IRS-aided Predictable High-Mobility Vehicular Communication with Doppler Effect Mitigation

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***Abstract*—In this paper, we propose a novel anti-Doppler spread technique in IRS-aided predictable high-mobility vehicular communication. The predictable information of vehicle (e.g., speed, trajectory, and timetable) can be used to design a sub-optimal phase shift set by solving a multi-objective problem, which can make IRS capable of maximizing instantaneous signal-to-noise ratio (SNR), minimizing Doppler spread, and keeping delay spread to a relatively low range simultaneously. The technique we propose in this paper is more cost-effective, more practical, and of higher energy efficiency since the phase shift set in every element of the IRS for one vehicle pass can be designed and stored in advance, instead of processing in real-time. Simulation results show the effectiveness of the propose phase shift set as compared to benchmark schemes.**

***Index Terms*—intelligent reflecting surfaces, Doppler effect, high mobility vehicular communication**

I. Introduction

Intelligent reflecting surfaces(IRSs), also known as reconfigurable intelligent surfaces(RISs), and Large intelligent surfaces(LISs), has emerged as an innovative technique that enables to control the propagation electromagnetic environment [1], [2]. specifically, IRS is a planar surface made up of an array of sub-wavelength passive reflecting elements, each of which can cause a tunable amplitude and/or phase shift on the incident signal in real-time independently [3], thereby the reflected electromagnetic wave can be appropriately reoriented towards the desired direction to improve the performance of transmission.

The vehicle with high mobility has severe Doppler effect and results in superimposed fast fading channel between the transmitter and the receiver [4]. Moreover, considering the occlusion of multiple buildings in urban areas, the channel is more complicated and unpredictable than the conventional low-mobility case. Therefore, the spectral efficiency and outage probability of the whole system would be greatly degraded. Recently, due to its ability of controlling the complicated propagation environment, IRS has attracted widespread attentions in vehicular communication to enhance the quality of transmission [5]–[7]. Furthermore, the authors in [8] show that by applying the real-time tunable IRS, the rapid variations in the received signal intensity due to the Doppler effect can be effectively reduced. [9] proposes a new and efficient channel estimation scheme by aiding an IRS on a high-speed moving vehicle to assist the communication between its passenger and a static base station (BS), and the simulation results show that IRS can not only maximize the beamforming gain but also convert the combined channel of the system from fast to slow fading. In [10], IRS channel estimation is investigated with the consideration of Doppler effect. Besides, the authors in [11] introduce a continuous time propagation model of IRS under predictable receiver mobility for LEO satellite communications.

The investigations of the works in [8]–[10] assume that moving positions and timetable are unknown during the whole transmission, thereby the IRS has to use complicated techniques to estimate channel and to obtain the phase shift set in real-time. However, in many practical scenarios, we can assume the mobility information is known a priori. For example, for high-speed train, its speed, trajectory and rail timetable are already known in advance. In other words, the position is predictable all the times.

In this paper, we exploit the predictable information of high mobility vehicle to optimize the phase shift of IRS in order to maximize the instantaneous SNR of the received signal for the vehicle, minimize Doppler spread, and keep delay spread to a very low level. The satellite communication propagation model in [11], however, is not suitable for vehicular communication. This is because compares to received power, instantaneous SNR is more important in vehicular communication. Therefore, we are considering a new framework that is fit for practical vehicular communication. Specially, we first adopt the continuous time propagation model in [11], and then transform three single-objective optimization problems into a multi-objective problem in a novel high mobility vehicle scenario. Consequently, we obtain a sub-optimal phase shift set, which can maximize the instantaneous SNR, totally eliminate the Doppler spread, and with almost no increase in delay spread simultaneously. Moreover, different from [8]–[10], which need channel estimation to mitigate Doppler effect and multipath fading, our scheme in this paper does not need channel estimation. Thus, it is more practical and cost-effective. Simulation results show that by applying our scheme, an IRS with 128×128 elements can bring instantaneous SNR gain from 0.67 dB to 4.21 dB in a typical scenario. Also, the Doppler spread and the delay spread can be eliminated to zero and kept to 10^{-7} s order of magnitude, respectively. To the best of our knowledge, this is the first work on Doppler mitigation for IRS-aided predictable high-mobility vehicular communication system.

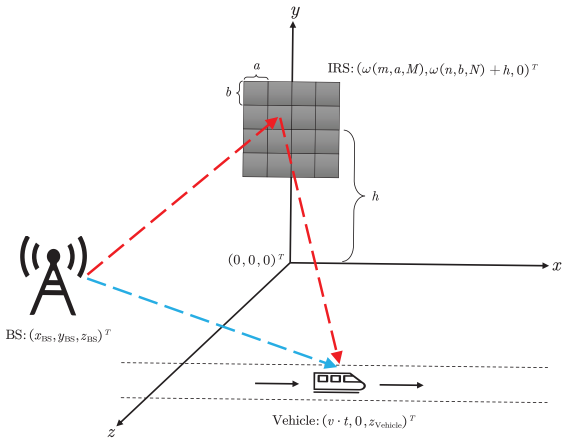

Fig. 1. IRS-aided high-mobility vehicle communication system.

Notation: Lower-case boldface letters (e.g., $\boldsymbol{p}$) denote the column vector. Upper-case calligraphic letters (e.g., $\mathcal{T}$) denote the phase shift set. The functions $(x \bmod y)$, $\lceil \cdot \rceil$ and $\lfloor \cdot \rfloor$ refer to the remainder of the division of x by y, the ceiling, and the floor of their argument, respectively. Further, $\mathbb{Z}$ is the set of integer, $|\cdot|$ is the absolute value, and $\|\cdot\|$ denotes the l_2 norm.

II. System Model

We consider a downlink IRS-aided high mobility vehicular communication system (e.g., high-speed train), where both of the fixed BS and the moving vehicle are equipped with a single isotropic antenna. As shown in Fig. 1, the IRS is a uniform planar array (UPA) composed of $M \times N$ elements which are placed on a rectangular grid spaced a and b apart in the $x-y$ plane in the three-dimensional (3D) Cartesian coordinate system, and the height of the IRS is h. Moreover, the vehicle moves at a very high speed of v m/s, the trajectory of the vehicle is straight along the positive x-axis direction, and we know when it arrives, using the schedule information.

A. Geometry of the IRS in the 3D Cartesian Coordinate System

Assume the geometric center of the IRS in Fig. 1 is $(0, h, 0)$. The coordinate column vector of the (m, n)th element is

$$\boldsymbol{p}_{m,n} = \Big[\omega(m, a, M), \omega(n, b, N) + h, 0\Big]^T, \quad (1)$$

where $\omega(m, a, M) = a(m - 0.5((M+1) \bmod 2))$, $m \in \Omega(M)$ and $n \in \Omega(N)$ with [11]

$$\Omega(M) = \left\{((M+1) \bmod 2) - \left\lfloor \frac{M}{2} \right\rfloor, \ldots, \left\lfloor \frac{M}{2} \right\rfloor\right\}, \quad (2)$$

and $\Omega(N)$ is obtained in the same way as $\Omega(M)$. Fig. 2 gives a simple example for IRS element notations when $M = N = 2$, $a = b = 1m$ and $h = 20m$.

B. Amplitude Gains and Time Delays for Direct Path

We consider both channels over the direct path (i.e., from the BS to the vehicle) and the IRS path (i.e., from the BS to the IRS and from the IRS to the vehicle) are line-of-sight (LOS), and ignore hardware impairment. Therefore, if the transmitter locates in $\boldsymbol{p}_{\rm T}$, the amplitude gain that can be observed at an arbitrary position $\boldsymbol{p}$ is [12]

$$A = \frac{\sqrt{G_{\rm T}(\theta_{\rm T}, \varphi_{\rm T})}}{\sqrt{4\pi}\|\boldsymbol{p} - \boldsymbol{p}_{\rm T}\|}, \quad (3)$$

where the angles $\theta_{\rm T} = \theta(\boldsymbol{p}_{\rm T}, \boldsymbol{p}) \in [0, \pi]$ and $\varphi_{\rm T} = \varphi(\boldsymbol{p}_{\rm T}, \boldsymbol{p}) \in [0, 2\pi]$ are polar and azimuth angles, respectively. $G_{\rm T}(\theta_{\rm T}, \varphi_{\rm T})$ refers to the antenna gain of the transmitter in the direction of observation point, and $\|\boldsymbol{p} - \boldsymbol{p}_{\rm T}\|$ is the Euclidean distance between transmitter and observation point.

Based on (3), if the instantaneous time is t, we obtain the direct path amplitude gain A_0 as

$$A_0 = \frac{\lambda_c \sqrt{G_{\rm T}^{\rm R}(t) G_{\rm R}^{\rm T}(t)}}{4\pi\|\boldsymbol{p}_{\rm R}(t) - \boldsymbol{p}_{\rm T}\|}, \quad (4)$$

where $G_{\rm T}^{\rm R}(t)$ is the transmitter antenna gain when the observation point is receiver antenna, $G_{\rm R}^{\rm T}(t)$ is the receiver antenna gain when the observation point is transmitter antenna. λ_c denotes the wavelength of the signal, and $\|\boldsymbol{p}_{\rm R}(t) - \boldsymbol{p}_{\rm T}\|$ is the Euclidean distance between transmitter and receiver at time t.

Accordingly, the signal that is received by the receiver antenna at time t is

$$S(t) = \frac{\lambda_c \sqrt{G_{\rm T}^{\rm R}(t) G_{\rm R}^{\rm T}(t)}}{4\pi\|\boldsymbol{p}_{\rm R}(t) - \boldsymbol{p}_{\rm T}\|} x_p\Big(t - \tau_0(t)\Big), \quad (5)$$

where $x_p(t)$ is a passband signal and $\tau_0(t)$ is the time delay from the transmitter to the receiver in the direct path with

$$\tau_0(t) = \frac{\|\boldsymbol{p}_{\rm R}(t) - \boldsymbol{p}_{\rm T}\|}{c_0}, \quad (6)$$

where c_0 is the speed of light. Since the transmitter and receiver antennas are all isotropic, then $G_{\rm T}^{\rm R}(t) = G_{\rm R}^{\rm T}(t) = 1$. Thus, (5) can be rewritten as

$$S_1(t) = \frac{\lambda_c}{4\pi\|\boldsymbol{p}_{\rm R}(t) - \boldsymbol{p}_{\rm T}\|} x_p\Big(t - \tau_0(t)\Big). \quad (7)$$

C. Amplitude Gains and Time Delays for IRS Path

Consider the observation point is at the (m, n)th IRS element at first. Similar to (5), we assume the energy of the signal is totally reflected by the surface, and the controllable phase shift by the (m, n)th IRS element at time t is $\phi_{m,n}(t) > 0$. If $G_{\rm T}^{m,n}$ is the transmitter antenna gain when the observation point is IRS, $G_{m,n}^{\rm T}$ is the IRS antenna gain when the observation point is transmitter antenna, then the reflected signal is

$$S_{\rm Reflected}(t) = \frac{\lambda_c \sqrt{G_{\rm T}^{m,n} G_{m,n}^{\rm T}}}{4\pi\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\rm T}\|} x_p\Big(t - \tau_{\rm T}^{m,n} - \frac{\phi_{m,n}(t)}{2\pi f_c}\Big), \quad (8)$$

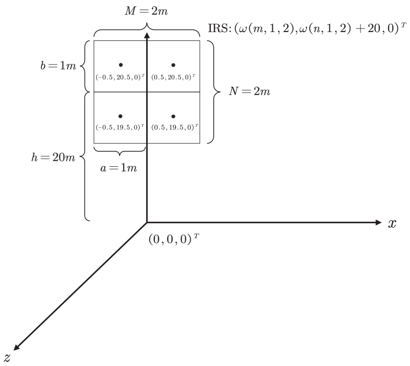

Fig. 2. A simple example for IRS element notations when $M = N = 2$, $a = b = 1m$ and $h = 20m$.

where f_c is carrier frequency, $\tau_{\mathrm{T}}^{m,n} = \frac{\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{T}}\|}{c_0}$ denotes the time delay in the transmitter to the (m,n)th IRS element path, and $\frac{\phi_{m,n}(t)}{2\pi f_c}$ is the time delay that caused by the controllable phase shift $\phi_{m,n}(t)$ of IRS at time t.

If we add the receive antenna to the system, then the signal from the IRS to the receiver in the IRS path is

$$S_{\mathrm{IRS}}(t) = \frac{\lambda_c \sqrt{G_{\mathrm{R}}^{m,n}(t) G_{m,n}^{\mathrm{R}}(t)}}{4\pi \|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|} x_p\left(t - \tau_{m,n}^{\mathrm{R}}(t)\right), \tag{9}$$

where $\tau_{m,n}^{\mathrm{R}}(t) = \frac{\|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|}{c_0}$ is the time delay from the (m,n)th IRS element to the receiver at time t, $G_{\mathrm{R}}^{m,n}(t)$ is the receiver antenna gain when the observation point is the IRS, and $G_{m,n}^{\mathrm{R}}(t)$ is the (m,n)th IRS element gain when the observation point is the receiver antenna.

Based on (8) and (9), and assume $G_{\mathrm{T}}^{m,n} = G_{\mathrm{R}}^{m,n}(t) = 1$, we then obtain the amplitude gain $A_{m,n}$ and time delay $\tau_{m,n}(t)$ for the (m,n)th IRS element path as

$$A_{m,n} = \frac{\lambda_c^2 \sqrt{G_{m,n}^{\mathrm{T}} G_{m,n}^{\mathrm{R}}(t)}}{16\pi^2 \|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{T}}\| \|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|}, \tag{10}$$

$$\tau_{m,n}(t) = \frac{\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{T}}\|}{c_0} + \frac{\|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|}{c_0}. \tag{11}$$

It is worth noting that in this paper we assume the vehicle is in the geometric far-field, and the IRS is with the planar antenna gain pattern, then $G_{m,n}^{\mathrm{T}} = \frac{4\pi}{\lambda_c^2} ab\cos(\theta_{\mathrm{T}})$ and $G_{m,n}^{\mathrm{R}}(t) = \frac{4\pi}{\lambda_c^2} ab\cos(\theta_{\mathrm{R}}(t))$ where $\theta_{\mathrm{T}} = \arccos(\frac{z_{m,n} - z_{\mathrm{T}}}{\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{T}}\|})$ and $\theta_{\mathrm{R}}(t) = \arccos(\frac{z_{m,n} - z_{\mathrm{R}}}{\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{R}}(t)\|})$ [13].

Based on the system model, $z_{m,n} = 0$, thereby (10) can be rewritten as

$$A_{m,n} = \frac{ab\sqrt{z_{\mathrm{T}} z_{\mathrm{R}}}}{4\pi (\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_T\| \|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|)^{\frac{3}{2}}}, \tag{12}$$

where z_{T} and z_{R} are the z-axis coordinates for the transmitter and the receiver, respectively. Consequently, the corresponding complex baseband received signal is

$$\begin{aligned} y(t) = {} & A_0(t) e^{-j2\pi f_c \tau_0(t)} x(t - \tau_0(t)) \\ & + \sum_{m,n} A_{m,n}(t) e^{-j2\pi f_c \tau_{m,n}(t) - j\phi_{m,n}(t)} \\ & x\left(t - \tau_{m,n}(t) - \frac{\phi_{m,n}(t)}{2\pi f_c}\right) + n(t), \end{aligned} \tag{13}$$

where $n(t) \sim \mathcal{CN}(0, \sigma^2)$ denotes the receiver noise.

III. Problem Formulation

Assuming that the transmitter sends a constant signal $x(t)$ with power P_{T}, the instantaneous received signal power $P_{\text{receiver}}(t)$ can be calculated, using (13), as

$$\begin{aligned} P_{\text{received}}(t) = P_{\mathrm{T}} \Bigg| & A_0(t) e^{-j2\pi f_c \tau_0(t)} + \\ & \sum_{m,n} A_{m,n}(t) e^{-j2\pi f_c \tau_{m,n}(t) - j\phi_{m,n}(t)} \Bigg|^2. \end{aligned} \tag{14}$$

Then *the instantaneous signal-to-noise ratio (SNR)* of the received signal $y(t)$ is

$$\mathrm{SNR}(t) = \frac{P_{\text{received}}(t)}{\sigma^2}, \tag{15}$$

where B is the bandwidth.

The maximum difference in instantaneous frequency over all essential propagation paths is *Doppler Spread* [4]. In this paper, the Doppler spread is the largest frequency difference over the two paths. First, at time t, the Doppler spread between the direct path and the IRS is

$$D_{s,0}(t) = f_c \max_{m,n} \left| \frac{d}{dt}\left(\tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c}\right) - \frac{d}{dt}\tau_0(t) \right|. \tag{16}$$

Similarly, the Doppler spread of the IRS itself at time t is

$$\begin{aligned} D_{s,\mathrm{IRS}}(t) = f_c \max_{m,n,m',n'} \Bigg| & \frac{d}{dt}\left(\tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c}\right) - \\ & \frac{d}{dt}\left(\tau_{m',n'}(t) + \frac{\phi_{m',n'}(t)}{2\pi f_c}\right) \Bigg|. \end{aligned} \tag{17}$$

Therefore, we have the Doppler spread over all paths at time t as

$$D_s(t) = \max\{D_{s,0}(t), D_{s,\mathrm{IRS}}(t)\}. \tag{18}$$

The maximum difference in propagation time over all substantial propagation paths is *Delay Spread* [4]. Like Doppler spread, we also define the delay spread as the largest time difference over the two paths. The delay spread between the direct path and the IRS at time t is

$$T_{d,0}(t) = \max_{m,n} \left\{ \tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c} \right\} - \tau_0(t), \tag{19}$$

and the delay spread across the IRS at time t is

$$T_{d,\text{IRS}}(t) = \max_{m,n}\left\{\tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c}\right\} - \min_{m,n}\left\{\tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c}\right\}. \tag{20}$$

Then we have the delay spread over all paths at time t as

$$T_d(t) = \max\{T_{d,0}(t), T_{d,\text{IRS}}(t)\}. \tag{21}$$

Using (15), (18) and (21), we aim to 1) maximize the instantaneous signal-to-noise ratio $\text{SNR}(t)$ at the receiver, 2) minimize Doppler spread $D_s(t)$ and 3) keep delay spread $T_d(t)$ to a low range. In other words, at time t, we want to find a *Pareto optimal solution* [14] for the multi-objective problem below

$$\phi_{m,n}^{opt}(t) = \max_{\forall m,n:\phi_{m,n}(t)}\left[\text{SNR}(t) \quad - D_\text{s}(t) \quad - T_\text{d}(t)\right]. \tag{22}$$

IV. Pareto Optimization for Phase Shift of IRS

A. Basic Concept

In this section, we want to obtain a Pareto optimal solution $\phi_{m,n}^{opt}(t)$ for (22). A point is considered to be the Pareto optimal solution if it achieves an objective vector where no component can be improved without worsening at least one other objective. Such a point is called a Pareto optimal solution and the set of all such vectors is the Pareto optimal solution set [14]. For (22), the only solution resource set available is the tunable phase shifts by the IRS elements, which can be modeled by a compact set $\mathcal{X}$. Consider the simplest case, however, if the discrete phase shifts are allowed to be 0 and π (i.e., 1bit), then we still have a very large resource set $\mathcal{X} \in \mathbb{R}^{2^{MN}}$, where M and N denote the columns and rows of the IRS elements, respectively. It is easy to find that if the IRS has a large number of elements (e.g., 128×128) and discrete phase shifts (e.g., 3bits), the solution resource set $\mathcal{X}$ will be extremely large, which makes finding the appropriate solution phase vector $\phi_{m,n}^{opt}(t) \in \mathcal{X}$ almost impossible.

Fortunately, the conclusion above is based on the assumption that the three objective problems (i.e., $\max\{\text{SNR}(t)\}$, $\min\{D_\text{s}(t)\}$ and $\min\{T_\text{d}(t)\}$) are processed without any predictions and are all not ordered, which can be improved according to our assumption in this paper. The most important objective in (22) is to maximize the instantaneous SNR at time t. Minimizing Doppler and delay spread are the second and third objectives. Therefore, instead of searching for the solution vector in an extremely large solution resource set $\mathcal{X}$, we first obtain the solution vector set $\mathcal{S}(t) \subseteq \mathcal{X}$ that can maximize $\text{SNR}(t)$ at time t. Based on $\mathcal{S}(t)$, we then try to search for a solution set $\mathcal{D}(t) \subseteq \mathcal{S}(t)$ which can minimize $D_\text{s}(t)$. Finally a solution (or solution set) $\mathcal{T}(t) \subseteq \mathcal{D}(t)$ which can minimize $T_\text{d}(t)$ can be selected from $\mathcal{D}(t)$.

B. Pareto Optimization

Consider all sets are continuous, we first obtain the solution set $\mathcal{S}(t)$ that can maximize $\text{SNR}(t)$ at time t by aligning in the phases of the two paths in (14), then we have [3]

$$\phi_{m,n}(t) = 2\pi f_c(\tau_0(t) - \tau_{m,n}(t)) + 2\pi k_{m,n}(t), \tag{23}$$

where $k_{m,n}(t)$ is a piecewise constant function taking integer values, which denotes the additional full carrier signal period delays. It is noteworthy that $k_{m,n}(t)$ does not affect $\text{SNR}(t)$. Therefore, we first obtain the solution phase vector set $\mathcal{S}(t) = \phi_{m,n}(t)$.

Next, we want to find the solution phase vector set $\mathcal{D}(t)$ from $\mathcal{S}(t)$. The set $\mathcal{D}(t)$ aims to minimize the Doppler spread and guarantees the $\text{SNR}(t)$ is also maximized simultaneously. The derivative of (23) is

$$\frac{d}{dt}\phi_{m,n}(t) = 2\pi f_c\frac{d}{dt}(\tau_0(t) - \tau_{m,n}(t)) + 2\pi\frac{d}{dt}k_{m,n}(t). \tag{24}$$

Because of the definition of $k_{m,n}(t)$, it is easy to find that $\frac{d}{dt}k_{m,n}(t) = 0$ except the time t changes by an integer value. However, we also note that $k_{m,n}(t)$ results in a 2π phase shift when the time t changes by an integer value, and it does not affect the instantaneous frequency. Therefore, $\frac{d}{dt}k_{m,n}(t) = 0$, and we then have

$$\frac{d}{dt}\phi_{m,n}(t) = 2\pi f_c\frac{d}{dt}(\tau_0(t) - \tau_{m,n}(t)). \tag{25}$$

Thus, $D_{s,0}(t)$ is minimized to zero due to (25). It is interesting to note that (25) also guarantees the $D_{s,\text{IRS}}(t)$ diminishes to zero [11]. Therefore, $\mathcal{D}(t)$ is the same set as $\mathcal{S}(t)$.

We finally want to search for an optimal solution (or solution set) to minimize the delay spread $T_\text{d}(t)$. From (19)-(21) we can simplify the delay spread $T_\text{d}(t)$ as

$$T_d(t) = T_{d,0}(t) = \max_{m,n}\left\{\tau_{m,n}(t) + \frac{\phi_{m,n}(t)}{2\pi f_c}\right\} - \tau_0(t). \tag{26}$$

From (23) and (26), it is not difficult to find that maximizing $\text{SNR}(t)$ and minimizing $T_d(t)$ cannot be achieved simultaneously. Therefore, the problem (22) has an sub-optimal solution phase vector set instead of a global single optimal solution. Substitute (23) into (26), we have

$$T_d(t) = \max_{m,n}\left\{\frac{k_{m,n}(t)}{f_c}\right\}. \tag{27}$$

Thus, $k_{m,n}(t)$ should be chosen as small as possible. From (6) and (11), it can be seen that $\tau_{m,n}(t) \geq \tau_0(t)$, then $k_{m,n}(t) \geq f_c(\tau_{m,n}(t) - \tau_0(t)) \geq 0$. Therefore, after optimization, we choose $k_{m,n}(t) = \lceil f_c(\tau_{m,n}(t) - \tau_0(t))\rceil$, and the delay spread $T_d(t)$ is

$$T_d(t) = \frac{1}{f_c}\max_{m,n}\lceil f_c(\tau_{m,n}(t) - \tau_0(t))\rceil. \tag{28}$$

Finally, based on (23), the sub-optimal solution phase vector set $\mathcal{T}(t)$ can be selected as

$$\mathcal{T}(t) = 2\pi\Bigg(\Big(f_c(\tau_0(t) - \tau_{m,n}(t))\Big) \bmod 1\Bigg). \quad (29)$$

V. Simulation Results

In this section, numerical examples are provided to validate the effectiveness of the sub-optimal solution phase vector set $\mathcal{T}(t)$. As mentioned in Section II, we consider a three-dimensional (3D) Cartesian coordinate system where an IRS located in $x-y$ plane, $M = N = 128$ and $a = b = \frac{\lambda}{5}$. The height of the IRS is $h = 20$ m, the vehicle moves at a speed of $v = 83.3$ m/s, and the trajectory of the vehicle is straight along the positive x-axis direction. Moreover, the BS is located in (-50 m,100 m, 50 m), the vehicle is located in $(v \cdot t,\ 0,\ 100\text{ m})$ where $t \in \mathbb{Z}$ and the range is from 1 s to 100 s. The start point of the vehicle is (-4000 m, 0, 100 m), the carrier frequency is 2.4 GHz and the bandwidth is 1 MHz. Transmit power P_T is 1 W and σ^2 is 10^{-6} W. We also assume unobstructed view, no atmospheric effects, and all paths are LOS.

A. Instantaneous Signal-to-Noise Ratio (SNR) of the Received Signal for One Vehicle Pass

We compare the IRS with the propose sub-optimal phase shift set $\mathcal{T}(t)$ for one vehicle pass to the cases of 1) without IRS, 2) IRS with random phase shift and 3) isotropic IRS at first. The isotropic IRS case is used as an upper bound that serves as a best case deployment, baseline case is the IRS with random phase shift and lower bound is the transmission without IRS. From Fig. 3, it can be seen that the baseline (i.e., IRS with random phase shift) and the lower bound (i.e., without IRS) have the same performance on the instantaneous SNR, this is because if the IRS with random phase shift, there is no difference between it and normal scatters in the practical environment. However, the instantaneous SNR of the IRS with propose optimal phase shift over the baseline and the lower bound from 0.67 dB to 4.21 dB for one vehicle pass. It is worth noting that in the case of isotropic IRS, we have to recompute the number of elements to make sure the effective antenna area matches the size of the IRS. We also observe that at time $t = 48$, the performances of all kinds of IRS are the best. This is expected since in (12), the minimum value of $\|\boldsymbol{p}_{\mathrm{R}}(t) - \boldsymbol{p}_{m,n}\|$ is achieved at $t = 48$. In other words, if we reduce the distance between the IRS and the vehicle in the z-axis direction, the performance of the whole vehicular communication system on instantaneous SNR will increase. Also, we can decrease $\|\boldsymbol{p}_{m,n} - \boldsymbol{p}_{\mathrm{T}}\|$ to achieve the same goal. However, if the distance is too small, the whole system is in near-field, and (12) does not exist then. Therefore, finding an appropriate location of deployment for IRS is an important and challenging problem.

Fig. 4 shows the performances on instantaneous SNR for the propose scheme when the IRS equips 64×64, 128×128 and 196×196 elements, respectively. Obviously the IRS with 196×196 elements has the highest instantaneous SNR. However, we also have to consider computational complexity and energy efficiency for IRS.

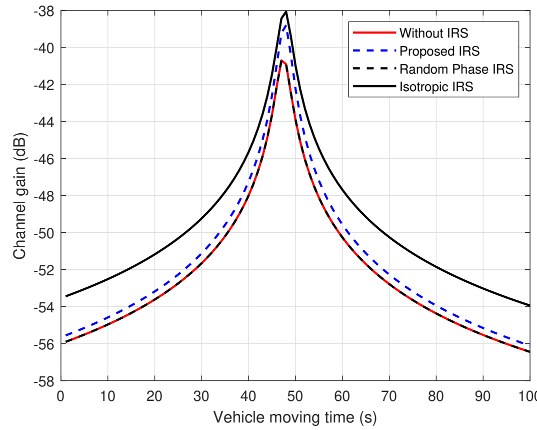

Fig. 3. Instantaneous SNR over time for one vehicle pass.

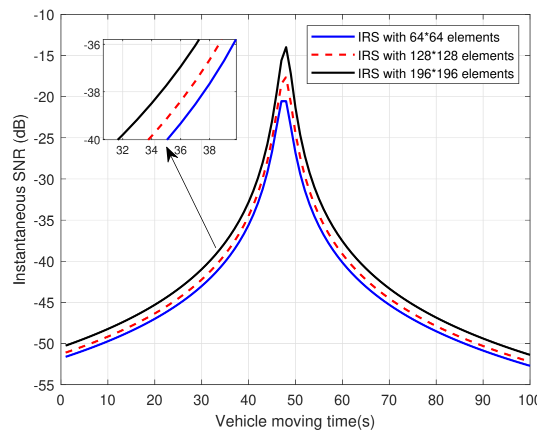

Fig. 4. Instantaneous SNR versus different number of elements of IRS over time for one vehicle pass.

B. Delay Spread for One Vehicle Pass

Fig. 5 plots the delay spread of 1) propose IRS, 2) random phase IRS, and 3) their difference for one vehicle pass. It can be observed that there is almost no difference between those two delay spreads. Therefore, although the sub-optimal phase shift set $\mathcal{T}(t)$ cannot diminish the delay spread to zero, it keeps the spread to a relatively low range.

C. Optimal Phase Shift for One Vehicle Pass

In Fig. 6, we choose four elements from 128×128 elements of the IRS to make visual illustrations of the designed sub-optimal phase shifts for one vehicle pass. Since the information of vehicle is known a priori, the sub-optimal phase shift by every element can be stored in the ROM of the IRS in advance. Therefore, compare to ordinary IRSs [3], [5]–[10], the IRS with the scheme we propose in this paper does not need channel estimation or other complicated processes. In other words, it is more cost-effective, more practical, and of higher energy efficiency.

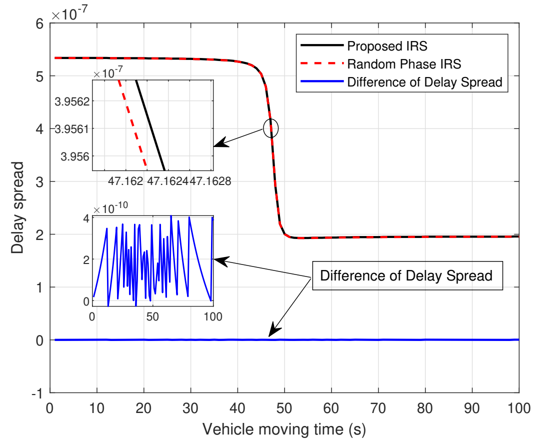

Fig. 5. Delay spread of propose and random phase IRS and their difference for one vehicle pass.

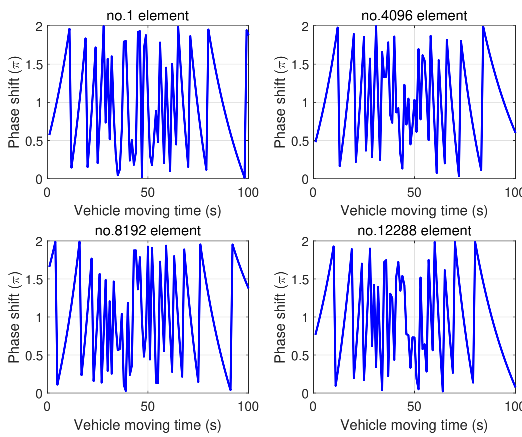

Fig. 6. Sub-optimal phase shifts by the 1st, 4096th, 8192th and 12288th element of a 128×128 IRS for one vehicle pass.

VI. Conclusions

In this paper, we propose a novel design for the optimal phase shift of IRS that takes full advantage of the predictable information of high-mobility vehicle in the vehicular communication system. A sub-optimal phase shift set of a multi-objective problem is obtained, which can maximize the instantaneous signal-to-noise ratio (SNR), totally eliminate the Doppler spread, and keep the delay spread to a relatively low range simultaneously. Simulation results show an instantaneous SNR gain from 0.68 dB to 4.21 dB for an IRS with 128×128 elements can be achieved over one vehicle pass. In the future, the research directions include 1) the influence by predictive information error for the propose IRS system, 2) how to optimize the phase set if the scenario has extra scatters, 3) how to utilize IRS to eliminate delay spread and Doppler shift and 4) the impact of random damage and discrete phase shifts for elements of IRS.

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