A geometry-based stochastic channel model and its application for intelligent reflecting surface assisted wireless communication

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Abstract
Intelligent reflecting surface (IRS) is a new concept originating from metamaterials, which can achieve beamforming through controllable passive reflecting. This device makes it possible to engineer the wireless communication environment, and has drawn increasing attention. However, the associated channel models in current literature are mainly borrowed from conventional wireless channel models directly, omitting the unique features of IRS. In this paper, a geometry-based stochastic channel model for IRS-assisted wireless communication system is employed. The model has certain accuracy and low computational complexity. In particular, it captures the correlations of subchannels associated with different IRS elements, which is typically not considered in current works. Based on this channel model and the derived channel spatial correlation functions (CFs), an iterative reflection coefficients configuration method is proposed exploiting statistical channel state information to maximise the ergodic channel capacity. The impacts of the IRS spatial positions as well as the number of the IRS elements on the ergodic channel capacity is investigated through simulations. It is found that to obtain a larger ergodic channel capacity, the IRS should be placed in the vicinity of either the transmitter side or the receiver side, which is a useful guideline for practical deployment.

1 INTRODUCTION

In the fifth generation (5G) wireless communication, massive multiple-input multiple-output (MIMO), millimeter wave (mmWave) were proposed in the physical layer as key technologies to improve the network capacity and achieve the requirements of 1000 times increase in data rate [1]. For instance, mmWave frequencies offer an enormous amount of spectrum, ranging from 30 GHz to 300 GHz, which is a feasible solution to the current shortage of spectrum resource. The large bandwidth can easily provide a data rate up to gigabits per second. Therefore, exploiting more spectrum resources is a clear path to increase the channel capacity. Meanwhile, massive MIMO makes it possible to overcome the severe path loss introduced by high frequency. It can also reduce the interference by flexible beamforming. Hence, exploiting spatial multiplexing gain is yet another path to achieve increased channel capacity [2, 3]. However, all those technologies only focus on the design of transmitters and receivers. The wireless channels, on the other hand, are only to be adapted without being changed. As higher frequencies are to be used in 5G and beyond 5G wireless communications systems, the coverage of base stations (BS) may be much smaller than that of 4G systems. Thus more BSs are needed, which complicates the interplay between the transceivers and...
the wireless propagation environments. With such visions, it would be desirable for the wireless channels, or environments, to be “engineerable” under active control.

Fortunately, with the development of meta-materials in recent years, controlling the effective wireless channels to some extent becomes a reality. Meta-materials are artificial structures that entitle non-nature characteristics [4, 5]. Intelligent reflecting surface (IRS), or large intelligent surface/antennas (LISA), is a representative meta-material that has found its interesting role in wireless communications. IRS is a planar surface composed of large amount of passive reflecting elements. Each element can independently change the amplitude and phase of the incident signal wave, thus making a collaborative reflecting beamforming [6]. By carefully configuring the reflection coefficients of all elements, IRS can be used to enhance the wireless communication performance in various aspects. IRS can be arranged around the BSs on a large scale to play the role of auxiliary communication [7, 8]. Besides, IRS can also shield the monitoring of illegal users and realise the confidential communication [9, 10]. In addition, IRS is a passive device and has lower power consumption and more simple hardware structure than traditional active relay, which means the cost is quite low. Current researches on IRS-assisted wireless communication mainly focused on the configuration of IRS reflection coefficients to optimise various kinds of performance metrics, such as ergodic spectral efficiency, weighted sum-rate and secrecy rate. Moreover, the deterioration in performance caused by discretisation in reflection coefficients had also been studied. Some comprehensive surveys of those works can be found in [11–13]. It is worth noting that most of those researches were based on the assumption that accurate channel state information (CSI) of the channels between IRS and involved transmitters and receivers was available. However, since IRS is a passive device and the number of the reflecting elements on IRS turns to be very large, the channel estimation becomes a challenging problem. Some researches concentrated on how to obtain instantaneous CSI [14, 15]. On the other hand, establishing a suitable channel model is also desired to serve as a basis for the evaluation of the performance of different channel estimation algorithms. From the perspective of system performance analysis, channel modelling for IRS-assisted communication system is also of great importance.

As a new research topic in IRS-assisted wireless communication systems, there were limited works in channel modelling at current stage. Most researches inherited existing models such as Rician fading channel model [7, 9, 16] and Rayleigh fading channel model [17, 18]. For instance, [7] used Rician fading model for the IRS associated channels, in which the line-of-sight (LOS) part was set as a deterministic process while the non-line-of-sight (NLOS) part was assumed to be i.i.d. complex Gaussian distributed, which may not be the case for IRS since the elements are regularly spaced in a limited area. Correlations were not captured in such model. Therefore, more accurate channel model tailored for IRS-assisted wireless communication system should be developed. Authors of [19] proposed a free space path loss model of the IRS-assisted wireless communication system based on the physical and electromagnetic characteristics of the IRS, which was a deterministic model and the statistical channel model is still desired. Recently, [20] employed a geometry-based stochastic channel model (GSCM) for IRS-assisted wireless communication system. The GSCM combines the statistical model with the deterministic model. On one hand, it has better accuracy than the statistical channel and can be better combined with MIMO technology [21, 22]. On the other hand, it has lower computational complexity than the deterministic channel modelling. The scatterers’ positions are generated randomly according to some specific probability distributions. Therefore, this model can be adapted to different scenarios with certain accuracy. This modelling method had been applied to V2V communication [23, 24], massive MIMO communication etc., and performs well. Although GSCM was investigated in [20], several deficiencies make it unsuitable for system performance assessment. Firstly, modelling of path loss was not considered in [20], which plays a decisive role in the performance analysis of IRS-assisted communication systems. Secondly, when applying GSCM, only an isolated single IRS element was investigated. The geometry layout relationship among IRS elements, that is, the number of IRS elements as well as their size and spacing, which can be regarded as the essential feature that distinguishes IRS plane from conventional scattering objects, were left untouched in channel modelling and characterisation. As a consequence, the channel correlations between different IRS elements, which is the basis for capacity analysis and motivation of employing GSCM for IRS-assisted system in this work, is still unknown.

In this paper, by overcoming the above issues, we employ a GSCM for IRS-assisted wireless communication in a more comprehensive way. Based on the GSCM, the spatial correlations between different IRS elements under certain system configurations are identified. Based on the spatial correlations, the ergodic channel capacity is obtained by optimally setting the IRS reflection coefficients, which is done by an iterative algorithm that has low complexity. The contributions of this paper can be summarised as follows:

i. The GSCM for IRS-assisted wireless communication system is fully and generally characterised for each individual subchannel, considering the large-scale fading effect due to pass loss, the small-scale fading effect due to scattering, and most importantly, the geometry relationship between different IRS elements.

ii. Facilitated by the GSCM, the spatial correlation functions (CFs) of up to the third and fourth orders between different subchannels are identified, thus capturing the correlation effects between subchannels induced by different IRS elements, which are usually omitted in current literature.

iii. Furthermore, according to the spatial CFs, an iterative algorithm of optimising the reflecting coefficients to achieve the maximum ergodic capacity is proposed, which entitles closed-form expression at each iteration and thus has low complexity and converges quickly. In addition, the algorithm only depends on the spatial CFs and does not rely on specific channel distributions.
Various simulations are conducted to reveal the potential gain of IRS and its impacts on the channel capacity with different configurations, including its spatial positions, the total number of elements, carrier frequencies, as well as the distance of adjacent elements.

The rest of this paper is organised as follows. Section 2 gives a general description of the proposed GSCM. Section 3 gives the spatial CFs of different subchannels derived from the proposed channel model. A low-complexity coefficients optimisation algorithm is detailed in Section 4. Simulations are presented in Section 5. Conclusions are drawn and future works are discussed in Section 6.

**Notation:** Vectors and matrices are denoted by boldface lowercase letters and boldface uppercase letters, respectively. For a vector \( \mathbf{a} \), \( \mathbf{a}[n] \) denotes its \( n \)th element, and for a matrix \( \mathbf{B} \), \( \mathbf{B}[m,n] \) denotes its element in the \( m \)th row and \( n \)th column. \( \mathbb{C}^{M \times N} \) denotes the space of \( M \times N \) complex-valued matrices. \( \mathbb{N} \) denotes the set of nonnegative integers. \( \mathbb{E}\{\cdot\} \) denotes the expectation operation and \( (\cdot)^* \) denotes conjugate operation. For a complex-valued vector \( \mathbf{x} \), \( \|\mathbf{x}\| \) denotes its Euclidean norm. For any general matrix \( \mathbf{X} \), \( \mathbf{X}^T \) and \( \mathbf{X}^H \) denotes the transpose and Hermitian transpose, respectively. For any general vector \( \mathbf{x} \), \( \text{diag}(\mathbf{x}) \) denotes the diagonal matrix whose entries on the main diagonal are given by \( x \).

## 2 | CHANNEL MODELLING

In this section, we apply the GSCM to an IRS-assisted wireless communication system with a single antenna at the transmitter (Tx) and a single antenna at the receiver (Rx). While this model can be extended to the more general MIMO scenario, possibly with much complicated expressions, the basic ideas behind them are essentially the same. Therefore, this paper focuses on the SISO scenario with the aim of better exposing the basic idea of the proposed model and its convenience in assessing the performance of IRS-assisted communication systems. The basic idea of the channel model is derived from ellipse channel model, that is, the effective scatterers with the same delay are located on the same ellipse [25]. Using different probability density function (PDF) of the scatterers can adapt the model to various types of environments.

### 2.1 | Overview of the subchannels and geometry

Figure 1 shows the diagram of an IRS-assisted wireless communication system. The LOS channel between the transmitter and the receiver may be blocked such that an IRS module is deployed to compensate for the potential signal quality degradation at the receiver side. The Tx, IRS, and Rx are placed in the \( x-y \) plane of a Cartesian coordinate system, where the \( x \)-axis is defined as the line connecting the Tx and the Rx. Without loss of generality, here we assume the lower edge of the IRS is parallel with the \( x \)-axis. The IRS module consists of \( P \) elements forming a rectangular plane, where \( M \) and \( N \) are the number of columns and rows of the regularly arranged elements of the IRS, respectively, that is, \( P = M \times N \). Each element is also a smaller rectangular with length \( d_x \) and width \( d_y \) along the \( x \) and \( y \) axes, respectively. The element in the \( m \)th row and \( n \)th column is referred to as \( u_{mn} \). The coordinates of the center point of \( u_{mn} \) is therefore \( ((m - 1/2)d_x, (n - 1/2)d_y) \), where \( m \in [1,M] \) and \( n \in [1,N] \).

Three subchannels can be identified in Figure 1, that is, the scalar direct Tx-Rx channel \( h_3 \in \mathbb{C}^{1 \times 1} \), the vector Tx-IRS...
channel $h_1 \in \mathbb{C}^{P \times 1}$, and the vector IRS-Rx channel $h_2 \in \mathbb{C}^{L \times 1}$. Modelling of those channels will be detailed later. Different from conventional wireless communication, each element of the IRS is capable of re-scattering the incident signal with an individual reflection coefficient which can be dynamically adjusted by the IRS controller for desired signal reflection behaviors. The IRS controller itself can be configured by the transmitter or the receiver through a dedicated wireless or wired control channel. Therefore, the effective channel can be actively controlled to some extent in IRS-assisted wireless communication. We denote $v_{mn}$ as the reflection coefficient of the IRS element $u_{mn}$. It is assumed that $|v_{mn}| = 1$ while the phase of $v_{mn}$ can be flexibly adjusted in $[0, 2\pi)$. The effective channel between the Tx and the Rx can be expressed as

$$h = h_3 + h_2^T \Theta h_1,$$

(1)

where $\Theta \in \mathbb{C}^{P \times P}$ denotes the reflection coefficient matrix, $\Theta$ is a diagonal matrix whose entries on the main diagonal are given by $\nu = [\nu_1 \cdots \nu_{MN}] \in \mathbb{C}^{P \times 1}$.

### 2.2 Channel model between the Tx and IRS

The vector channel between the Tx and IRS can be expressed as

$$h_1 = [b_{1,11} \cdots b_{1,mm} \cdots b_{1,MM}]^T \in \mathbb{C}^{P \times 1},$$

(2)

where $b_{1,um}$ represents the complex channel impulse response (CIR) between Tx and the element $u_{mn}$. It is assumed that the LOS and NLOS propagation links from the Tx to $u_{mn}$ are independent of each other. Therefore, the complex CIR between the transmit antenna and the element $u_{mn}$ can be expressed as

$$b_{1,um} = b_{1,um}^{\text{LOS}} + b_{1,um}^{\text{NLOS}},$$

(3)

It is important to mention that the complex CIR $b_{1,um}^{\text{LOS}}$ is a deterministic process due to that the transmitter and the IRS are fixed. And when the signals transmit through different scatterers in NLOS channels, the complex CIR $b_{1,um}^{\text{NLOS}}$ is a random process because of the random variations in phase caused by the scatterers [26]. In this paper, we consider the case in which the Tx, the IRS, and the Rx are static. Therefore the complex CIRs $b_{1,um}^{\text{LOS}}$ and $b_{1,um}^{\text{NLOS}}$ are stationary.

The LOS part $b_{1,um}^{\text{LOS}}$ can be expressed as

$$b_{1,um}^{\text{LOS}} = \sqrt{\frac{\Omega_1}{\Omega_1 + 1}} e^{-j2\pi f \xi_{\text{T,R}} \cos(\theta_1 - \theta_{um})} \times b_{1,um}^{\text{LOS}},$$

(4)

where $\Omega_1$ denotes the Rician factor of this subchannel, $f$ is the carrier frequency, and $c$ is the speed of light. $\xi_{\text{T,R}}$ denotes the distance between Tx and $u_{mn}$, which can be expressed as

$$\xi_{\text{T,R}} = \sqrt{\xi_{\text{IRS}}^2 + \xi_{\text{um}}^2 + 2\xi_{\text{T,IR}} \xi_{\text{um}} \cos(\theta_1 - \theta_{um})},$$

(5)
where $\xi_{\text{TIRS}}$ denotes the distance between Tx and the lower left corner of IRS, and can be expressed as

$$\xi_{\text{TIRS}} = \sqrt{D^2 + H^2}. \quad (6)$$

$D_1$ denotes the horizontal distance between the Tx and the left edge of the IRS. $H$ denotes the vertical distance between the lower edge of the IRS and the $x$-axis. $\xi_{\text{mn}}$ denotes the distance between $u_{\text{mn}}$ and the lower left corner of IRS, and can be expressed as

$$\xi_{\text{mn}} = \sqrt{\left(\frac{n - 1}{2} d_x\right)^2 + \left(\frac{n - 1}{2} d_y\right)^2}. \quad (7)$$

And $\theta_1$ denotes the tilt angle between the $x$-axis and the line connecting the center point of the Tx antenna array and the edge of the IRS system. $\theta_{\text{mn}}$ denotes the included angle between the $x$-axis and the line connecting $u_{\text{mn}}$ and the lower left corner of the IRS, and can be expressed as

$$\theta_{\text{mn}} = \arctan\left(\frac{n - 1}{2} d_y}{m - 1/2 d_x}\right). \quad (8)$$

$\beta_{\text{LOS}}^{\text{Tmn}}$ denotes the large-scale path loss of the LOS channel from the Tx to $u_{\text{mn}}$, which can be expressed as

$$\beta_{\text{LOS}}^{\text{Tmn}} = \sqrt{\frac{A_{\text{IRS}} G_r}{4\pi}} \times \frac{1}{\xi_{\text{Tmn}}}, \quad (9)$$

where $A_{\text{IRS}}$ is the effective area of the IRS, which can be approximated by the physical area of IRS [27]. $G_r$ denotes the gain of the transmitter.

And the NLOS part $b_{\text{NLOS}}^{\text{Tmn}}$ can be expressed as

$$b_{\text{NLOS}}^{\text{Tmn}} = \sqrt{\frac{1}{\Omega_1 + \sum_{k_1=1}^{K_1} \chi_1}} \times \left(e^{-j2\pi f_s (\xi_{\text{Tmn}} + \xi_{\text{mn}})}/e^{j\varphi_{0}} \times \beta_{\text{NLOS}}^{\text{Tmn}}\right), \quad (10)$$

where $\chi_1$ denotes the number of visual scatterers in the link between Tx and IRS, $\chi_1$ denotes the path gain caused by visual scatterers, and $\xi_{\text{Tmn}}$ is the distance from the transmit antenna to the visual scatterer $j^{(k_1)}$, which can be expressed as

$$\xi_{\text{Tmn}} = \frac{b_1^2}{b_1^2 \cos^2 \left(\frac{\alpha_{(k_1)}}{\gamma} - \theta_1\right) + a_1 b_1^2} \times \frac{1}{\xi_{\text{Tmn}}}. \quad (11)$$

where $a_1$ and $b_1$ are, respectively, the semi-lengths of the major axis and minor axis of the ellipse channel model between the Tx and IRS. Furthermore, $\xi_{\text{mn}}$ is the distance from $u_{\text{mn}}$ to the visual scatterer $j^{(k_1)}$, which can be expressed as

$$\xi_{\text{mn}} = \sqrt{\frac{2}{b_1^2} + \frac{2}{b_1^2}}. \quad (12)$$

where $\xi_{\text{IRSR}}$ denotes the distance between the lower left corner of IRS and the visual cluster $j^{(k_1)}$, and can be expressed as

$$\xi_{\text{IRSR}} = \frac{2}{b_1^2} \cos \left(\frac{\alpha_{(k_1)}}{\gamma} - \theta_1\right) + a_1 b_1^2, \quad (13)$$

where $\alpha_{(k)}^{(k_1)}$ denotes the angle between $x$-axis and line connecting lower left corner of the IRS and the visual scatterer $j^{(k_1)}$. $\varphi_0$ is a random phase angle and represents the variations in phase caused by the scatterers, which follows a uniform distribution in the interval from $-\pi$ to $\pi$ [24, 26]. $\beta_{\text{NLOS}}^{\text{IRSR}}$ denotes the large-scale path loss of the NLOS channel from the Tx to $u_{\text{mn}}$ via visual scatterer $j^{(k_1)}$, and can be expressed as

$$\beta_{\text{NLOS}}^{\text{IRSR}} = \sqrt{\frac{A_{\text{IRS}} G_r}{4\pi}} \times \frac{1}{\xi_{\text{Tmn}} + \xi_{\text{mn}}}. \quad (14)$$

2.3 Subchannel model between the IRS and Rx

Using a similar approach, the channel between the IRS and the Rx can be described as

$$h_2 = [b_{11,R} \cdots b_{m,R} \cdots b_{MN,R}]^T \in \mathbb{C}^{P \times 1}. \quad (15)$$

where $b_{m,R}$ represents the complex CIR of the propagation link between the unit cell $u_{\text{mn}}$ and the Rx. Different from [20] in which only the NLOS propagation links were considered, here we incorporate both the LOS and NLOS links from $u_{\text{mn}}$ to the Rx which is more realistic. Specifically, the complex CIR can be expressed as

$$h_{\text{mn}} = h_{\text{LOS}}^{\text{mn}} + h_{\text{NLOS}}^{\text{mn}}, \quad (16)$$

where

$$h_{\text{LOS}}^{\text{mn}} = \sqrt{\frac{\Omega_2}{\Omega_2 + 1}} e^{-j2\pi f_s \xi_{\text{mn}}}/e^{j\varphi_{0}} \times \beta_{\text{LOS}}^{\text{IRSR}}, \quad (17)$$

where $\Omega_2$ denotes the Rician factor of this subchannel, $\xi_{\text{mn}}$ is the distance for the direct propagation link from the unit cell $u_{\text{mn}}$ to the Rx, which can be expressed as

$$\xi_{\text{mn}} = \sqrt{\frac{2}{b_1^2} + \frac{2}{b_1^2} \cos \left(\frac{\alpha_{(k_1)}}{\gamma} - \theta_2\right), \quad (18)$$
where $\xi_{\text{IRS,R}}$ denotes the distance between Rx and the lower left corner of IRS, and can be expressed as

$$\xi_{\text{IRS,R}} = \sqrt{D_2^2 + H^2}, \quad (19)$$

where $D_2$ denotes the horizontal distance between the left edge of the IRS and the Rx. $\theta_2$ is the tilt angle between the $x$-axis and the line connecting the Rx and $\xi_{\text{IRS,R}}$. $\beta_{\text{LOS}}^n$ denotes the large-scale path loss of the LOS channel from $\mu_{\text{nn}}$ to Rx, which can be expressed as

$$\beta_{\text{LOS}}^n = \sqrt{\frac{A_{\text{IRS}} G_r}{4\pi}} \times \frac{1}{\xi_{\text{nn,R}}}, \quad (20)$$

where $G_r$ denotes the gain of receiver antenna.

The NLOS part can be expressed as

$$b_{\text{NLOS}}^{\text{nn},R} = \sqrt{\frac{1}{\Omega_2}} \sum_{k_2=1}^{\mathcal{K}_2} \chi_2 \times \left( \sqrt{\left(\frac{\xi_{\text{IRS,K}_2}}{\xi_{\text{nn,K}_2}}\right)^2 + \left(\frac{\sqrt{2}}{\xi_{\text{nn,K}_2}} \xi_{\text{nn}} \cos \left(\alpha_{\text{IRS,K}_2} + \theta_2 - \theta_{\text{nn}}\right)\right)^2} \right), \quad (21)$$

where $\mathcal{K}_2$ denotes the number of visual scatterers in the link between IRS and Rx, $\chi_2$ denotes the path gain caused by visual scatterers, $\xi_{\text{nn,K}_2}$ is the distance from the element $\mu_{\text{nn}}$ to the visual scatterer $s_{(k_2)}$, and can be expressed as

$$\xi_{\text{nn,K}_2} = \sqrt{\xi_{\text{IRS,K}_2}^2 + \xi_{\text{nn}}^2 - 2 \xi_{\text{IRS,K}_2} \xi_{\text{nn}} \cos \left(\alpha_{\text{IRS,K}_2} + \theta_2 - \theta_{\text{nn}}\right)}, \quad (22)$$

where $\xi_{\text{IRS,K}_2}$ denotes the distance between $\mu_{\text{nn}}$ and the visual scatterer $s_{(k_2)}$, and can be expressed as

$$\xi_{\text{IRS,K}_2} = \sqrt{\frac{d_2^2 - b_2^2 \cos \left(\alpha_{\text{IRS,K}_2} + \theta_2\right)}{b_2 \cos \left(\alpha_{\text{IRS,K}_2} + \theta_2\right)} + a_2^2 r^2}, \quad (23)$$

where $\xi_{\text{IRS,K}_2}$ denotes the distance between the lower left corner of IRS and the visual scatterer $s_{(k_2)}$, $\alpha_{\text{IRS,K}_2}$ denotes the angle between $x$-axis and the line connecting lower left corner of the IRS and the visual scatterer $s_{(k_2)}$. And $\xi_{\text{R,K}_2}$ is the distance from the visual scatterer $s_{(k_2)}$ to the Rx, which can be expressed as

$$\xi_{\text{R,K}_2} = \sqrt{\frac{d_2^2 - b_2^2 \cos \left(\alpha_{\text{R,K}_2} - \theta_2\right)}{b_2 \cos \left(\alpha_{\text{R,K}_2} - \theta_2\right)} + a_2^2 r^2}, \quad (24)$$

where $a_2$ and $b_2$ are, respectively, the semi-lengths of the major axis and minor axis of the ellipse channel model between the IRS and Rx. $\alpha_{\text{R,K}_2}$ is the angle of arrival (AoA) for the Rx.

Similarly, $b_{\text{NLOS}}^{\text{nn},K_{2,R}}$ denotes the large-scale path loss of the NLOS channel from $\mu_{\text{nn}}$ to Rx by visual scatterer $s_{(k_2)}$, and can be expressed as

$$b_{\text{NLOS}}^{\text{nn},K_{2,R}} = \sqrt{\frac{A_{\text{IRS}} G_r}{4\pi}} \times \frac{1}{\xi_{\text{nn,K}_2,R}}. \quad (25)$$

### 2.4 Subchannel model between Tx and Rx

The subchannel between the Tx and Rx can be described as scalar $b_3$, which denotes the complex CIR of the propagation link from the Tx to the Rx via the reflection of the visual scatterers $s_{(k_3)}$. It can be expressed as

$$b_3 = \sum_{k_3=1}^{\mathcal{K}_3} \chi_3 e^{j\phi_3} e^{j\psi_{k_3}} \times \beta_{\text{LOS}}^{\text{nn},R}, \quad (26)$$

where $\mathcal{K}_3$ denotes the number of visual scatterers in the link between Tx and Rx, $\chi_3$ denotes the path gain caused by visual scatterers, $\xi_{\text{dx}}$ denotes the distance from the Tx to the visual scatterer $s_{(k_3)}$, and can be expressed as

$$\xi_{\text{dx}} = \sqrt{d_3^2 - b_3^2 \cos \alpha_{\text{d}}^n + a_3^2 b_3^2 \cos \alpha_{\text{d}}^n + a_3^2 \sin^2 \alpha_{\text{d}}^n}, \quad (27)$$

where $a_3$ and $b_3$ are, respectively, the semi-lengths of the major axis and minor axis of the ellipse channel model between the Tx and IRS. $\alpha_{\text{d}}^n$ is the angle of departure (AoD) for the Tx. $\xi_{\text{r,K}_3}$ is the distance from the visual scatterers $s_{(k_3)}$ to the Rx, which can be expressed as

$$\xi_{\text{r,K}_3} = \sqrt{d_3^2 - b_3^2 \cos \alpha_{\text{r}}^n + a_3^2 b_3^2 \cos \alpha_{\text{r}}^n + a_3^2 \sin^2 \alpha_{\text{r}}^n}, \quad (28)$$

where $\alpha_{\text{r}}^n$ denotes the AoA for the Rx. And $\beta_{\text{NLOS}}^{\text{dx,K}_{3,R}}$ denotes the large-scale path loss of the NLOS channel from the Tx to Rx by visual scatterer $s_{(k_3)}$, which can be expressed as

$$\beta_{\text{NLOS}}^{\text{dx,K}_{3,R}} = \sqrt{\frac{G_r A}{4\pi \xi_{\text{dx}} + \xi_{\text{r,K}_3}}}. \quad (29)$$

By substituting (2), (15), and (26) into (1), the effective channel model can be obtained, which can be expressed as

$$h = h_3 + h_3' h_1 = \sum_{m=1}^{\infty} \sum_{n=1}^{\infty} b_{\text{T,m}a\text{nn},R} b_{\text{nn},R} + b_3. \quad (30)$$

In the following, based on the above channel model, we will analyse some properties of the channel and their impacts on the channel capacity.
2.5 | Feasible region

The proposed GSCM for IRS-assisted wireless communication system is tailored for SISO narrowband wireless communication system, so the multipath effect could not be included in this model. In this subsection, we will analyse the reasonable locations to place the IRS and select appropriate parameters to ensure that the Rx can receive the signal with sufficiently small delay with respect to the direct path.

We denote $W$ as the operation bandwidth of the system, and $T_s = \frac{1}{W}$ as the sampling interval or the time duration for each symbol. In order to ensure that the Rx perceives a narrow band channel, we require that the delay spread of the signal, caused by the reflections of the scatterers, must be small enough. Here, we assume that the maximum time delay of different propagation links is within $\frac{T_s}{10}$. Under this requirement, based on the geometry shown in Figure 3, the parameters should satisfy the following inequalities

$$c_3 \leq c_1 + c_2 \leq c_3 + \frac{T_s}{10} \times \frac{c}{2}, \quad (31)$$
$$c_1 \leq a_1 \leq \frac{T_s}{10} \times \frac{c}{2} + c_3 - a_2, \quad (32)$$
$$c_2 \leq a_2 \leq \frac{T_s}{10} \times \frac{c}{2} + c_3 - a_1, \quad (33)$$
$$c_3 \leq a_3 \leq \frac{T_s}{10} \times \frac{c}{2} + c_3. \quad (34)$$

According to (31), we can draw the feasible region to place IRS such that narrow band channel model suffices, which is an ellipse with Tx and Rx as the focal points as shown in Figure 3. According to (32) and (33), we can determine the feasible region of $a_1$ and $a_2$, which is an isosceles right triangle as shown in Figure 4. We choose the midpoint on the hypotenuse in simulation to make both $a_1$ and $a_2$ big enough, where

$$a_1 = \frac{1}{2} \left( c_1 - c_2 + c_3 + \frac{T_s}{10} \times \frac{c}{2} \right), \quad (35)$$
$$a_2 = \frac{1}{2} \left( -c_1 + c_2 + c_3 + \frac{T_s}{10} \times \frac{c}{2} \right). \quad (36)$$

As for $a_3$, we can determine its value based on (34). Here we set $a_3$ as its upper bound. In the following, we set the bandwidth as $W = 2$ MHz, that is, $T_s = 0.5 \mu$s.

3 | SPATIAL CFS

The spatial CFs can be used to measure the correlation between two different channels, which can be expressed as

$$\rho_{b_p,qh_{ij}} = \mathbb{E}\left\{ b_{p,q} b_{ij}^* \right\}, \quad (37)$$

where $b_{p,q}$ denotes CIR of the channel between nodes $p$ and $q$, $h_{ij}$ denotes CIR of the channel between nodes $i$ and $j$, where $p, q, i$ and $j$ can represent the Tx, the Rx or the IRS element $m$. Note that $\varphi_0$ is the only random variable, thus $\mathbb{E}\{\cdot\}$ in (37) only applies to $\varphi_0$. Substituting (3) and (16) into (37), the spatial CFs of any pair of the aforementioned subchannels can be derived from the proposed channel model as

$$\rho_{b_{mn,h_{Eij}}} = \frac{1}{2\pi} \int_{-\pi}^{\pi} b_{T,\text{in}}(\varphi_0) b_{h_{Eij}}^*(\varphi_0) d\varphi_0, \quad (38)$$
$$\rho_{b_{mn,h_{ij}}^{R}} = \frac{1}{2\pi} \int_{-\pi}^{\pi} b_{\text{in},R}(\varphi_0) b_{h_{ij}}^{R}(\varphi_0) d\varphi_0, \quad (39)$$
$$\rho_{b_{mn,h_{ij}}^{R}} = \frac{1}{2\pi} \int_{-\pi}^{\pi} b_{\text{in},R}(\varphi_0) b_{h_{ij}}^{R}(\varphi_0) d\varphi_0. \quad (40)$$
The discrete AoAs $\alpha_1^{(k)}$, $\alpha_{IRS}^{(k)}$, and $\alpha_R^{(k)}$ can be replaced as continuous variables $\alpha_i^{(1)}$, $\alpha_{IRS}^{(2)}$ and $\alpha_R^{(3)}$ with PDFs $f(\alpha_i^{(1)})$, $f(\alpha_{IRS}^{(2)})$ and $f(\alpha_R^{(3)})$, respectively. When the number of scatterers approaches infinity, that is, $\mathcal{K}_1 \to \infty$, $\mathcal{K}_2 \to \infty$, $\mathcal{K}_3 \to \infty$, we use the von Mises PDF to approximate the distributions [28]. The von Mises PDF is defined as

$$f(\alpha) = \frac{1}{2\pi L_0(\kappa)} e^{\kappa \cos(\alpha - \mu)},$$

where $L_0(\cdot)$ denotes the zeroth-order modified Bessel function of the first kind, $\mu(\cdot) \in [-\pi, \pi]$ is the mean value of the angle $\alpha$, and $\kappa(\cdot) \in \mathbb{N}$ represents the environment-related parameter. It represents isotropic scattering when $\kappa = 0$, and $f(\alpha) = 0.5\pi$. While $\kappa = \infty$, $f(\alpha) = \delta(\alpha - \mu)$ represents extremely nonisotropic scattering. In this paper, we assume $\kappa = 0$, $\mu = 0$, and the results are given in Appendix. Although the forms seem very complex, the integrals of different AoAs and AoDs can be carried out separately, such that the final results can be easily evaluated by efficient numerical integral processing.

As an extension, higher order spatial CF quantities, involving more than two subchannels, can be defined and efficiently calculated using similar approach, which will also be used in the optimisation of the IRS reflection coefficients. The detailed calculations of those higher order spatial CFs are omitted here. In Section 4, up to the fourth order of spatial CFs will be utilised to assess the ergodic rate performance. Here, we illustrate some calculations of those higher order spatial CFs are omitted here. In this paper, we assume $\kappa = 0$, $\mu = 0$, and the results are given in Appendix. Although the forms seem very complex, the integrals of different AoAs and AoDs can be carried out separately, such that the final results can be easily evaluated by efficient numerical integral processing.

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The parameters we used in this paper can be roughly divided into three groups. Firstly, the parameters about the size of the IRS elements, such as $d_s$, $d_l$ and $P$, are borrowed from [6]. Secondly, the parameters about the wireless environments such as $\Omega_1, \Omega_2, \chi_i (i = 1, 2, 3)$, $c_i$, are determined according to the communication scenario we choose. In this paper we assume that IRS works in the microcell of a mobile communication system whose communication range is limited to about 100 m. Finally, the channel model we established is tailored for narrow band, and the IRS should be deployed in some feasible region. The parameters about the location of IRS, such as $c_1, c_2, a_1, a_2$, can be calculated according to the above parameters as we discussed in Section 2.5.

Firstly, we evaluate the spatial CFs of the subchannels involved IRS as we mentioned in (38), (39) and (40). Figure 5 depicts the second order spatial CFs for different subchannels. It can be found that the correlation of those subchannels is obvious. Subchannels corresponding to the adjacent IRS elements show similar spatial CFs to other subchannels, and because the IRS elements are arranged regularly, the spatial CFs of the relative subchannels also present a spatial periodicity.

Figure 5 clearly reveals the fact that the spatial correlations of IRS associated channels are significant and cannot be simply omitted in system performance assessment. Moreover, we investigate the impact of carrier frequency $f_c$ on the spatial CF. It should be mentioned that the size of the IRS element must be smaller than half wavelength to ensure good approximation of the far field electromagnetic radiation model for both LOS and NLOS channels. The size of IRS element is set as $d_s = d_l = 1$ cm, which means the model can theoretically characterise the channel with carrier frequency up to 15 GHz. As shown in Figure 6, the spatial periodicity still exists when the frequency is low, but when frequency increases, the variation becomes dramatic and the spatial periodicity turns indistinctive.

### 4 Reflection Coefficients Optimisation Based on Spatial CFs

The GSCM proposed in Section 2 has many applications. For example, based on the spatial CFs which are derived from this model, we can analyse the ergodic channel capacity of IRS-assisted wireless communication, which is a key performance metric of a communication system.

#### 4.1 Problem formulation

To proceed, we write the received signal as

$$y = (h_1 + h_2 \Theta h_1) x + \zeta$$

(42)
where $x$ denotes the information symbol with power $P_x$, $z$ is the noise component, which is circularly symmetric complex Gaussian (CSCG) distributed with zero mean and variance $\sigma_z^2$.

$y$ denotes the received signal at the receiver. We redefine the subscripts of the main diagonal entries in matrix $\Theta$ for the convenience of calculations, that is, $v = [v_0, \cdots, v_{P-1}]^T \in \mathbb{C}^{P \times 1}$.

The ergodic capacity is given by

$$R(\Theta) = \mathbb{E}\left\{ \log_2 \left( 1 + \frac{\|b_3 + h_2^T \Theta h_1\|^2 P_x}{\sigma_z^2} \right) \right\},$$

(43)

where $R(\Theta)$ denotes the “achieved” capacity with reflecting coefficients $v$ (with unit bits/s/Hz), and the expectation is taken over the joint distribution of all involved channels. It can be seen that $R$ is “tunable” by configuration of $\Theta$. Therefore, in the following we investigate the maximisation of $R$ by optimising $\Theta$.

### 4.2 Optimisation algorithm

Direct calculation of (43) is difficult due to the logarithm operation. Instead, noting $\mathbb{E}\{\log_2(J)\}$ is a concave function, we can apply Jensen’s inequality to get an upper bound of (43), which is given by [7]:

$$R(\Theta) \leq \log_2 \left( 1 + \gamma \mathbb{E}\{\|b_3 + h_2^T \Theta h_1\|^2\} \right),$$

(44)

where $\gamma = P_x/\sigma_z^2$ is the transmit signal-to-noise ratio (SNR). Now maximising $R$ is equivalent to maximising (45)

$$J = \mathbb{E}\left\{ \|b_3 + h_2^T \Theta h_1\|^2 \right\}$$

$$= \mathbb{E}\left\{ \langle h_3 + h_2^T \Theta h_1 \rangle^H (h_3 + h_2^T \Theta h_1) \right\}$$

$$= \mathbb{E}\left\{ \|b_3\|^2 + h_1^H \Theta^H (h_2^T)^H h_3 + \|h_2^T \Theta h_1\| \right\}$$

$$+ \mathbb{E}\left\{ \langle h_2^H \Theta^H (h_2^T)^H h_2 \rangle \right\}$$

$$= \mathbb{E}\left\{ \|b_3\|^2 + \nu^H H_1^H \left( h_2^T \right)^H b_3 + \|h_2^T H_1 v\| \right\}$$

$$+ \mathbb{E}\left\{ \langle h_2^H H_1^H \left( h_2^T \right)^H H_1 \rangle \right\}$$

$$= E_{\nu} \left\{ \|b_3\|^2 + \nu^H H_1^H \left( h_2^T \right)^H b_3 + \|h_2^T H_1 v\| \right\}$$

$$+ \mathbb{E}_{\nu} \left\{ \langle H_1^H \left( h_2^T \right)^H H_1 \rangle \right\}$$

(45)

where

$$H_1 = \text{diag}(h_1)$$

(46)
and the \([m, n]^{th}\) entry of \(A = \mathbb{E}\{h_1 h_2^T\}\) is given by
\[
A[m, n] = \mathbb{E}\left\{ b_{1_m} b_{2_n} \right\},
\] (47)
which can be obtained using the approach in Section 3, where \(b_{1_m}\) denotes the \(m^{th}\) element in vector \(h_1\), \(b_{2_n}\) denotes the \(n^{th}\) element in vector \(h_2\). Similarly, the entries of \(b\), as well as the scalar \(v\), can all be obtained as defined in (45). Therefore, \(J\) can be expressed as
\[
J = \mathbf{v}^H \mathbf{A} \mathbf{v} + \mathbf{v}^H \mathbf{b} + \mathbf{b}^H \mathbf{v} + c,
\] (48)
where \(A, b, c\) are known quantities.

In this paper, we propose a low-complexity coefficients optimisation algorithm. The basic idea is to optimise the coefficients in a sequential and iterative manner. At each iteration, when updating a coefficient, others are treated as known and fixed ones. Then, for the very coefficient, a closed-form expression is derived. At each iteration, some quantities can be reused for every coefficient calculation. Therefore, the overall computational complexity is relatively low.

Specifically, at each iteration, for any coefficient \(v\), we decompose \(v\) by
\[
v = \tilde{v}_n + e_n v_n,
\] (49)
where
\[
\tilde{v}_n = [v_{n1}, v_{n2}, \ldots, 0, 0, \ldots, 0]^T,
\]
\[
e_n = [0, 0, 1, 0, \ldots, 0]^T.
\] (50)

\[
J = (\tilde{v}_n + e_n v_n)^H \mathbf{A} (\tilde{v}_n + e_n v_n) + (\tilde{v}_n + e_n v_n)^H \mathbf{b} + \mathbf{b}^H (\tilde{v}_n + e_n v_n) + c = (\tilde{v}_n^H \mathbf{A} \tilde{v}_n + \tilde{v}_n^H \mathbf{A} e_n v_n) + (\tilde{v}_n^H + v_n^H e_n^H) \mathbf{b} + \mathbf{b}^H (\tilde{v}_n + e_n v_n) + c = (\tilde{v}_n^H \mathbf{A} \tilde{v}_n + \tilde{v}_n^H \mathbf{b} + \tilde{v}_n^H v_n^H e_n^H \mathbf{A} e_n v_n + c)
\] (51)

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]

\[
J = \begin{bmatrix}
v_n^T (e_n^H \mathbf{A} e_n + b^H) + (\tilde{v}_n^H \mathbf{A} e_n + b^H) v_n
\end{bmatrix}
\]

\[
\begin{bmatrix}
a_n
\end{bmatrix}
\]
5 | NUMERICAL RESULTS

In this section, we investigate the performance of the IRS-assisted communication system based on the proposed model, and present numerical results of the proposed optimisation algorithm.

5.1 | Capacity versus SNR

Firstly, we evaluate the difference between the “achieved” ergodic capacity and its upper bound. The parameters are mainly borrowed from Table 1. To highlight the effect of IRS, we choose a larger sized IRS, where \( d_x = d_y = 5 \) cm. Here, the capacity upper bound is obtained by (44) using optimal reflection coefficients generated by Algorithm 1. The “achieved” ergodic capacity is obtained by inserting the aforementioned optimal reflection coefficients into (43) using numerical integration method. As shown in Figure 7, the capacity upper bound is very close to the “achieved” capacity using the optimal reflection coefficients, which means the proposed capacity upper bound can approximate the system’s actual capacity to a large extent. Note that the “achieved” capacity can be regarded as a special capacity lower bound corresponding to the very optimal reflection coefficients. So, we analyse the capacity upper bound instead of the capacity itself in the following of this paper for sake of convenience.

Secondly, we evaluate the capacity of the IRS-assisted wireless communication system with respect to SNR. We use the same parameters for simulation as in Figure 7. Moreover, we vary the number of IRS elements, including 10 \( \times \) 10, 15 \( \times \) 15, 25 \( \times \) 25 and 30 \( \times \) 30, and compare the capacities of the respective IRS-assisted wireless communication systems with optimised reflection coefficients using proposed algorithm. The result is shown in Figure 8. It can be observed in Figure 8 that as we enlarge the number of IRS elements, the energy reflected to Rx by the IRS also increases, which leads to the increased capacity. For a 30 \( \times \) 30 IRS, it can increase the capacity by about 20% compared to the system without IRS.

Finally, we place a 10 \( \times \) 10 IRS in different positions, including \( D_1 = 10 \) m, \( D_1 = 20 \) m, \( D_1 = 30 \) m and \( D_1 = 40 \) m, and compare the performance of three different systems: the IRS-assisted system with optimised reflection coefficients using proposed algorithm (which is denoted by “optimal coefficients” in the legend), IRS-assisted system with random reflection coefficients (which is denoted by “random coefficients” in the legend), and the conventional system without IRS (which is denoted as “without IRS” in the legend). As shown in Figure 9, IRS with optimised reflection coefficients has significant performance improvement, which increases the rate by about 2 bits/s/Hz over the system without IRS when the SNR = 40 dB and \( D_1 = 10 \) m. As the SNR increases, the gain remains approximately constant. At the same time, using random reflection coefficients has almost no improvement as compared to
the system without IRS. By placing the IRS in different locations, we find that the capacities are also slightly different. This suggests that there should be an optimal location to place the IRS for each SNR value.

5.2 | Capacity versus the location of IRS

To fully reveal the impact of IRS location on the system performance, in this subsection, we investigate how the ergodic capacity varies according to the geological position of the IRS. We set the Tx at (0,0), Rx at (55,0), and vary the coordination of the IRS within the feasible region as we have mentioned in Section 2.5. Figure 10 depicts the variation in capacity of the IRS-assisted wireless communication system. Capacity gain over that with random reflecting coefficients is also given. The results strongly suggest that the position of IRS has great impact on the capacity of IRS-assisted wireless communication system. In general, the capacity and the capacity gain is pronounced only when the IRS is placed at the vicinity of either the transmitter or the receiver. When the IRS is placed far away from the transmitter and the receiver, the performance is not significant. Moreover, we find that there are four pronounced regions near the receiver in Figure 10, which can be viewed as the optimal location to place the IRS in this case.

5.3 | How much power can the IRS reflect to the Rx

As we have assumed in Section 2, the direct TX-RX channel is blocked and IRS is deployed to assist the communication by establish a reflection link. A natural question is, how much power can the IRS reflect to the Rx? In this subsection, we investigate this issue by simulations. To quantify the gain of IRS, we define $\eta$ as the ratio of the powers of the following two channels: one is the effective channel of the IRS-assisted wireless communication system, that is, $b_3 + h_2^\top \Theta h_1$, the other is the direct NLOS subchannel $b_3$, and $\eta$ can be expressed as

$$\eta = \frac{\mathbb{E}\{\|b_3 + h_2^\top \Theta h_1\|^2\}}{\mathbb{E}\{|b_3|^2\}} = \frac{J}{\mathcal{I}}.$$ (55)

The parameters are mainly borrowed from Table 1, and some of them are changed to $D_1 = 5$ m, $H = 5$ m and $\chi_3 = 1$. Note that the absolute value of $\eta$ is affected by the actual values of $\chi_1$, $\chi_2$ and $\chi_3$, which vary in different environments and need measurements. Nonetheless, we are still able to study how this ratio changes, relatively, with respect to other important factors. Figure 11 depicts the ratio when the size and the number of IRS elements change. It can be found that increasing the size and/or the number of the IRS elements can increase the ratio $\eta$, which accords with intuition: a larger sized IRS can receive more energy, as a result, it reflects more energy to Rx, and makes the ratio $\eta$ larger.

6 | CONCLUSIONS AND FUTURE WORKS

In this paper, we established a GSCM for IRS-assisted communication, which can describe the large-scale path loss and the correlations of different subchannels. According to the proposed channel model, we analysed some characteristics of the channels, such as spatial CFs and ergodic capacity of the system. And a reflection coefficients optimisation algorithm based on spatial CFs was provided. Numerical simulations depicted...
that the IRS can improve the capacity of the system, while the location of IRS should be around the transmitter or the receiver. The GSCM presented in this paper still has some limitations. For instance, in the near-field region, the model is not accurate. Thus, more accurate modelling considering near-field electromagnetic radiation model is desired in the future. Moreover, extending the model to a more general model with multiple users, wideband and non-stationary channel characteristics, is also worth studying. In addition, it would be interesting to study the capacity limit when the size of each IRS element becomes small enough such that the whole IRS can be viewed as a continuous material. Finally, how to assess the capacity lower bound of an IRS assisted communication system is also a worthy problem to study.

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\[\rho_{\text{Eam}^h i j} = \frac{1}{2\pi^2} \sqrt{\frac{A_{\text{IRSO}} \Omega c_R^2}{4\pi (\Omega + 1)}} \left( e^{-\frac{2\pi i}{\xi_{\text{Eam}}} \xi_{\text{Eam}}} \times e^{\frac{2\pi i}{\xi_{ij}} \xi_{ij}} + \right) \]
\[\frac{1}{4\pi^2} \int_{-\pi}^{\pi} \int_{-\pi}^{\pi} e^{-\frac{2\pi i}{\xi_{ij}} \xi_{ij} + \xi_{mn}} d(\alpha_{ij}^c) \, d\varphi_0 \]
\[\rho_{\text{Eam}^h i j} = \frac{1}{2\pi^2} \sqrt{\frac{A_{\text{IRSO}} G_r^2}{4\pi (\Omega + 1)}} \left( e^{-\frac{2\pi i}{\xi_{\text{Eam}}} \xi_{\text{Eam}}} \times e^{\frac{2\pi i}{\xi_{ij}} \xi_{ij}} + \right) \]
\[\frac{1}{4\pi^2} \int_{-\pi}^{\pi} \int_{-\pi}^{\pi} e^{-\frac{2\pi i}{\xi_{ij}} \xi_{ij} + \xi_{mn}} d(\alpha_{ij}^c) \, d\varphi_0 \]
\[\rho_{\text{Eam}^h i j} = \frac{1}{2\pi^2} \sqrt{\frac{A_{\text{IRSO}} G_r^2}{4\pi (\Omega + 1)}} \left( e^{-\frac{2\pi i}{\xi_{\text{Eam}}} \xi_{\text{Eam}}} \times e^{\frac{2\pi i}{\xi_{ij}} \xi_{ij}} + \right) \]
\[\frac{1}{4\pi^2} \int_{-\pi}^{\pi} \int_{-\pi}^{\pi} e^{-\frac{2\pi i}{\xi_{ij}} \xi_{ij} + \xi_{mn}} d(\alpha_{ij}^c) \, d\varphi_0 \]
\[\rho_{\text{Eam}^h i j} = \frac{1}{2\pi^2} \sqrt{\frac{A_{\text{IRSO}} G_r^2}{4\pi (\Omega + 1)}} \left( e^{-\frac{2\pi i}{\xi_{\text{Eam}}} \xi_{\text{Eam}}} \times e^{\frac{2\pi i}{\xi_{ij}} \xi_{ij}} + \right) \]
\[\frac{1}{4\pi^2} \int_{-\pi}^{\pi} \int_{-\pi}^{\pi} e^{-\frac{2\pi i}{\xi_{ij}} \xi_{ij} + \xi_{mn}} d(\alpha_{ij}^c) \, d\varphi_0 \]