Modular Extremely Large-Scale Array Communication: Near-Field Modelling and Performance Analysis

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Abstract: This paper investigates the wireless communication with a novel architecture of antenna arrays, termed modular extremely large-scale array (XL-array), where array elements of an extremely large number/size are regularly mounted on a shared platform with both horizontally and vertically interlaced modules. Each module consists of a moderate/flexible number of array elements with the inter-element distance typically in the order of the signal wavelength, while different modules are separated by the relatively large inter-module distance for convenience of practical deployment. By accurately modelling the signal amplitudes and phases, as well as projected apertures across all modular elements, we analyse the near-field signal-to-noise ratio (SNR) performance for modular XL-array communications. Based on the non-uniform spherical wave (NUSW) modelling, the closed-form SNR expression is derived in terms of key system parameters, such as the overall modular array size, distances of adjacent modules along all dimensions, and the user’s three-dimensional (3D) location. In addition, with the number of modules in different dimensions increasing infinitely, the asymptotic SNR scaling laws are revealed. Furthermore, we show that our proposed near-field modelling and performance analysis include the results for existing array architectures/modelling as special cases, e.g., the collocated XL-array architecture, the uniform plane wave (UPW) based far-field modelling, and the modular extremely large-scale uniform linear array (XL-ULA) of one-dimension. Extensive simulation results are presented to validate our findings.

Keywords: modular extremely large-scale array; practical deployment; projected apertures; non-uniform spherical wave; near-field modelling

I. INTRODUCTION

As a critical enabling technology for spectral-efficient communications, massive multiple-input multiple-output (MIMO) has been realized in the fifth-generation (5G) wireless communication networks [1–3]. To promote the ambitious process of 6G with the goals of extremely high-throughput, extremely high-reliability, and extremely low-latency [4–6], several promising transmission technologies have been studied, including extremely large-scale MIMO (XL-MIMO) [3, 7–10], Terahertz communication [11, 12], large intelligent surface (LIS) [13, 14], and intelligent reflecting surface (IRS) [15–17]. In particu-
lar, XL-MIMO that dramatically boosts antenna numbers and physical size beyond the current massive MIMO systems is expected to be able to greatly increase the system spatial resolution and spectral efficiency, which has attracted fast-growing attention recently. Besides XL-MIMO, other terms have also been adopted in the literature to refer to this technology, such as extremely large aperture arrays (ELAAs) [3], ultra-massive MIMO (UM-MIMO) [11], and extremely large aperture massive MIMO (xMaMIMO) [10].

As the array size continues to grow, two general MIMO architectures can be used to accommodate the large antenna arrays, namely the collocated extremely large-scale array (XL-array) [7, 18] and distributed antenna system (DAS) [3, 19]. Collocated XL-array is an array architecture where all antenna elements with inter-element distance typically designed at the wavelength scale are regularly deployed on a shared contiguous platform, like standard antenna arrays. With the array size going large and/or the user’s link distance becoming small, the uniform plane wave (UPW) wireless channel model is no longer valid for XL-array communications [13, 18, 20–23]. More generally, the non-uniform spherical wave (NUSW) can more accurately characterize the distance and phase variations across elements. Several preliminary efforts along such a direction have been made in [7, 8, 23–27]. For example, in [7], based on the NUSW model, a collocated extremely large-scale uniform linear array (XL-ULA) was investigated for the single-user near-field communication. In [25] and [26], the effect of distance and phase variations across all modular elements has been revealed for XL-MIMO channel estimation. By synthetically considering the varying signal amplitudes, effective projected apertures, and the polarization mismatch, the authors in [27] studied the near-field channel modelling for XL-MIMO communications. Furthermore, in [8], based on the consideration of both elevation and azimuth angles, the authors pursued a generic three-dimensional (3D) channel modelling for collocated extremely large-scale uniform planar array (XL-UPA) communications. However, since the conventional collocated XL-array architecture in [7, 8, 12, 25–27] requires a contiguous platform for deployment, the array size is typically limited by the available surface of the deployment structure in practice [3]. Different from the architecture of collocated array, DAS is an array architecture that antennas are widely distributed over a vast geographical region with multiple separated sites, which are interconnected by the backhaul/fronthaul links, so as to perform joint signal processing and guarantee ubiquitously good communication [19]. Existing representative DAS architectures include network MIMO [28], cloud radio access network (C-RAN) [29], as well as virtual MIMO [30]. Recently, a novel DAS architecture termed cell-free massive MIMO was proposed, which consists of a great many distributed access nodes to simultaneously serve the users over a wide area [31–33]. However, for DAS, massive sites are used for array deployment, which usually requires the sophisticated site coordination and high backhaul/fronthaul capacity [32, 33].

To complement the architectures of existing collocated XL-array and DAS, we study a novel modular array architecture to accommodate extremely large arrays, termed modular XL-array. As illustrated in Figure 1, all array elements in modular XL-array are regularly mounted on a shared platform with both horizontally and vertically interlaced modules, like Lego-type building blocks [34, 35]. Each module is comprised of a moderate/flexible number of array antennas with the inter-element distance typically in the order of the signal wavelength, while different modules are separated by the relatively large inter-module distance, so as to enable conformal capability with the deployment structure in practice. For example, the modular XL-array with interlaced modules can be embedded into the discontinuous wall spaced by windows, like facade circumstances of shopping malls, factories or office buildings, etc. Note that the modular design can also be applied to IRS, where high passive beamforming gain usually requires extremely large-scale surface apertures [36]. In contrast to the existing collocated XL-array, modular XL-array has the characteristic of flexible deployment. Furthermore, it may also help to improve the communication coverage [34] and spatial resolution [3]. Note that to support the large-scale deployment of modular XL-array, existing cost-effective array techniques can be employed, like analog or hybrid beamforming, as well as analog-to-digital converters (ADCs) in low-resolution, etc. On the other side, compared to the architecture of DAS, modular XL-array typically performs joint signal processing, without having to exchange or coordinate so-
phisticated inter-site information, which may ease the requirement of synchronization and reduce hardware cost associated with the backhaul/fronthaul links for DAS [35]. Furthermore, as different modules of modular XL-array share a common site, it may ease the complex network planning and site selection. However, it is worth noting that the above three array architectures fit for different application scenarios. For example, collocated and modular XL-array may be used for supporting cellular hot spot, while DAS is a good candidate for providing uniformly good services everywhere over a relatively large area [3, 35]. Therefore, the three array architectures are expected to be complementary to each other, and their choices are dependent on the practical scenarios.

The modular concept for antenna arrays was first introduced in the antenna design community [37–39]. In [37], a microstrip antenna array was designed via a modular approach, where multiple antenna modules make up a large array to achieve higher gain. In [38], a modular design of phased array antenna was exploited to reduce the array size and achieve conformal capability. In [39], a ring concentric antenna array that is composed of circular rotatable sub-arrays was designed in a modular manner, which is practically economic. However, all the aforementioned works mainly focused on the antenna designs instead of their communication channel modelling and performance analysis. To fill this gap, modular massive MIMO (mmMIMO) system was studied in [40], where modules each with a few antennas were utilized to construct a large full-dimension MIMO (FD-MIMO) system. In [41], the authors presented a flexible modular architecture that accommodates multiple antennas connected by a node, and further analysed its distributed computation complexity. By considering mmMIMO architecture with distributed antenna modules, the authors in [34] showed that the average throughput for multi-user systems could be improved than the conventional FD-MIMO. In our previous work [35], we investigated the wireless communication with modular XL-array, for which near-field NUSW channel model was developed, and the maximal SNR expression was derived in closed-form. However, that work only focused on one-dimensional (1D) modular XL-ULA architecture.

In this study, we extend our previous work on 1D modular XL-ULA [35] to the more general two-dimensional (2D) modular XL-array architecture. By accurately taking into consideration the signal phase and amplitude variations, as well as projected apertures across all modular elements, a near-field NUSW channel model for 2D modular XL-array is studied, and its achievable signal-to-noise ratio (SNR) expression is derived in closed-form. Our studied near-field channel modelling and performance analysis include the results for existing array architectures/modelling as special cases, e.g., the conventional architecture of collocated XL-array and UPW based far-field modelling. Our primary contributions are outlined as follows:

- First, we provide the mathematical modelling for modular XL-array based architecture, where XL-array antennas are modularly mounted, and different interlaced modules are separated by horizontal and vertical distances much larger than the scale of wavelength to keep conformal with the practical building facades. With the generic near-field NUSW model, we then derive the maximum achievable SNR in closed-form based on maximal-ratio combining (MRC) beamforming. This result indicates that this achievable SNR depends on the geometric parameters of modular XL-array, such as the overall modular array size, distances between adjacent modules along different dimensions, and the user’s 3D position. Addi-
II. SYSTEM MODEL

- Next, we show that our derived closed-form SNR expression includes that for the existing collocated XL-array as a special case. Furthermore, we study the asymptotic performance limit of modular XL-array. It is revealed that as modular array size goes extremely large via increasing more modules in each of the two dimensions, the corresponding SNRs approach to different upper bounds, which depend on the module spacing and effective apertures. Besides, by considering the special case of the far-field UPW assumption, we show that the corresponding SNR depends on the projected apertures of the total array, which is commonly ignored in the far-field assumption.

- Furthermore, to obtain more useful insights, we analyse a special case of 1D modular XL-ULA architecture. The result indicates that this simplified SNR of the modular XL-ULA is determined by geometric angles, called angular span and angular difference [7]. Finally, massive numerical results are presented to highlight the significance of NUSW based near-field modelling for communicating with modular XL-array.

We organize the remainder of this paper as shown below. In Section II, we introduce the mathematical model of modular XL-array. Then, in Section III, the maximal SNR expression of modular XL-array is derived in closed-form, and its performance scaling is investigated. Next, in Section IV, we provide numerical results to verify our obtained findings. Finally, we summarize this paper in Section V.

Notations: || · || is the Euclidean norm and | · | represents the absolute value of a complex number. With the given mean vector $\mathbf{0}$ and covariance matrix $\mathbf{\Sigma}$, the distribution based on a circularly symmetric complex Gaussian vector is given by $\mathcal{CN}(\mathbf{0}, \mathbf{\Sigma})$. $\mathbb{C}^{M \times N}$ represents the space of $M \times N$ complex-valued matrices.

As shown in Figure 2, we investigate an uplink communication system based on a modular XL-array mounted at the base station (BS) that communicates with a one-antenna user. For simplicity, we assume that each module is of ULA architecture. Generally, modular XL-array with the central point as the origin is located on the $y$-$z$ plane. The overall number of modules is $N = N_y N_z$, where $N_y$ and $N_z$ are the number of modules along the $y$-axis and $z$-axis, respectively. Each ULA module is equipped with $M$ antenna elements, and inter-element distance is set as $d$, which is at the scale of wavelength, e.g., $d = \frac{\lambda}{2}$, with $\lambda$ denoting the signal wavelength. Furthermore, let $D_y = K_y d$ and $D_z = K_z d$ denote the module separation distances along the $y$-axis and $z$-axis, respectively, where $K_y \geq 1$ and $K_z \geq 1$ are integers. For the special case when $K_y = 1$ and $K_z = 1$, the novel architecture of modular array reduces to the conventional collocated UPA architecture [8]. With the above notations, the physical length of each ULA module is thus denoted as $L_y = (M - 1)d$, and the physical sizes of modular XL-array along $y$–axis and $z$–axis are given by $L_y = K_y (N_y - 1) d$ and $L_z = [K (N_z - 1) + (M - 1)] d$, respectively, with $K = M + K_z - 1$.

The conventional modelling for the antenna array typically views each antenna element as a sizeless point, while we explicitly take the antenna element

Figure 2. The wireless communication with a 2D modular XL-array.
size into account, which is denoted as $\sqrt{A} \times \sqrt{A}$, where $\sqrt{A} \leq d$ [8, 27]. The effective aperture for each antenna element is given by $A_e = eA$, where $0 < e \leq 1$ denotes the aperture efficiency. Note that $A_e = eA = \frac{\sqrt{A}}{2\pi}$ holds for the hypothesis of isotropic antenna element. Denote by $\xi = \frac{A}{\pi^2} \leq 1$ the array occupation ratio that indicates the fraction of each module area occupied by the antenna elements [8].

For ease of notations, $N_y, N_z$ and $M$ are assumed as odd numbers. Therefore, the central position of the $m$th antenna element in module $(n_y,n_z)$ with $n_y = 0, \pm 1, \ldots, \pm N_y-1$, $n_z = 0, \pm 1, \ldots, \pm N_z-1$, and $m = 0, \pm 1, \ldots, \pm M-1$, can be expressed as $w_{n_y,n_z,m} = [0, y_{n_y}, z_{n_z,m}]^T$, where $y_{n_y} = n_y K_y d$ and $z_{n_z,m} = n_z [D_2 + (M - 1)d] + md = (Kn_z + md)$. Besides, let $r$ be the distance between the central position of modular XLA-array and the user, and $\theta \in [0, \pi]$ and $\phi \in [-\frac{\pi}{2}, \frac{\pi}{2}]$ represent the zenith and azimuth angles for modular array, respectively. As such, the user’s 3D position is given by $q = [r \Psi, r \Phi, r \Omega]^T$, where $\Psi \triangleq \sin \theta \cos \phi$, $\Phi \triangleq \sin \theta \sin \phi$ and $\Omega \triangleq \cos \theta$. Hence, the distance from the single user to the $m$th antenna element in module $(n_y,n_z)$ is expressed as

$$r_{n_y,n_z,m} = ||q - w_{n_y,n_z,m}|| = r \sqrt{1 + \mu(n_y,n_z,m)\epsilon + \nu(n_y,n_z,m)\epsilon^2},$$

(1)

where $\mu(n_y,n_z,m) = -2m\Omega - 2Kn_z\Omega - 2Kny\Phi$, $\nu(n_y,n_z,m) = m^2 + K' n_z^2 + 2Kmn_z + K_n^2 n_z^2$, and $\epsilon \triangleq \frac{d}{r}$. Due to antenna separation with the wavelength scale, $\epsilon \ll 1$ holds.

Let $S_{n_y,n_z,m} = [n_y K_y d - \frac{\sqrt{A}}{2}, n_y K_y d + \frac{\sqrt{A}}{2}] \times [(Kn_z + m)d - \frac{\sqrt{A}}{2}, (Kn_z + m)d + \frac{\sqrt{A}}{2}]$ denote the surface area of the $m$th antenna element in module $(n_y,n_z)$. For simplicity, we focus on the basic line-of-sight (LoS) propagation in free space. Therefore, the channel power gain from the user to the $m$th antenna element in module $(n_y,n_z)$ can be written as [8, 27]

$$g_{n_y,n_z,m}(r,\theta,\phi) =$$

$$\int_{S_{n_y,n_z,m}} \frac{1}{4\pi||q - s||^2} \frac{(q - s)^T \hat{u}_x}{||q - s||} ds,$$

(2)

where $\hat{u}_x$ means a unit vector along the positive $x$-axis that is the normal direction of array surface. In practice, the physical area of each array element $A$ is at the scale of wavelength, and hence the variations of the user’s link distance $||q - s||$ and signal direction $\frac{s - q}{||q - s||}$ within each element can be negligible. Then, we obtain the following approximated results, i.e.,

$$\frac{1}{4\pi||q - s||^2} \approx \frac{1}{4\pi\left|\left|q - w_{n_y,n_z,m}\right|\right|^2},$$

$$\frac{(q - s)^T \hat{u}_x}{||q - s||} \approx \frac{(q - w_{n_y,n_z,m})^T \hat{u}_x}{||q - w_{n_y,n_z,m}||}, \forall s \in S_{n_y,n_z,m}.$$  

(3)

As a result, the channel power gain can be rewritten as

$$g_{n_y,n_z,m}(r,\theta,\phi) \approx$$

$$\frac{eA}{4\pi\left|\left|q - w_{n_y,n_z,m}\right|\right|^2} \frac{eA (q - w_{n_y,n_z,m})^T \hat{u}_x}{\left|\left|q - w_{n_y,n_z,m}\right|\right|^2},$$

(4)

Element projected aperture

$$= \frac{eA r \sin \theta \cos \phi}{4\pi\left|\left|q - w_{n_y,n_z,m}\right|\right|^2} \frac{eA r \sin \theta \sin \phi}{4\pi\left[1 + \mu(n_y,n_z,m)\epsilon + \nu(n_y,n_z,m)\epsilon^2\right]^2}.$$  

Accordingly, the array response vector can be expressed as $a(r,\theta,\phi) \in \mathbb{C}^{(NM)\times 1}$, which contains the terms, i.e.,

$$a_{n_y,n_z,m}(r,\theta,\phi) = \sqrt{g_{n_y,n_z,m}(r,\theta,\phi)} e^{-j\frac{2\pi}{\lambda} r_{n_y,n_z,m}}.$$  

(5)

Note that the conventional far-field UPW modelling, i.e., $a_{n_y,n_z,m}(r,\theta,\phi) = \sqrt{\frac{\lambda}{4\pi} e^{-j \frac{2\pi}{\lambda} (r-n_y K_y d \Phi-(Kn_z+m)d\Omega)}}$, where $\beta_0$ is the reference channel gain at the distance of 1 m, only takes into account the variation of phase across the array elements. By contrast, the near-field NUSW modelling in (5) accurately characterizes the variations of signal phase, amplitude, and projected aperture across array elements.

We consider the uplink communication where the single-antenna user transmits the signals to the BS.
with modular XL-array. The similar analysis can also be applied to the downlink modular XL-array communications. After the receive beamforming, i.e., $v \in \mathbb{C}^{(N_M) \times 1}$ with $||v|| = 1$, the received signal obtained by the BS is formulated as

$$y = v^H a(r, \theta, \phi) \sqrt{P} s + v^H z,$$  \hspace{1cm} (6)

where $P$ and $s$ with $\mathbb{E}[|s|^2] = 1$ denote the available transmit power and the information-bearing symbol of the user, respectively. $z$ is the additive white Gaussian noise (AWGN) that follows the distribution based on a circularly symmetric complex Gaussian vector with mean vector $0$ and covariance matrix $\sigma^2 I_{NM}$, denoted as $z \sim \mathcal{CN}(0, \sigma^2 I_{NM})$.

### III. CLOSED-FORM SNR AND PERFORMANCE ANALYSIS

In this section, a new expression for the maximum SNR is derived in closed-form at first, and then a simplified SNR result is obtained when $(\theta, \phi) = (\frac{\pi}{2}, 0)$. We further study the special cases of the derived expressions, including the collocated XL-array architecture, asymptotic SNR limit, far-field assumption, as well as 1D modular XL-ULA.

Based on the formulation of signal propagation (6), the achievable SNR can be written as

$$\gamma_{\text{NUSW}} = \hat{P} |v^H a(r, \theta, \phi)|^2,$$  \hspace{1cm} (7)

where $\hat{P} = \frac{P}{\sigma^2}$ denotes the transmit SNR. Under the optimization of MRC receive beamforming, i.e., $v^* = \frac{a(r, \theta, \phi)}{|a(r, \theta, \phi)|}$, the obtained maximum SNR is $\gamma_{\text{NUSW}} = \hat{P} |a(r, \theta, \phi)|^2$. By substituting $a(r, \theta, \phi)$ into (7), the maximum SNR can be rewritten as

$$\gamma_{\text{NUSW}} = \frac{\hat{P} e A \Psi}{4\pi D_y |D_z + (M-1)d|} \times \left[ F\left(\frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} - \Omega\right) - F\left(\frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} + \Omega\right)\right] \times \left[ F\left(\frac{\bar{L}_y}{2r} + \Phi, \frac{\bar{L}_z}{2r} - \Omega\right) - F\left(\frac{\bar{L}_y}{2r} + \Phi, \frac{\bar{L}_z}{2r} + \Omega\right)\right],$$  \hspace{1cm} (9)

where $F(x, y) \triangleq \text{arcsinh} \left(\frac{x}{\sqrt{y^2 + z^2}}\right) + \frac{y}{\sqrt{y^2 + z^2}} \text{arctan} \left(\frac{xy}{\sqrt{y^2 + z^2}}\right)$, $\bar{L}_y = K y N_y d = L_y + D_y$, $\bar{L}_z = (K N_z + M)d = L_z + M d + D_z$, $\bar{L}_\epsilon = M d = L_\epsilon + d$, $\hat{L}_y = (K N_z - M)d = \hat{L}_z - 2 L_\epsilon$, and $\text{arcsinh}(x) \triangleq \ln(x + \sqrt{1 + x^2})$ is hyperbolic arcsine function. Note that $\hat{L}_y \approx L_y$ and $\hat{L}_z \approx L_z$ hold when $N_y \gg 1$ and $N_z \gg 1$, $L_\epsilon \approx L_\epsilon$ is true when $M \gg 1$, and the term $\frac{\epsilon P d r \Psi}{4\pi D_y K}$ outside the bracket of (9) can be also rewritten as $\frac{\epsilon P d r \Psi}{4\pi D_y K}$ by letting $D_z = K d$.

**Proof.** Please refer to Appendix A.

**Theorem 1.** The maximum SNR for modular XL-array communications in (8) can be obtained in closed-form as

$$\gamma_{\text{NUSW}} \approx \frac{\epsilon P d r \Psi}{4\pi D_y |D_z + (M-1)d|} \times \left[ F\left(\frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} - \Omega\right) + F\left(\frac{\bar{L}_y}{2r} + \Phi, \frac{\bar{L}_z}{2r} - \Omega\right)\right] \times \left[ F\left(\frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} + \Omega\right) + F\left(\frac{\bar{L}_y}{2r} + \Phi, \frac{\bar{L}_z}{2r} + \Omega\right)\right],$$  \hspace{1cm} (9)

where $F(x, y) \triangleq \text{arcsinh} \left(\frac{x}{\sqrt{y^2 + z^2}}\right) + \frac{y}{\sqrt{y^2 + z^2}} \text{arctan} \left(\frac{xy}{\sqrt{y^2 + z^2}}\right)$, $\hat{L}_y = K y N_y d = L_y + D_y$, $\hat{L}_z = (K N_z + M)d = L_z + M d + D_z$, $\hat{L}_\epsilon = M d = L_\epsilon + d$, $\hat{L}_y = (K N_z - M)d = \hat{L}_z - 2 L_\epsilon$, and $\text{arcsinh}(x) \triangleq \ln(x + \sqrt{1 + x^2})$ is hyperbolic arcsine function. Note that $\hat{L}_y \approx L_y$ and $\hat{L}_z \approx L_z$ hold when $N_y \gg 1$ and $N_z \gg 1$, $L_\epsilon \approx L_\epsilon$ is true when $M \gg 1$, and the term $\frac{\epsilon P d r \Psi}{4\pi D_y K}$ outside the bracket of (9) can be also rewritten as $\frac{\epsilon P d r \Psi}{4\pi D_y K}$ by letting $D_z = K d$.
indicate the variations of the modular array size $\hat{L}_y$ and $\hat{L}_z$, and the projected distances toward the array surface $r\Phi$ and $r\Omega$. Moreover, the closed-form result (9) can be widely applicable for different practical mounting structures by flexibly adjusting the module separation distances $D_y$ and $D_z$, or the number of modules $N$ and antenna elements in each module $M$.

### 3.1 Special Case with User Located on X-axis

To gain useful insights from the closed-form expression (9), we consider the following special case.

**Corollary 1.** If the user’s location is on the x-axis, i.e., $\theta = \frac{\pi}{2}$ and $\phi = 0$, the SNR expression (9) reduces to

$$\gamma_{\text{NUSW}} \left( \theta = \frac{\pi}{2}, \phi = 0 \right) = \frac{e\xi \hat{P} dr}{\pi D_y[D_z + (M - 1)d]} \times \left[ F_1 \left( \frac{\hat{L}_y}{2r}, \frac{\hat{L}_z}{2r} \right) - F_1 \left( \frac{\hat{L}_y}{2r}, \frac{\hat{L}_z}{2r} \right) \right], \tag{10}$$

where $F_1(x, y) \triangleq \arcsinh \left( \frac{x}{\sqrt{1+y^2}} \right)$ + $y \arctan \left( \frac{xy}{\sqrt{1+x^2+y^2}} \right)$.

As described in Figure 3, denote by $D_1 \sim D_5$ the vertices of one quarter of modular XL-array. We have $D_1D_2 = D_3D_4 = OD_5 = \frac{\hat{L}_y}{2}$, $OD_1 = D_2D_3 = \frac{\hat{L}_z}{2}$ and $D_1D_4 = D_2D_3 = \hat{L}_e$. An alternative expression of (10) can be equivalently written as

$$\gamma_{\text{NUSW}} \left( \theta = \frac{\pi}{2}, \phi = 0 \right) = \frac{e\xi \hat{P} dr}{\pi D_y[D_z + (M - 1)d]} \times \left[ \arcsinh (\tan \eta_2) + \tan \beta_2 \arctan (\tan \beta_2 \sin \eta_2) - \arcsinh (\tan \eta_1) - \tan \beta_1 \arctan (\tan \beta_1 \sin \eta_1) \right], \tag{11}$$

where $\eta_1 = \arctan \frac{\hat{L}_y}{\sqrt{r^2 + (\frac{\hat{L}_z}{2r})^2}}$, $\eta_2 = \arctan \frac{\hat{L}_z}{\sqrt{r^2 + (\frac{\hat{L}_y}{2r})^2}}$, $\beta_1 = \arctan \frac{\hat{L}_z}{2r}$, and $\beta_2 = \arctan \frac{\hat{L}_y}{2r}$. The expression (11) can be used to interpret the obtained SNR result in (10) in terms of geometrical relationships. Specifically, the obtained SNR depends on horizontal angular spans, i.e., $\eta_1$, $\eta_2$, and vertical angular spans, i.e., $\beta_1$ and $\beta_2$. As horizontal angular span $\eta_2$ or vertical angular span $\beta_2$ increases, the term $\arcsinh (\tan \eta_2) + \tan \beta_2 \arctan (\tan \beta_2 \sin \eta_2)$ will become large, leading to a higher SNR value. Similarly, the increase of horizontal angular span $\eta_1$ or vertical angular span $\beta_1$ will lead to smaller SNR.

### 3.2 Degeneration to Collocated XL-Array

For the special case when the module separation distances $D_y$ and $D_z$ are equal to inter-element spacing $d$, the SNR of modular XL-array given in (9) degenerates to that of conventional collocated XL-array [8], as shown in Corollary 2.

**Corollary 2.** When $D_y = D_z = d$, the SNR in (9) for modular XL-array communication reduces to

$$\gamma_{\text{NUSW}} = \frac{e\xi \hat{P}}{4\pi} \left[ G \left( \frac{N_y d}{2r} - \Phi, \frac{N_z M d}{2r} - \Omega \right) + G \left( \frac{N_y d}{2r} - \Phi, \frac{N_z M d}{2r} + \Omega \right) \right. \tag{12}$$

$$\left. + G \left( \frac{N_y d}{2r} + \Phi, \frac{N_z M d}{2r} - \Omega \right) + G \left( \frac{N_y d}{2r} + \Phi, \frac{N_z M d}{2r} + \Omega \right) \right].$$

Figure 3. The geometric relationships for modular XL-array with the user located on the x-axis.
where \( G(x, y) \triangleq \text{arctan} \left( \frac{xy}{\sqrt{x^2+y^2}} \right) \).

Proof. Please refer to Appendix B.

An useful observation of (12) is that the terms in the bracket account for the effect of the variation of the overall planar array size \( N_yd \) and \( N_zMd \), as well as the practical projected distance toward array surface \( r\Phi \) and \( r\Omega \). It is also observed from Corollary 2 that compared to the obtained SNR result for collocated XL-array that only depends on the planar array size \( N_yd \) and \( N_zMd \) [8], that for modular XL-array is associated with several key geometric parameters, including the planar physical size \( L_y \) and \( L_z \), as well as module separations along different dimensions \( D_y \) and \( D_z \). As a result, the conventional collocated XL-array that needs extremely large continuous surface is typically limited by the actual mounting structure, while module separations along different dimensions \( D_y \) and \( D_z \) render the modular architecture more flexible for practical deployment.

### 3.3 Near-Field Asymptotic SNR Scaling

As the modular XL-array size grows, the asymptotic limit of (9) is revealed in the following corollary, by considering the following three different cases.

**Corollary 3.** Case 1: As the number of modules along z-axis \( N_z \) increases indefinitely while fixing \( N_y \), we have

\[
\begin{align*}
\lim_{N_z \to \infty} \gamma_{\text{NUSW}} &= \frac{\overline{P}MeA}{2\pi D_y[(M-1)d+D_z] \times} \\
& \left[ \text{arctan} \left( \frac{\tilde{L}_y - 2r\Phi}{2r\Psi} \right) + \text{arctan} \left( \frac{\tilde{L}_y + 2r\Phi}{2r\Psi} \right) \right].
\end{align*}
\]

Case 2: As \( N_y \) increases indefinitely while fixing \( N_z \), we have

\[
\begin{align*}
\lim_{N_y \to \infty} \gamma_{\text{NUSW}} &= \frac{\overline{P}MeA}{2\pi D_y[(M-1)d+D_z] \times} \\
& \left[ \text{arctan} \left( \frac{\tilde{L}_z - 2r\Omega}{2r\Psi} \right) + \text{arctan} \left( \frac{\tilde{L}_z + 2r\Omega}{2r\Psi} \right) \right] \\
& - \frac{\overline{P}MeAr\Psi}{\pi D_y[(M-1)d+D_z]} \left( \frac{\tilde{L}_e}{(\tilde{L}_z - \tilde{L}_e - 2r\Omega)^2 + (2r\Psi)^2} \right) \\
& + \frac{\tilde{L}_e}{(L_z - \tilde{L}_e + 2r\Omega)^2 + (2r\Psi)^2}. \quad (14)
\end{align*}
\]

Case 3: As both \( N_y \) and \( N_z \) increase indefinitely, we have

\[
\begin{align*}
\lim_{N_y, N_z \to \infty} \gamma_{\text{NUSW}} &= \frac{\overline{P}}{2D_y[(M-1)d+D_z]} \frac{MeA}{\text{Module effective apertures}}. \quad (15)
\end{align*}
\]

Proof. Please refer to Appendix C.

Corollary 3 indicates that as the size of modular XL-array increases, the resulting SNR would approach to constant values, rather than increasing unboundedly. Such results are expected and comply with the energy conservation law. Specifically, it is observed that the asymptotic SNR limits in (13) and (14) are not only dependent on the module separation distances \( D_y \) and \( D_z \), and the effective apertures of each module, i.e., \( MeA \), but also determined by the user’s 3D location \( q \) and the array size \( \tilde{L}_y \) or \( \tilde{L}_z \). Instead, the asymptotic SNR limit in (15) only decreases with the module separation distances \( D_y \) and \( D_z \), while increasing with the effective apertures of each module \( MeA \). Besides, for the special case when \( D_y = D_z = d \), (15) degenerates to the asymptotic SNR for collocated XL-array, i.e., \( \gamma_{\text{collocated}} = \frac{\overline{P}}{2D_d} \) [8]. Furthermore, to explicitly show the relationship between the SNR asymptotic limits of collocated and modular XL-array architectures, the SNR ratio is defined as

\[
\Gamma = \frac{\gamma_{\text{collocated}}}{\lim_{N_y, N_z \to \infty} \gamma_{\text{NUSW}}} = \frac{K_y(K_z + M - 1)}{M}. \quad (16)
\]

Such SNR ratio is a constant value dependent on distances between adjacent modules with \( D_y = K_yd \) and \( D_z = K_zd \), as well as the number of antennas within each module \( M \).

According to the hypothesis of isotropic array elements typically spaced by the half-wavelength, i.e., \( A_e = eA = \frac{\lambda^2}{4\pi} \) and \( d = \frac{\lambda}{2} \), (15) can be simplified as

\[
\begin{align*}
\lim_{N_y, N_z \to \infty} \gamma_{\text{NUSW}} &= \frac{\overline{P}M}{2\pi K_y(K_z + M - 1)}. \quad (17)
\end{align*}
\]

### 3.4 Degeneration to Far-Field UPW Model
In the following corollary, we further study the far-field behavior of the generic SNR expression in (9).

**Corollary 4.** When \( r \Psi \gg \hat{L}_y \), \( r \Psi \gg \hat{L}_z \) and \( \frac{\hat{L}_y}{r \Psi} \), \( \frac{\hat{L}_z}{r \Psi} \ll 1 \), the SNR expression in (9) reduces to

\[
\gamma_{\text{NUSW}} \approx \gamma_{\text{UPW}} = \frac{P}{4\pi r^2} \frac{N_y N_z M eA}{\Psi} \quad \text{Projected apertures for XL-array}.
\]

(18)

**Proof.** Please refer to Appendix D.

As seen from Corollary 4, with the conventional far-field UPW assumption, the beamforming gain increases linearly and unboundedly with the overall number of modular elements, i.e., \( N_y N_z M \), which is consistent with the SNR result under the UPW assumption in [8, 27]. Corollary 4 also indicates that our derived SNR expression in (9) is general, since it can be applied to both near- and far-field communications. In addition, we can observe from (18) that under the far-field UPW assumption, the resulting SNR only depends on the projected apertures of the total modular array, i.e., \( N_y N_z M eA\Psi \), while being independent of the module separation distances \( D_y \) and \( D_z \). The reason is that the user’s projected distance along \( r \Psi \) is much longer than the array size \( \hat{L}_y \) and \( \hat{L}_z \), so that the effect of the module separation distances \( D_y \) and \( D_z \) is negligible.

By contrast, the SNR under the existing far-field UPW assumption without considering the total projected apertures is

\[
\gamma_{\text{UPW}} = \frac{P N_y N_z M}{\Psi} \beta_0,
\]

(19)

where \( \beta_0 \) represents the channel gain at a reference distance of \( r_0 = 1 \) m. We find that for the isotropic antenna elements, with \( A_e = e A = \frac{4\pi}{\lambda^2} \) and \( \beta_0 = \left( \frac{\lambda}{\Delta r} \right)^2 \), (18) differs from (19) by \( \Psi \). This shows that the UPW based far-field modelling over-estimates the SNR result in (18) as the latter takes into consideration the projected apertures since \( \Psi \ll 1 \).

### 3.5 Special Case of 1D Modular XL-ULA

To obtain more insights of the closed-form expression in (9), we consider the special case when \( N_y = 1 \) and \( K_y = 1 \), for which (9) reduces to the result of 1D modular XL-ULA.

**Corollary 5.** For a special case of modular XL-ULA, i.e., \( N_y = 1 \) and \( K_y = 1 \), the SNR expression in (9) degenerates to

\[
\gamma_{\text{NUSW,1D}} = \frac{e \xi P d \cos \phi}{4\pi [D_z + (M - 1)d] \sin \theta} \left[ \frac{H \left( \frac{L_z}{2r} - \Omega \right)}{\gamma} - H \left( \frac{L_z}{2r} - \Omega \right) + H \left( \frac{L_z}{2r} + \Omega \right) - H \left( \frac{L_z}{2r} + \Omega \right) \right] \quad \text{for } \Omega = 0.
\]

(20)

where \( H(x) \triangleq \sqrt{\sin^2 \theta + x^2} \).

**Proof.** Please refer to Appendix E.

As shown in Corollary 5, the obtained SNR depends on the geometric structure of modular XL-ULA, like the physical length \( \hat{L}_z \), the module separation distance \( D_z \), and the user’s 3D location \( q \). This result is also applicable to different discrete deployment surfaces via adjusting the module separation distance \( D_z \) accordingly. Furthermore, the resulting SNR is able to be expressed in an alternative form with the four geometric angles shown in Figure 4, i.e.,

\[
\gamma_{\text{NUSW,1D}} = \frac{e \xi P d \cos \phi}{4\pi [D_z + (M - 1)d]} \left[ \frac{\cos \alpha_3 + \cos \alpha_4}{\cos \alpha_3 \cos \alpha_4} - \frac{\cos \alpha_1 + \cos \alpha_2}{\cos \alpha_1 \cos \alpha_2} \right] \times \left[ \frac{\cos(\Delta_{s,2}) \cos(\Delta_{d,2})}{\cos(\Delta_{s,1}) + \cos(\Delta_{d,1})} - \frac{\cos(\Delta_{d,2}) \cos(\Delta_{d,2})}{\cos(\Delta_{s,2}) + \cos(\Delta_{d,2})} \right],
\]

(21)

where \( \Delta_{s,1} = \alpha_1 + \alpha_2 \), \( \Delta_{d,1} = \alpha_2 - \alpha_1 \), \( \Delta_{s,2} = \alpha_3 + \alpha_4 \), \( \Delta_{d,2} = \alpha_4 - \alpha_3 \), \( \alpha_1 = \arctan \left( \frac{L_z - r \cos \theta}{r \sin \theta} \right) \), \( \alpha_2 = \arctan \left( \frac{L_z + r \cos \theta}{r \sin \theta} \right) \), \( \alpha_3 = \arctan \left( \frac{L_z - r \cos \theta}{r \sin \theta} \right) \), and \( \alpha_4 = \arctan \left( \frac{L_z + r \cos \theta}{r \sin \theta} \right) \). Note that (21) can be easily obtained by using the trigonometric functions. This result indicates that (21) depends on the angular spans \( \Delta_{s,1} \) and \( \Delta_{s,2} \), as well as the angular differences \( \Delta_{d,1} \) and \( \Delta_{d,2} \).
Lemma 1. As $N_z$ goes infinitely, the asymptotical SNR of modular XL-ULA is given by

$$
\lim_{N_z \to \infty} \gamma_{\text{NUSW, 1D}}^{\text{NUSW, 1D}} = \frac{P \cos \phi}{2\pi[D_z + (M-1)d]r \sin \theta} M e A_{\text{Module effective apertures}}.
$$

(22)

Proof. Similar to the proof of Corollary 3, the asymptotical limit of (20) can be obtained.

Lemma 2. When $r \Psi \gg \tilde{L}_z$, the SNR of modular XL-ULA is in accordance with the SNR result based on the far-field UPW assumption, i.e.,

$$
\gamma_{\text{NUSW, 1D}} \approx \gamma_{\text{UPW, 1D}} = \frac{P}{4\pi r^2} N_z M e A_{\text{Projected apertures for XL-ULA}}.
$$

(23)

Proof. This lemma’s proof is similar to Corollary 4, which can be omitted.

Lemma 2 shows that for modular XL-ULA, the SNR under NUSW modelling can be applied to both near- and far-field communications. Besides, compared to the SNR under conventional UPW modelling without considering the effect of the practical projected apertures, i.e., $\gamma_{\text{UPW, 1D}} = \frac{P N_z M e A}{4\pi r^2}$, the asymptotical SNR in (23) also depends on the projected apertures, i.e., $N_z M e A \Psi$. However, such two SNR results $\gamma_{\text{UPW, 1D}}$ are independent of the modular separation distance $D_z$, due to the UPW assumption that user’s projected distance along $x$-axis is much longer than the array size, i.e., $r \Psi \gg \tilde{L}_z$.

IV. SIMULATION RESULTS

In this section, we present extensive numerical results to validate our obtained analytic findings for modular XL-array communications. Unless otherwise specified, the module numbers along the $y$-axis and $z$-axis are set as $N_y = 64$ and $N_z = 64$, respectively, and the number of antennas within each module is given by $M = 9$. The aperture efficiency is $e = 1$, as well as the transmit SNR is set as $P = 90$ dB. Note that while the near-field modelling and performance analysis developed in this paper are applicable to all carrier frequencies, as an example, we consider the 2.38 GHz carrier frequency in subsequent simulations. As such, the inter-element distance is denoted as $d = \frac{\lambda}{2} = 0.0628$ m. The module separation distances along two dimensions are given by $D_y = D_z = 10d = 0.628$ m, and the distance from the single user to the central position of modular XL-array is set as $r = 25$ m.

Figure 4 (a) shows the SNRs $\gamma_{\text{NUSW}}$ and $\gamma_{\text{UPW}}$ versus the overall number of modular elements $N_y N_z M$, by increasing the number of modules $N_z$ along the $z$-axis. Note that the user’s direction is $(\theta, \phi) = (60^\circ, 30^\circ)$, and the “SNR limit” shown in the figure is obtained by letting $N_y, N_z \to \infty$, corresponding to (15). As can be seen from Figure 5 (a), the closed-form SNR expression in (9) and the original summation expression in (8) perfectly match with each other, which demonstrates the accuracy of Theorem 1. When the overall number of modular elements $N_y N_z M$ is relatively small, both $\gamma_{\text{NUSW}}$ and $\gamma_{\text{UPW}}$ grow linearly with $N_y N_z M$, and this is according with Corollary 4. Nevertheless, as $N_y N_z M$ becomes large by increasing $N_z$, the SNR result under UPW modelling grows linearly unboundedly, while the SNR result under NUSW modelling approaches to a constant value. These scaling laws are in accordance with (13) and (18), respectively. It is also observed that the SNR in (19) that ignores the projected apertures across all modular elements over-estimates the SNR obtained by (18). Sim-

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ilar to Figure 5 (a), Figure 5 (b) describes the SNRs $\gamma_{\text{NUSW}}$ and $\gamma_{\text{UPW}}$ versus the number of total modular elements $N_y N_z M$ by increasing the module numbers $N_y$ along the $y$-axis. It is shown that similar observations as Figure 5 (a) can be obtained. Moreover, it is also seen from Figure 5 that the SNR limit of (15) with both $N_y$ and $N_z$ going to infinity is larger than its counterparts when only $N_y$ or $N_z$ grows, as expected.

Figure 6 plots the SNRs $\gamma_{\text{NUSW}}$ and $\gamma_{\text{UPW}}$ versus zenith angle $\theta$ by fixing the azimuth angle $\phi = 30^\circ$. It is observed that for relatively small values of $\sin \theta$, e.g., $\theta = 10^\circ$ or $\theta = 170^\circ$, the SNR under UPW modelling in (18) under-estimates the result in (9), while the reverse is true as $\sin \theta$ is large, e.g., $\theta = 90^\circ$. It is worth noting that the UPW-based result in (19) without considering the effect of projected apertures is independent of the user’s direction $\theta$, since isotropic array elements are assumed.

Figure 7 plots the SNRs $\gamma_{\text{NUSW}}$ and $\gamma_{\text{UPW}}$ versus module separation distances $D_y = D_z$ with two different zenith angles $\theta = 30^\circ$ or $\theta = 120^\circ$, by fixing azimuth angle $\phi = 30^\circ$. The inter-module separation $D_y = D_z$ starting from $\frac{\lambda}{2}$ corresponds to the conventional array with collocated architecture. It is obviously seen from Figure 7, the SNR based on farfield UPW model is independent of $D_y$ and $D_z$. By contrast, when $D_y$ and $D_z$ are relatively small, the SNRs based on NUSW model for different values of $\theta$ show different trends, while as $D_y$ and $D_z$ increase, the SNRs based on NUSW model for different values of $\theta$ exhibit the general decreasing trends. This illus-
Figure 8. SNRs with different architectures versus the overall number of modular elements.

Figure 9. SNRs for different models versus the distance r.

Figure 10. SNRs with different models for XL-ULA communication versus the overall number of modular elements.

and then the two models reach similar performance.

Last, Figure 10 shows the SNRs $\gamma_{\text{NUSW,1D}}$ and $\gamma_{\text{UPW,1D}}$ versus the overall number of modular elements $N_y N_z M$ with $N_y = 1$ and $(\theta, \phi) = (60^\circ, 30^\circ)$. It is observed from Figure 10 that for relatively small array element numbers of $N_z M$, both $\gamma_{\text{NUSW,1D}}$ and $\gamma_{\text{UPW,1D}}$ increase linearly with $N_z M$, which is consistent with the result in (23). However, as $N_z M$ further increases with $N_z$, different from the result based on UPW modelling that grows linearly and eventually becomes unbounded, the SNR result under the NUSW model approaches a limit. It is also observed that the conventional $\gamma_{\text{UPW,1D}} = \frac{P N M \beta_0}{r^2}$ by ignoring the projected apertures across modular elements overestimates the result obtained by the proposed model.
V. CONCLUSION

In this paper, we developed the channel model and conducted the performance analysis for the wireless communication with modular XL-array, where all adjacent modules have relatively large separations to keep conformal with practical building facades. By considering the signal phase and amplitude variations, as well as projected apertures across all modular elements, a generic array model is proposed. By performing the optimal MRC beamforming at the receiver, the SNR expression was derived in closed-from with respect to system parameters, such as the modular array size, distances of adjacent modules, and the user’s 3D location. Additionally, with the number of modules in two different dimensions going infinitely, the corresponding SNR asymptotic limits were obtained. Furthermore, our proposed modelling and performance analysis were shown to involve the achievable SNR results for the collocated XL-array, UPW based far-field modelling, and 1D modular XL-ULA architecture as special cases. Extensive numerical results validated the significance of NUSW based near-field modelling for accurate characterization of the modular XL-array performance, which was shown to outperform the collocated XL-array for wider coverage due to more flexible array deployment. Note that modular XL-array can be used for supporting cellular hot spot, while our theoretical derivation and analysis in this paper are only applicable for the ideal LoS propagation. There are still many practical issues, such as coupling between the user and the nearby environment, radiation pattern of transmit antenna, as well as impact of RF channel, that need to be considered in future works.

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APPENDIX

A Appendix A: Proof of Theorem 6

Similar to the derivation in [8], by observing the structure of triple summation in (8), we define the continuous function \( f(a, b, c) \) as

\[
\frac{1}{[1-2\alpha \Omega - 2Kb \Omega - 2K \nu \epsilon + \alpha^2 + K^2 b^2 + 2K ab + K^2 \epsilon^2]^2},
\]

over the domain \( A = \{ a, b, c \mid -\frac{M_\epsilon}{2} \leq a \leq \frac{M_\epsilon}{2}, -\frac{N_y \epsilon}{2} \leq b \leq \frac{N_y \epsilon}{2}, -\frac{N_y \epsilon}{2} \leq c \leq \frac{N_y \epsilon}{2}\} \).

Due to \( \epsilon \ll 1 \), we obtain \( f(a, b, c) \approx f(m \epsilon, n_x \epsilon, n_y \epsilon) \), \( \forall (a, b, c) \in [(m - \frac{1}{2}) \epsilon, (m + \frac{1}{2}) \epsilon] \times [(n_x - \frac{1}{2}) \epsilon, (n_x + \frac{1}{2}) \epsilon] \times [(n_y - \frac{1}{2}) \epsilon, (n_y + \frac{1}{2}) \epsilon] \). According to the triple integral theorem, we have

\[
\sum_{n_y=-N_y \epsilon}^{N_y \epsilon-1} \sum_{n_x=-N_x \epsilon}^{N_x \epsilon-1} \sum_{m=-M_\epsilon}^{M_\epsilon-1} f(m \epsilon, n_x \epsilon, n_y \epsilon)\epsilon^3
\]

\[
\approx \int \int \int_A f(a, b, c) \ da \ db \ dc.
\]

(A.1)

By substituting \( f(a, b, c) \) into (A.1), (A.2) can be obtained as shown below. Note that (a) and (b) hold due to the integral expressions 2.264.5 and 2.02.10 in [42], respectively, as well as (c) holds owing to the integral expressions 2.01.18 and 2.284 in [42]. We substitute (A.2) into (A.1), and then obtain the SNR result in (9). Theorem 1 is thus proved.
Due to $D_y = D_z = d$, we have $K_y = 1$ and $K = M$. Thus, the first term inside the bracket of (9) is

$$F\left(\frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} - \Omega\right)$$

$$= F\left(\frac{N_y d}{2r} - \Phi\left(M N_z + M\right) d - \Omega\right)$$

$$\approx F\left(\frac{N_y d}{2r} - \Phi, \frac{N_z d}{2r} - \Omega\right) + M d \frac{\partial F\left(\frac{N_y d}{2r} - \Phi, \frac{N_z d}{2r} - \Omega\right)}{\partial y}$$

$$= F\left(\frac{N_y d}{2r} - \Phi, \frac{M N_z d}{2r} - \Omega\right) + M d \frac{\partial F\left(\frac{N_y d}{2r} - \Phi, \frac{M N_z d}{2r} - \Omega\right)}{\partial y}$$

$$= F\left(\frac{N_y d}{2r} - \Phi\left(M N_z + M\right) d - \Omega\right)$$

$$+ M d \frac{\partial F\left(\frac{N_y d}{2r} - \Phi\left(M N_z + M\right) d - \Omega\right)}{\partial y}$$
where (a) holds due to the first-order Taylor series expansion for $y_0 = \frac{M_N d}{2r} - \Omega$ with relatively small $\frac{M d}{2r}$. Similarly, we have the approximated results for other seven terms given in the bracket of (9). After substituting all approximated results into (9), we obtain (12). Such a corollary is therefore proved.

C Appendix C: Proof of Corollary 3

1) When $N_z \to \infty$, (9) is written as the following form by adding four limit terms, i.e.,

$$
\lim_{N_z \to \infty} \gamma_{NUSW} = \frac{e^\xi Pd \Psi}{4\pi D y [D_z + (M - 1)d]} \times \left\{ \lim_{N_z \to \infty} \left[ F\left(B_1, \frac{Kd}{2r} N_z + B_3\right) - F\left(B_1, \frac{Kd}{2r} N_z - B_4\right) \right] 
+ \lim_{N_z \to \infty} \left[ F\left(B_2, \frac{Kd}{2r} N_z + B_3\right) - F\left(B_2, \frac{Kd}{2r} N_z - B_4\right) \right] 
+ \lim_{N_z \to \infty} \left[ F\left(B_2, \frac{Kd}{2r} N_z + B_4\right) - F\left(B_2, \frac{Kd}{2r} N_z - B_3\right) \right] \right\},
$$

(C.1)

where $B_1 = \frac{L_N}{2r} - \Phi$, $B_2 = \frac{L_d}{2r} + \Phi$, $B_3 = \frac{L_N}{2r} - \Omega$ and $B_4 = \frac{L_d}{2r} + \Omega$. For the first limit term inside the bracket, let $g_1(N_z) = \frac{Kd}{2r} N_z + B_3$, $g_2(N_z) = \frac{Kd}{2r} N_z - B_4$, $g_3(N_z) = \frac{Kd}{2r} N_z - \Omega$, and $g_4(N_z) = \sqrt{\Psi^2 + B_1^2 + g_3^2(N_z)}$ whose highest order is one, and thus we have (C.2), i.e.,

$$
\begin{align*}
&\lim_{N_z \to \infty} \left[ F\left(B_1, g_1(N_z)\right) - F\left(B_1, g_2(N_z)\right) \right] \\
&= \lim_{N_z \to \infty} \left[ \arcsinh\left(\frac{B_1}{\sqrt{\Psi^2 + g_1^2(N_z)}}\right) - \arcsinh\left(\frac{B_1}{\sqrt{\Psi^2 + g_2^2(N_z)}}\right) \right] \\
&+ \lim_{N_z \to \infty} \frac{g_2(N_z)}{\Psi} \left[ \arctan\left(\frac{B_1 g_1(N_z)}{\Psi \sqrt{\Psi^2 + B_1^2 + g_1^2(N_z)}}\right) - \arctan\left(\frac{B_1 g_2(N_z)}{\Psi \sqrt{\Psi^2 + B_1^2 + g_2^2(N_z)}}\right) \right] \\
&+ \lim_{N_z \to \infty} \frac{L_c}{r} \arctan\left(\frac{B_1 g_1(N_z)}{\Psi \sqrt{\Psi^2 + B_1^2 + g_1^2(N_z)}}\right) \\
&= \frac{L_c}{r} \left[ \lim_{N_z \to \infty} \frac{-B_1 g_3(N_z)}{\Psi^2 + g_3^2(N_z)} g_4(N_z) + \lim_{N_z \to \infty} \left[ \frac{\Psi^2 B_1 + B_1^2}{B_1^2 g_3(N_z) + \Psi^2 g_3^2(N_z)} g_2(N_z) \right] \right] + \frac{L_c}{r} \arctan\left(\frac{B_1}{\Psi}\right),
\end{align*}
$$

(C.2)

where (a) holds according to the first-order Taylor series expansion with relatively small $\frac{M d}{2r}$, as well as (b) follows from fractional polynomial limit theorem. Due to the similar form of four terms inside the bracket, we have (13).

2) When $N_y \to \infty$, (9) can be also given in the following form by summing four limit terms, i.e.,

$$
\lim_{N_y \to \infty} \gamma_{NUSW} \approx \frac{e^\xi Pd \Psi}{4\pi D y [D_z + (M - 1)d]} \times \left\{ \lim_{N_y \to \infty} \left[ F\left(K_y d, \frac{2r}{2r} N_y - \Phi, B_7\right) - F\left(K_y d, \frac{2r}{2r} N_y - \Phi, B_8\right) \right] \\
+ \lim_{N_y \to \infty} \left[ F\left(K_y d, \frac{2r}{2r} N_y + \Phi, B_7\right) - F\left(K_y d, \frac{2r}{2r} N_y + \Phi, B_8\right) \right] \\
+ \lim_{N_y \to \infty} \left[ F\left(K_y d, \frac{2r}{2r} N_y + \Phi, B_7\right) - F\left(K_y d, \frac{2r}{2r} N_y - \Phi, B_6\right) \right] \right\},
$$

(C.3)

where $B_5 = \frac{L_N}{2r} - \Omega$, $B_6 = \frac{L_d}{2r} - \Omega$, $B_7 = \frac{L_d}{2r} + \Omega$ and $B_8 = \frac{L_d}{2r} + \Omega$. For the first limit term inside the bracket, let $h_1(N_y) = \frac{K_y d}{2r} N_y - \Phi$, $h_2(N_y) = \sqrt{\Psi^2 + B_1^2 + h_3^2(N_y)}$ and $B_9 = \frac{L_N}{2r} - \Omega$, we have (C.4), i.e.,
\[
\begin{align*}
\lim_{N_y \to \infty} [F(h_1(N_y), B_5) - F(h_1(N_y), B_6)] \\
= \lim_{N_y \to \infty} \left[ \arcsinh \left( \frac{h_1(N_y)}{\sqrt{\Psi^2 + B_5^2}} \right) - \arcsinh \left( \frac{h_1(N_y)}{\sqrt{\Psi^2 + B_6^2}} \right) \right] + \lim_{N_y \to \infty} B_6 \left[ \arctan \left( \frac{B_3 h_1(N_y)}{\Psi \sqrt{\Psi^2 + B_5^2 + h_1^2(N_y)}} \right) \right] \\
- \arctan \left( \frac{B_6 h_1(N_y)}{\Psi \sqrt{\Psi^2 + B_5^2 + h_1^2(N_y)}} \right) \right] + \lim_{N_y \to \infty} \frac{\tilde{L}_e}{r} \arctan \left( \frac{B_3}{\Psi} \right) \\
= \frac{\tilde{L}_e}{r} \left[ \lim_{N_y \to \infty} \left( \frac{-B_3 h_1(N_y)}{\Psi^2 + B_5^2 + h_1^2(N_y)} \right) + \lim_{N_y \to \infty} \left[ \Psi^2 h_1(N_y) + h_1^2(N_y) \right] B_6 \right] + \frac{\tilde{L}_e}{r} \arctan \left( \frac{B_3}{\Psi} \right) \\
= -\frac{\tilde{L}_e^2}{2r^2(\Psi^2 + B_5^2)} + \frac{\tilde{L}_e}{r} \arctan \left( \frac{B_3}{\Psi} \right), \tag{C.4}
\end{align*}
\]

where (a) and (b) follow from conditions (a) and (b) of (C.2), respectively. Due to the same form of four terms inside the bracket, we have (14).

\[
\begin{align*}
\lim_{N_y \to \infty, N_z \to \infty} \gamma_{\text{NUSW}} \\
\approx \frac{e \xi \hat{P} d \Psi}{4\pi D_y[D_z + (M - 1)d]} \left\{ \lim_{N_y \to \infty} \left[ F \left( \frac{K_y d}{2r} N_y - \Phi, \frac{K d\tau}{2r} N_y + B_3 \right) - F \left( \frac{K_y d}{2r} N_y - \Phi, \frac{K d\tau}{2r} N_y + B_4 \right) \right] \right\} \\
+ \left\{ \lim_{N_y \to \infty} \left[ F \left( \frac{K_y d}{2r} N_y + \Phi, \frac{K d\tau}{2r} N_y + B_3 \right) - F \left( \frac{K_y d}{2r} N_y + \Phi, \frac{K d\tau}{2r} N_y + B_4 \right) \right] \right\} \\
+ \left\{ \lim_{N_y \to \infty} \left[ F \left( \frac{K_y d}{2r} N_y + \Phi, \frac{K d\tau}{2r} N_y + B_3 \right) - F \left( \frac{K_y d}{2r} N_y + \Phi, \frac{K d\tau}{2r} N_y + B_4 \right) \right] \right\}. \tag{C.5}
\end{align*}
\]

For the first limit term inside the bracket, by assuming

\[
h_3(N_y) = \sqrt{\Psi^2 + h_1^2(N_y) + g_3^2(\tau N_y)},
\]

we have

\[
\begin{align*}
\lim_{N_y \to \infty} [F(h_1(N_y), g_1(\tau N_y)) - F(h_1(N_y), g_2(\tau N_y))] \\
= \lim_{N_y \to \infty} \left[ \arcsinh \left( \frac{h_1(N_y)}{\sqrt{\Psi^2 + g_1^2(\tau N_y)}} \right) - \arcsinh \left( \frac{h_1(N_y)}{\sqrt{\Psi^2 + g_2^2(\tau N_y)}} \right) \right] \\
+ \lim_{N_y \to \infty} \frac{g_2(\tau N_y)}{\Psi} \left[ \arctan \left( \frac{h_1(N_y) g_1(\tau N_y)}{\Psi \sqrt{\Psi^2 + h_1^2(N_y) + g_1^2(\tau N_y)}} \right) \right] \\
- \arctan \left( \frac{h_1(N_y) g_2(\tau N_y)}{\Psi \sqrt{\Psi^2 + h_1^2(N_y) + g_2^2(\tau N_y)}} \right) \right] + \lim_{N_y \to \infty} \frac{\tilde{L}_e}{r} \arctan \left( \frac{h_1(N_y) g_1(\tau N_y)}{\Psi \sqrt{\Psi^2 + h_1^2(N_y) + g_1^2(\tau N_y)}} \right) \\
= -\frac{\tilde{L}_e^2}{r} \left[ \lim_{N_y \to \infty} \left[ \Psi^2 + g_3^2(\tau N_y) \right] h_3(N_y) \right] + \lim_{N_y \to \infty} \left[ \Psi^2 h_1(N_y) + h_1^2(N_y) \right] g_2(\tau N_y) \\
+ \lim_{N_y \to \infty} \left[ \Psi^2 h_1(N_y) + h_1^2(N_y) \right] g_2(\tau N_y) \right] + \frac{\pi \tilde{L}_e}{2r \Psi}, \tag{C.6}
\end{align*}
\]

where \(a\) and \(b\) follow from conditions (a) and (b) of (C.2), respectively.
where (a) holds thanks to the first-order Taylor series expansion with relatively small $\frac{L_y}{2r\Psi}$ and
$$\lim_{x,y,\rightarrow \infty} \arctan \left( \frac{xy}{\Psi \sqrt{\Psi + x^2 + y^2}} \right) = \frac{\pi}{2},$$
and (b) follows from fractional polynomial limit theorem as well. Since other three terms inside the bracket share the similar form, we have (15). Thus, Corollary 3 is completely proved.

**D Appendix D: Proof of Corollary 4**

If $r\Psi \gg \tilde{L}_y$ and $r\Psi \gg \tilde{L}_z$ hold, we can obtain $\frac{\tilde{L}_y}{2r\Psi} \ll 1$ and $\frac{\tilde{L}_z - \tilde{L}_e}{2r\Psi} \ll 1$. (9) is thus expressed as

$$\gamma_{\text{NUSW}} \approx \frac{e^\xi \tilde{P} dr \Psi}{4\pi D_y[D_z + (M - 1)d]} \times \left[ F_2 \left( \frac{\tilde{L}_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) - F_2 \left( \frac{\tilde{L}_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) \right] + \tilde{L}_e \frac{\arctan \left( \frac{L_y}{2r\Psi} - \frac{\Phi}{\Psi}, \frac{L_z - \tilde{L}_e}{2r\Psi} - \Omega \right)}{\sqrt{1 + \left( \frac{L_y}{2r\Psi} - \frac{\Phi}{\Psi} \right)^2 + \left( \frac{L_z - \tilde{L}_e}{2r\Psi} - \Omega \right)^2}} \right)$$

(D.2)

where (a) holds thanks to $\arctan x \approx x$ for $|x| \ll 1$ and $\frac{(L_y - \frac{\Phi}{\Psi})^2 + (L_z - \tilde{L}_e - \frac{\Omega}{\Psi})^2}{1 + (L_y - \frac{\Phi}{\Psi})^2 + (L_z - \tilde{L}_e - \frac{\Omega}{\Psi})^2} \ll 1$. Similarly, we can calculate the approximations of other seven terms in (D.2). Furthermore, by substituting all approximated results into (D.2), we have

$$\gamma_{\text{NUSW}} \approx \frac{e^\xi \tilde{P} M}{4\pi K K_y} \left[ F_3 \left( \frac{L_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) + \frac{\tilde{L}_e}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi} + \frac{\Omega_{\tilde{L}_z}}{\Psi} \right] + F_3 \left( \frac{L_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) + F_3 \left( \frac{L_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) + F_3 \left( \frac{L_y}{2r\Psi} - \frac{\Phi_{\tilde{L}_z}}{r^2\Psi}, \frac{\Omega_{\tilde{L}_z}}{\Psi} \right) \right]$$

(D.3)

where $F_3(x, y) = \frac{x y}{\sqrt{1 + x^2 + y^2}}$. Then, the denominator of the first term is briefly formulated as $g(u, v) \triangleq \sqrt{1 + (u - \frac{\Phi}{\Psi})^2 + (v - \frac{\Omega}{\Psi})^2}$, with $u \triangleq \frac{L_y}{2r\Psi}$ and $v \triangleq \frac{L_z - \tilde{L}_e}{2r\Psi}$. By utilizing the first-order Taylor approximation for small values $u$ and $v$, it degenerates to

$$g(u, v) \approx C_1 - \frac{\Phi}{C_1} u - \frac{\Omega}{C_1} v \quad C_1 \triangleq \sqrt{1 + \frac{\Phi^2}{\Psi^2} + \frac{\Omega^2}{\Psi^2}}$$

(D.4)

Similar to other three terms given in the bracket,
When $N_y = 1$ and $K_y = 1$, we express the first term inside the bracket of (9) as

$$\gamma_{\text{NUSW}} \approx \frac{\epsilon P M}{4\pi K K_y} \left[ F_4 \left( -\frac{\Phi}{\Psi}, -\frac{\Omega}{\Psi} \right) + F_4 \left( \frac{\Phi}{\Psi}, \frac{\Omega}{\Psi} \right) \right],$$

(D.5)

where $F_4 \triangleq \frac{(u^2+v^2)(v^2+\frac{\Phi^2}{\Psi^2})}{C_1 + \frac{\Omega^2}{C_1 \Psi^2} v^2}$. By summing $F_4 \left( -\frac{\Phi}{\Psi}, -\frac{\Omega}{\Psi} \right)$ and $F_4 \left( \frac{\Phi}{\Psi}, \frac{\Omega}{\Psi} \right)$ inside the bracket at first, we have

$$F_4 \left( -\frac{\Phi}{\Psi}, -\frac{\Omega}{\Psi} \right) + F_4 \left( \frac{\Phi}{\Psi}, \frac{\Omega}{\Psi} \right) = 2 \left( C_1 - \frac{\Omega^2}{C_1 \Psi^2} - \frac{\Phi^2}{C_1 \Psi^2} \right) uv + C_1 \frac{\Phi \Omega}{C_1 \Psi^2} - \frac{\Phi \Omega}{C_1 \Psi^2} (u^2 + v^2)
\approx 2 \left( C_1 - \frac{\Omega^2}{C_1 \Psi^2} - \frac{\Phi^2}{C_1 \Psi^2} \right) uv + C_1 \frac{\Phi \Omega}{C_1 \Psi^2} - \frac{\Phi \Omega}{C_1 \Psi^2} (u^2 + v^2)
\approx 2 \frac{C_1 - \Omega^2}{C_1 \Psi^2} C_1 \frac{\Phi \Omega}{C_1 \Psi^2} - \frac{\Phi \Omega}{C_1 \Psi^2} (u^2 + v^2)
\approx \left( \frac{\Omega^2}{C_1 \Psi^2} + \frac{\Phi^2}{C_1 \Psi^2} \right) \frac{C_1}{4\pi r^2}.
$$

(E.1)

Consequently, the proof of this corollary is done.

**Appendix E: Proof of Corollary 5**

When $N_y = 1$ and $K_y = 1$, we express the first term inside the bracket of (9) as

$$F \left[ \frac{\bar{L}_y}{2r} - \Phi, \frac{\bar{L}_z}{2r} - \Omega \right] = F \left[ \frac{d}{2r} - \Phi, \frac{\bar{L}_z}{2r} - \Omega \right]
\approx F \left[ x, \frac{\bar{L}_z}{2r} - \Omega \right] \left| \begin{array}{c} x = x_0 \\ \frac{\partial F}{\partial x} \left| x = x_0 \right. \end{array} \right.
\approx -F \left( -\Phi, \frac{\bar{L}_z}{2r} - \Omega \right)
\approx -\frac{d}{2r} \sin^2 \theta \left( \frac{\bar{L}_z}{2r} - \Omega \right)^2,$$

where (a) holds due to the first-order Taylor series expansion for $x_0 = -\Phi$ with relatively small $\frac{d}{2r}$. Similarly, the approximated results for other seven terms given in the bracket of (9) can be calculated. After substituting all approximations into (9), (20) can be obtained. Therefore, we complete the proof of Corollary 5.

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