2}{\beta_1^2 P_s} \mathbf{I}_N\right]^{-1} \mathbf{R}_d. \tag{50}$$

Similar to the previous subsection, the corresponding projection matrix $\mathbf{P}_{AN}$ has the same form as in (42).

V. SIMULATION RESULTS AND ANALYSIS

To evaluate the performance of the proposed scheme, in our simulation, the baseband system parameters are chosen as: quadrature phase shift keying (QPSK), $L_t = \lambda/2$, $N = 16$, $M = 2$, $K = 2$, $\beta_1^2 = 0.9$, $\Theta_d = \{45°, 135°\}$, $\Theta_e = \{60°, 120°\}$, $\sigma_w^2 = \sigma_d^2 = \sigma_e^2$, $K_c = 0.95$, and $\Delta\theta_{d,\max} = \Delta\theta_{e,\max} = 6°$. Below, the orthogonal projection method in [12] is used as a performance reference.

Fig. 2 illustrates the curves of BER versus direction angle of the proposed Max-SLNR plus Max-SANR method in Section III for the perfect desired and eavesdropper direction knowledge. From Fig. 2, it can be observed that at two desired directions 45° and 135°, the corresponding BER performance is slightly better than that of the orthogonal projection (OP) method in [12]. When the direction angle varies from 45° to 135°, the proposed method indicates a smaller fluctuation on BER compared to the OP in [12].

Fig. 3 plots the curves of BER versus direction angle of the proposed method in Section III for the perfect desired and unknown eavesdropper direction knowledge. From Fig. 3, we find that at two desired directions 45° and 135°, the corresponding BER performance is still slightly better than that of the OP method in [12]. Similar to Fig. 2, as the direction angle changes between the two desired directions, the BER ripples of the proposed method is smaller than those of the orthogonal projection method in [12]. In other words, the former is more stable than the latter during the angle interval.

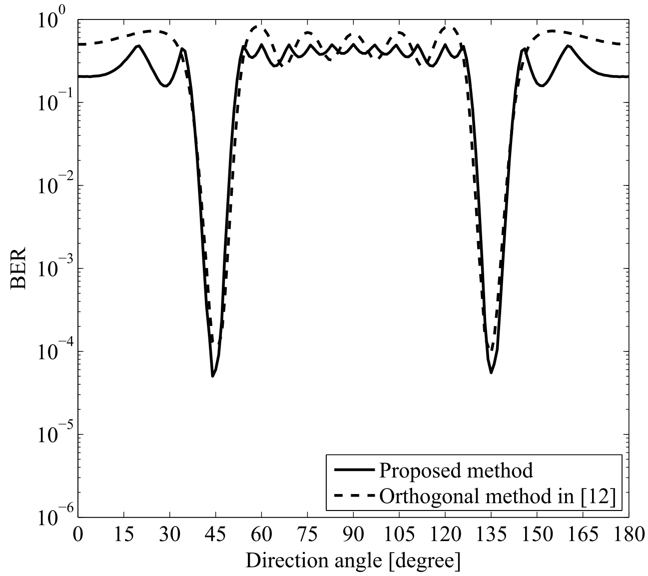

FIGURE 2. Curves of BER versus direction angle of the proposed method in Section III with perfect desired and undesired direction knowledge(SNR = 15dB).

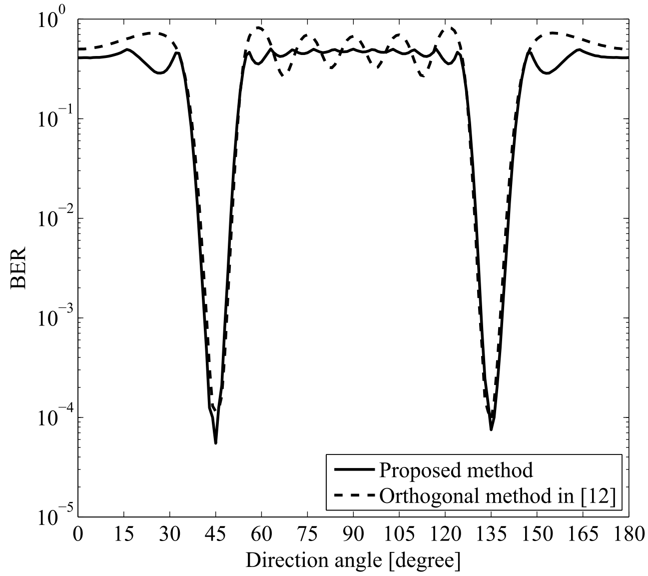

FIGURE 3. Curves of BER versus direction angle of the proposed method in Section III with perfect desired and unknown undesired direction knowledge(SNR = 15dB).

For the imperfect desired and undesired direction knowledge, Fig. 4 plots the curves of BER versus direction angle of the proposed robust Max-SLNR plus Max-SANR method in Section IV. From Fig. 4, it is noted that at the two desired direction angles of 45° and 135°, the corresponding BER performance is much better than that of the OP method in [12]. More importantly, outside two main lobes corresponding to the two desired directions, the BER is greater than 3×10^{-1}. Thus, it is very difficult for eavesdroppers to correctly intercept the transmit useful symbols outside two main lobes corresponding to the two desired directions .

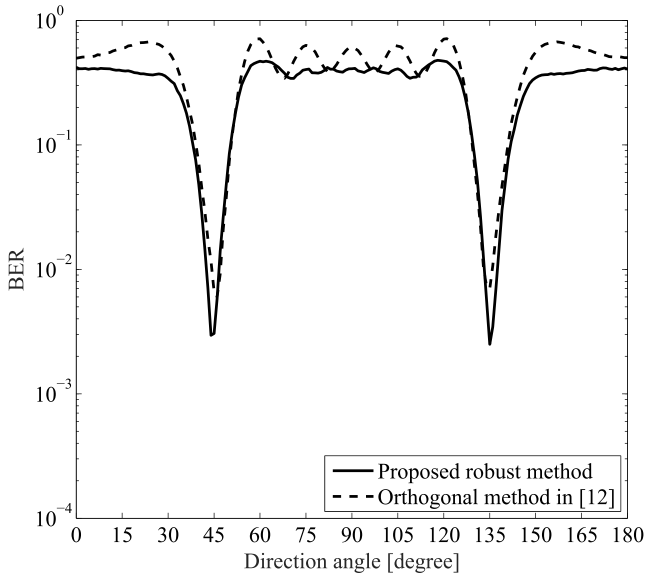

FIGURE 4. Curves of BER versus direction angle of the proposed robust method in Section IV with imperfect desired and undesired direction knowledge (SNR = 15dB).

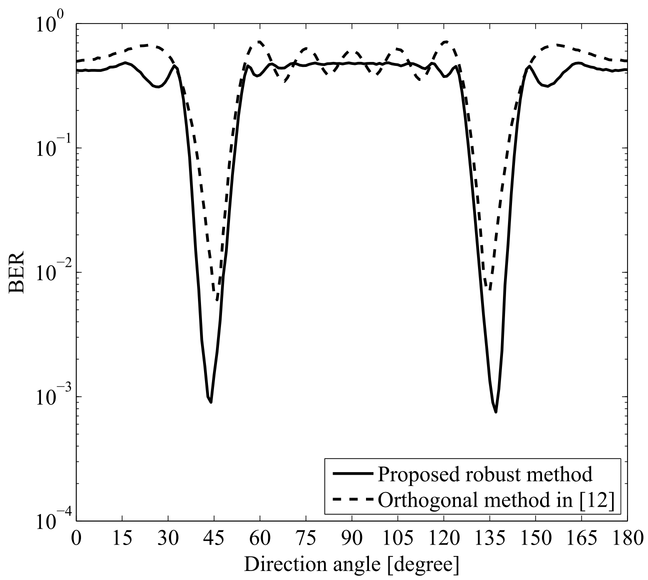

FIGURE 5. Curves of BER versus direction angle of the proposed robust method in Section IV with imperfect desired and unknown undesired direction knowledge(SNR = 15dB).

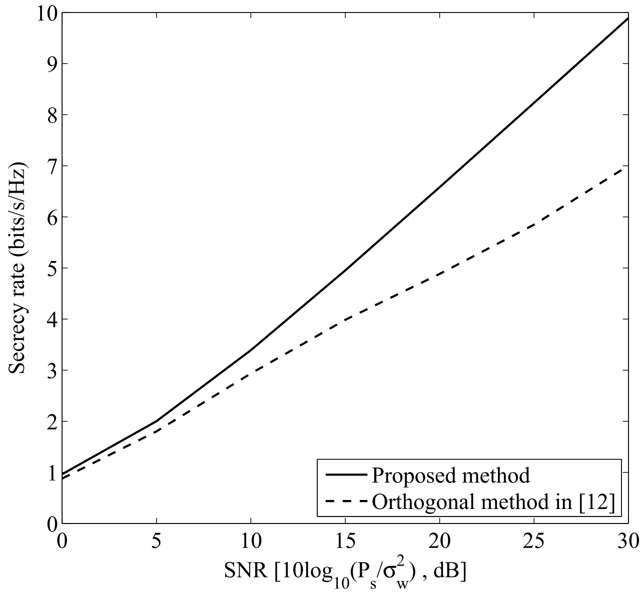

FIGURE 6. Curves of secrecy-rate versus SNR of the proposed method in Section III with perfect desired and undesired eavesdropper direction knowledge.

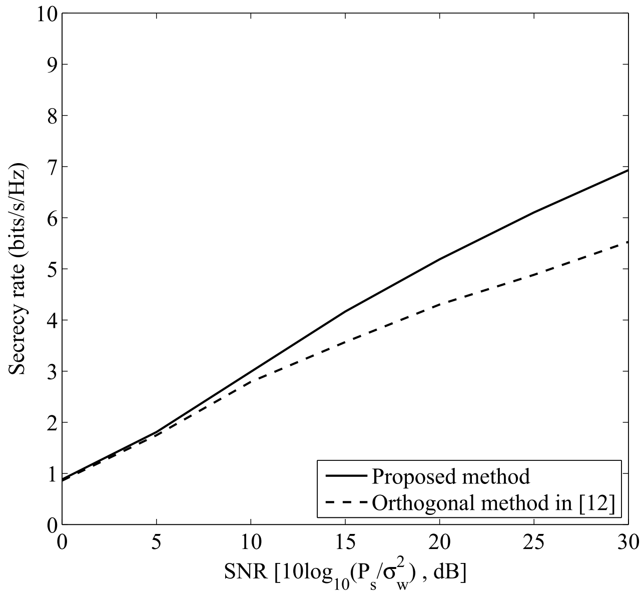

FIGURE 7. Curves of secrecy-rate versus SNR of the proposed method in Section III with perfect desired and unknown eavesdropper direction knowledge.

In the absence of the eavesdropper direction knowledge, Fig. 5 shows the curves of BER versus direction angle of the propose robust method with imperfect desired direction knowledge. At the two desired direction angles of 45° and 135°, the corresponding BER performance is about one order of magnitude, better than that of the OP method in [12].

In what follows, we will evaluate the performance of the proposed method from secrecy rate (SR) aspect. Fig. 6 demonstrates the curves of SR versus SNR of the proposed method in Section III with perfect desired and eavesdropper direction knowledge. From Fig. 6, it is noted that the SR performance of the proposed method is much better than those of the OP method in [12] as SNR increases. As SNR grows, the performance gap between the proposed scheme and OP becomes large. This guarantees that the confidential message is not easy to be intercepted by eavesdroppers.

In the presence of perfect desired and unknown eavesdropper direction knowledge, Fig. 7 plots the curves of SR versus SNR of the proposed method. From Fig. 7, it is obvious that the corresponding SR performance is substantially better than that of the OP method in [12]. The rate improvement of the

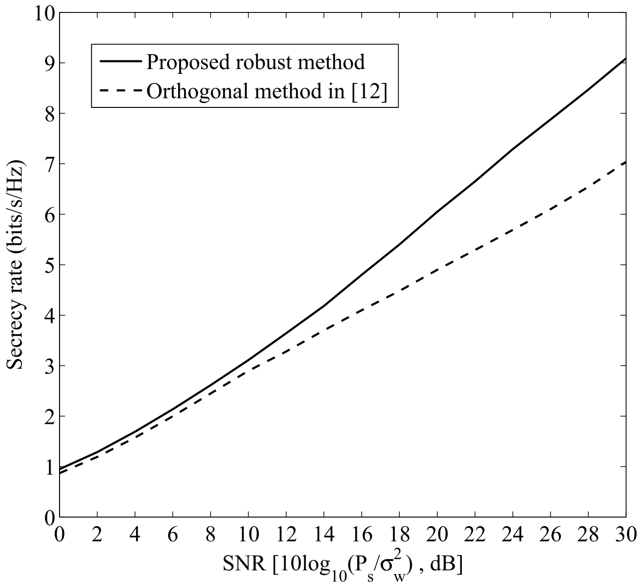

FIGURE 8. Curves of secrecy-rate versus SNR of the proposed robust method in Section IV with imperfect desired and undesired eavesdropper direction knowledge.

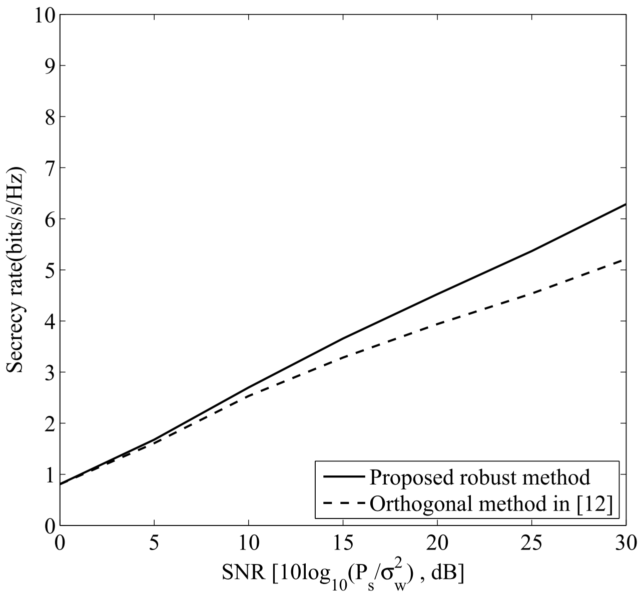

FIGURE 9. Curves of secrecy-rate versus SNR of the proposed robust method in Section IV with imperfect desired and unknown eavesdroppers direction knowledge.

proposed scheme over the OP method in [12] grows with the increase in the value of SNR.

Fig. 8 shows the curves of SR versus SNR of the proposed robust method in Section IV with fixed desired directions. From Fig. 8, for almost the entire SNR region, the secrecy-rate performance of the proposed robust method performs much better than that of the OP method in [12]. In particular, in the high SNR region, the proposed robust scheme makes a substantial improvement over OP in performance. For example, at SNR=30dB, the proposed robust method shows an about 30 percent SR improvement over the OP. This result is attractive. Considering that secrecy-rate is a measure of secrecy, and a larger secrecy rate means a higher security, consequently, we conclude that the proposed robust method is more secure than OP.

Fig. 9 shows the curves of SR versus SNR of the proposed robust method in Section IV with imperfect desired and unknown eavesdropper direction knowledge. In the medium and high SNR regions, the secrecy rate of the proposed robust method is much higher than those of the OP method in [12]. However, in the low SNR region, the proposed robust method performs slightly better than the OP.

VI. CONCLUSIONS

In this paper, we make an investigation of synthesis methods of MBDM in broadcasting systems. We firstly propose a hybrid beamforming scheme: Max-SLNR and Max-SANR in the case of perfect desired direction knowledge, where Max-SLNR and Max-SANR are used to designed the beamforming vector of the useful message and the projection matrix for AN, respectively. The proposed method shows a slightly better BER performance over existing OP method in the desired directions and at the same time keeps a low BER performance along eavesdropper directions. In the imperfect desired direction knowledge scenario, we propose a robust Max-SLNR and Max-SANR method by utilizing conditional expectation of the Grammian matrices for the desired and eavesdroppers directions, which achieves a substantial secrecy-rate performance improvement over the existing OP method in the medium and high SNR regions and indicates an obvious BER improvement over the existing OP method along the desired direction for a given SNR. For example, given SNR=30dB, compared to the OP, the proposed robust method can achieve not only about thirty percent secrecy-rate improvement but also one order of magnitude BER enhancement along the desired directions in the scenario of imperfect desired and undesired eavesdropper direction knowledge. The proposed scheme can be applied to the future wireless communication for robust secure information transmissions like broadcasting systems, mobile communications, and satellite broadcasting.

APPENDIX
DERIVATION OF MATRIX $\mathbf{R}_d$

Proof: Matrix $\mathbf{R}_d$ in (37) is expanded as a sum of M terms

$$\mathbf{R}_d = \sum_{m=1}^{M} \mathbb{E}\left[\mathbf{h}(\theta_d^m)\mathbf{h}^H(\theta_d^m)|\hat{\Theta}_d\right]. \tag{51}$$

Observing the above sum, it is clear that all terms inside the summation have a similar expression form. If we derive the expression for the mth term, then we readily write the expressions for other terms. Let us use Γ_m to represent the mth term of the above sum as follows

$$\Gamma_m = \mathbb{E}\left\{\mathbf{h}(\theta_d^m)\mathbf{h}^H(\theta_d^m)|\hat{\theta}_d^m\right\} \tag{52}$$

where Γ_m is an $N \times N$ matrix. Its (p, q) entry is given by the following expression

$$\begin{aligned}
&\Gamma_m(p,q)\\
&= \mathbb{E}\left\{\mathbf{h}_p(\theta_d^m)\mathbf{h}_q^H(\theta_d^m)|\hat{\theta}_d^m\right\}\\
&= \int_{-\Delta\theta_{d,\max}}^{\Delta\theta_{d,\max}} \frac{1}{\sqrt{N}} e^{\frac{-j2\pi(p-(N+1)/2)L_t\cos(\hat{\theta}_d^m-\Delta\theta_d^m)}{\lambda}}\\
&\quad \cdot \frac{1}{\sqrt{N}} e^{\frac{j2\pi(q-(N+1)/2)L_t\cos(\hat{\theta}_d^m-\Delta\theta_d^m)}{\lambda}} f(\Delta\theta_d^m)d(\Delta\theta_d^m)\\
&= \frac{1}{N}\int_{-\Delta\theta_{d,\max}}^{\Delta\theta_{d,\max}} e^{\frac{j2\pi(q-p)L_t\cos\hat{\theta}_d^m\cos\Delta\theta_d^m+j2\pi(q-p)L_t\sin\hat{\theta}_d^m\sin\Delta\theta_d^m}{\lambda}}\cdot\\
&f(\Delta\theta_d^m)d(\Delta\theta_d^m)\\
&= \frac{2}{N}\cdot\int_0^{\Delta\theta_{d,\max}}\cos\\
&\quad\times\frac{\gamma(q-p)\left(\cos\hat{\theta}_d^m\cos\Delta\theta_d^m+\sin\hat{\theta}_d^m\sin\Delta\theta_d^m\right)}{\lambda}\\
&\cdot f(\Delta\theta_d^m)d(\Delta\theta_d^m),
\end{aligned} \tag{53}$$

where $\gamma = 2\pi L_t$. In order to simplify (53), let us define

$$a_{pq} = \frac{\gamma(q-p)\cos\hat{\theta}_d^m}{\lambda}, \tag{54}$$

$$b_{pq} = \frac{\gamma(q-p)\sin\hat{\theta}_d^m}{\lambda}, \tag{55}$$

and

$$x = \frac{\Delta\theta_d^m}{c}, \tag{56}$$

with $c = \Delta\theta_{d,\max}/\pi$. Substituting (33), (54), (55) and (56) into (53) yields

$$\begin{aligned}
\Gamma_m(p,q) = &\frac{2c}{NK_d\sigma_{\theta_d}\sqrt{2\pi}}\cdot\int_0^{\pi}\cos\left[a_{pq}\cos(cx)+b_{pq}\sin(cx)\right]\\
&\times\exp\left[-\frac{(cx)^2}{2\sigma_{\theta_d}^2}\right]dx.
\end{aligned} \tag{57}$$

Under the condition $\Delta\theta_{d,\max} \ll \pi$, we have the following approximation

$$\sin(cx) \approx cx = \Delta\theta_d^i \leq \Delta\theta_{d,\max} \approx 0, \quad \cos(cx) \approx 1. \tag{58}$$

Then, the integration expression (53) has the following approximate form

$$\begin{aligned}
&\Gamma_m(p,q)\\
&\approx \frac{2}{N}\cdot\int_0^{\Delta\theta_{d,\max}}\cos\frac{2\pi(q-p)L_t\cos\hat{\theta}_d^m}{\lambda}\cdot f(\Delta\theta_d^m)d(\Delta\theta_d^m).
\end{aligned} \tag{59}$$

Similar to (57), by using the variable substitutions in (54), (54), and (56), we have

$$\Gamma_m(p,q) \approx \frac{2c\cos\left(a_{pq}\right)}{NK_d\sigma_{\theta_d}\sqrt{2\pi}}\cdot\int_0^{\pi}\exp\left[-\frac{(cx)^2}{2\sigma_{\theta_d}^2}\right]dx. \tag{60}$$

If we define

$$t = \frac{cx}{\sigma_{\theta_d}}, \tag{61}$$

then the above integral $\Gamma_i(p, q)$ is further simplified as

$$\begin{aligned}
\Gamma_m(p,q) &\approx \eta_{pq}\int_0^g \frac{1}{\sqrt{2\pi}}\exp\left(-\frac{t^2}{2}\right)dt\\
&= \eta_{pq}\left[\frac{1}{2}-\int_g^{+\infty}\frac{1}{\sqrt{2\pi}}\exp\left(-\frac{t^2}{2}\right)dt\right]\\
&= \eta_{pq}\left[\frac{1}{2}-\mathrm{Q}(g)\right],
\end{aligned} \tag{62}$$

where

$$\eta_{pq} = \frac{2\cos\left(a_{pq}\right)}{NK_d}, \quad g = \frac{\Delta\theta_{d,\max}}{\sigma_{\theta_d}}, \tag{63}$$

and the Q function is defined by [23]

$$\mathrm{Q}(z) = \int_z^{+\infty}\frac{1}{\sqrt{2\pi}}\exp\left(-\frac{t^2}{2}\right)dt. \tag{64}$$

In summary,

$$\mathbf{R}_d(p,q) = \sum_{m=1}^{M}\Gamma_m(p,q). \tag{65}$$

This completes the derivation of matrix $\mathbf{R}_d$. In the same manner, we can get the exact and approximate expressions of matrices $\mathbf{R}_e$ in (39), $\mathbf{A}$ in (44), and $\mathbf{P}_d$ in (43). Due to the limited paper length and similarity of the analysis, we omit their derivation processes here. ■

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