Coverage Improvements for Sub-Terahertz Systems Under Shadowing Conditions

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Abstract — Radio propagation in the millimeter wave and sub-terahertz domain is heavily affected by shadowing conditions. The communication link is blocked without any additional technical means being used. Coverage improvements can be provided by using reflectors, RIS arrays, and repeaters to direct radio waves around corners or obstacles. These concepts show different performance and complexity levels affecting their network deployment. This paper investigates the achievable radio range or the received power to compare specific deployment concepts under realistic propagation conditions. Overall, the repeater solution provides either the largest radio range or the lowest necessary total transmit power compared to reflectors or RIS arrays and, thereby, is the most sustainable approach. A RIS array requires an additional centralized signal processing capacity for calculating optimized RIS settings and results in the highest level of network deployment complexity.

Keywords — multipath propagation, radio range, received power, reconfigurable intelligent surface (RIS), shadowing

1. Introduction

Sub-terahertz (sub-THz) mobile communication systems are currently discussed within the research community and industry as one of the technical components required for the development of sixth generation mobile networks (6G), with a view of enabling very wideband radio systems and high aggregated carrier throughput rates in the order of several hundred gigabits per second or up to a terabit per second (e.g., [1]). The wide carrier bandwidth can only be provided at very high frequencies in the sub-THz domain. However, with the increasing carrier frequency of the radio system, the propagation conditions are affected by additional atmospheric, as well as rain- and foliage-generated attenuation [2] compared with lower frequency bands. Distant-dependent pathloss and multipath propagation are encountered as well, resulting in fading (e.g., Rayleigh or Rice) and shadowing. Pathloss is particularly high in the millimeter wave (mmWave) and sub-THz domains, and the communication link is interrupted by obstacles or shadowing conditions created, for instance, by walls and buildings [2]–[5]. Therefore, technical means are needed to improve coverage and enable communication around corners in urban environments or indoor scenarios.

This paper investigates and compares the concepts of simple metallic reflectors (mirrors), reconfigurable intelligent surfaces (RIS), and repeaters to understand the performance, complexity, benefits and drawbacks for a practical network deployment. Section 2 describes the shadowing scenario; Section 3 deals with the received power required by these technical means in order to overcome shadowing, while Section 4 compares the performance of the different concepts. Results are summarized in Section 5.

2. Shadowing Propagation Scenario

In [2], terahertz systems are investigated under idealized line-of-sight (LOS) conditions to obtain a basic understanding of the achievable range and channel capacity with the additional condition of legal radiation limits to avoid health effects.

Depending on the propagation conditions, and with additional atmospheric pathloss as well as potential rain- and/or foliage-generated attenuation taken into consideration, the range is significantly limited as the frequency increases, and channel capacity is decreasing at a rapid pace [2]. These additional attenuations need to be taken into account in the higher frequency bands. Furthermore, the intended much wider system carrier bandwidths or throughput rates achieved in the sub-THz domain contribute to reducing the achievable range even further.

The transmit power or EIRP is fixed in the following scenario to meet the radiation limit of 10 W/m^2 (at a reference distance), so as to ensure compliance with legal radiation limits [6], [7]. Radio range limitations resulting from radiation limits may be overcome by higher locations at which access point or base station antennas are installed [2]. However, this does not overcome the very high pathloss encountered due to shadowing.

2.1. Propagation Measurements and Models

Pathloss measurements and models existing in the sub-THz domain are investigated, for instance, in [8], [9], and [10]. In [8], the LOS model including additional attenuation for atmospheric pathloss is confirmed for the terahertz domain. Reflections from rough surfaces, such as buildings, are also considered, as they result in additional losses. Scattering increases along with the roughness of the reflecting surface. In the sub-THz domain, the roughness of surfaces existing in the environment, such as walls, is in the order of the wavelength ([3] p. H6 and [5] pp. 26). Paper [9] confirms basically the LOS model with a slightly higher pathloss decay factor of $n > 2$. Fading statistics are investigated in [10]. For different indoor scenarios the fading statistics are modelled by $\alpha - \mu$, Rice, Nakagami-m and log-normal distributions. The best fit was achieved with the use of the $\alpha - \mu$ distribution. The Rice distribution, which is considered in this paper, is another reasonable approach.

2.2. Shadowing Propagation Scenario

Shadowing results in very high pathloss preventing data transmission in the mmWave and sub-THz domains. In addition, compared to low frequency ranges, additional attenuation components have to be considered, such as atmospheric pathloss, rain- and foliage-generated attenuation, as well as reflections and scattering depending on the propagation scenario [2]. Shadowing results in statistical fluctuations of the received power (according to a log-normal distribution) and, therefore, in a significantly higher pathloss L_{ln} compared to free space propagation [2], [4] p. 104. These fluctuations correspond to slow fading along the propagation path with a correlation length of several carrier wavelengths λ at low frequencies and many wavelengths in the mmWave and sub-THz domains.

$$\begin{aligned} L_{ln}(r) = \overline{L_{ln}(r_0)} &+ 10\, n\, \log \frac{r}{r_0} + X_\sigma \\ &+ \frac{L_r(f, r_0 = 1\text{ km})}{1\text{ km}}\; r \\ &+ \frac{L_{at}(f, r_0 = 1\text{ km})}{1\text{ km}}\; r + L_{fol}(f, r)\ [\text{dB}] \end{aligned} \tag{1}$$

where:
L_{ln} – propagation loss for log-normal fading [dB],
r – distance [m],
r_0 – reference distance [m],
$\overline{L_{ln}(r_0)}$ – average pathloss at reference distance [dB],
n – pathloss coefficient depending on the environment,
X_σ – zero mean Gaussian process [dB] with a standard deviation of σ = 4 to 10 dB, describing random shadowing depending on the environment,
L_r – rain attenuation [11] pp. 91, [12]–[14] pp. 2–68, [15],
L_{at} – atmospheric attenuation [3] p. H12, [11], [15],
L_{fol} – foliage attenuation [16].
Equation (1) is valid for a constant carrier frequency. However, in the sub-THz domain, radio propagation under shadowing conditions results in much higher pathloss than for free space propagation depending on the radio environment and, thereby, in a blocked communication link. Therefore, technical system deployment means are needed which provide quasi-LOS conditions to increase the range under shadowing conditions while simultaneously complying with legal radiation limits.

3. Pathloss with a Single Reflection and Potential Secondary Reflections Around the Receiver

The technical means deployed between the transmitter and the receiver include artificial metallic reflectors, RIS or repeaters. Additional reflections and scattering caused by the environment, e.g., building or obstacles, are also taken into consideration, as these direct electromagnetic waves around corners. In the following sections, these three technical means are investigated. For simplification purposes, only free space propagation is taken into consideration, without any atmospheric, rain- and foliage-generated attenuation.

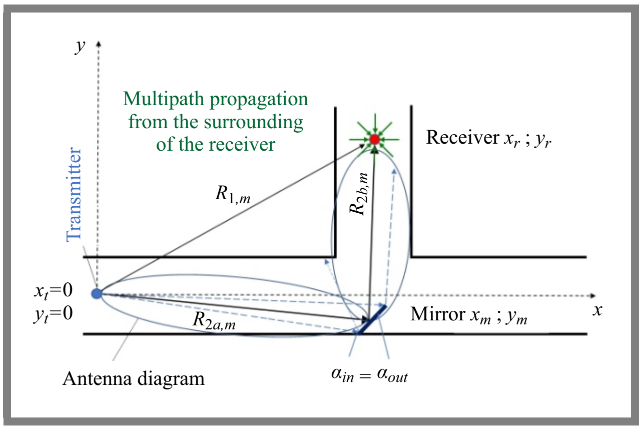

Fig. 1. Propagation scenario for indoor environments or street canyons, with the receiver not being visible from the transmitter and a reflector being used to ensure coverage.

3.1. Reflector Much Bigger Than the Wavelength

A metallic reflector with good conductivity and a physical size of $\gg \lambda$ is assumed to mitigate shadowing effects. The high conductivity of the reflector ensures that the field resistance of the metal is much lower than the field resistance of the free space. Therefore, the reflection coefficient complies with [17] pp. 134 for horizontal and vertical polarization:

$$\varrho_m \to -1\ . \tag{2}$$

The angle of the outgoing wave α_{out} is the same as that of the incoming wave α_{in}. Figure 1 shows the propagation scenario with a corner, as experienced in indoor environments or street canyons. The receiver is obstructed and is not directly visible from the transmitter. Secondary reflections from walls around the receiver are experienced as well, where $|\varrho| \ll 1$ may be assumed. Therefore, the path reflected by the reflector is much stronger than the secondary reflections (similarly as in a Rice channel). The direct path $R_{1,m}$ is obstructed, for instance by buildings, and the effective length of the reflected propagation path at the reflector (as shown in Fig. 1) is given by:

$$R_{2m} = R_{2a,m} + R_{2b,m}\ , \tag{3}$$

where the reflected path is mirrored at the reflector. Communication at mmWave or sub-THz frequency ranges is only feasible in such scenarios via the reflected path.
If the incoming wave at the reflector is a plane wave, the reflected wave is a plane wave as well – see [5] pp. 24, [17] pp. 134. Radiation density at reflector S_{2a} and field strength E_{2a} are given by the Friis formula with the field resistance of $Z_0 = 120\Omega$ [5] p. 18, where $EIRP_T = P_T\, G_T$ is limited by applicable legal radiation limits (e.g., [2]):

$$S_{2a} = P_T\, G_T\, \frac{1}{4\pi {R_{2a,m}}^2} = EIRP_T\, \frac{1}{4\pi {R_{2a,m}}^2} = \frac{{E_{2a}}^2}{Z_0}\ . \tag{4}$$

The reflected field strength E'_{2a} and radiation density S'_{2a} result from the reflection coefficient used in Eq. (2):

$$E'_{2a} = \varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi} \; E_{2a} \; , \tag{5}$$

$$\begin{aligned} S'_{2a} &= \left|\varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi}\right|^2 P_T \, G_T \, \frac{1}{4\pi {R_{2a,m}}^2} \\ &= \frac{{E'_{2a}}^2}{Z_0} = \left|\varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi}\right|^2 \frac{E^2_{2a}}{Z_0}. \end{aligned} \tag{6}$$

For $\varrho_m < |{-1}|$, the reflector may add attenuation to the propagation loss. This results in a specific relation between the received and transmitted powers, such as:

$$\begin{aligned} \left.\frac{P_R}{P_T}\right|_{reflector} &= \left|\varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi}\right|^2 G_T \, \frac{1}{4\pi {R_{2,m}}^2} \, A_R \\ &= \left|\varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi}\right|^2 G_T \, \frac{1}{4\pi (R_{2a,m}+R_{2b,m})^2} \, A_R \\ &= \left|\varrho_m \, \mathrm{e}^{-\mathrm{j}\Delta\phi}\right|^2 G_T \, G_R \left(\frac{\lambda}{4\pi R_{2,m}}\right)^2 \alpha \, \frac{1}{(R_{2a,m}+R_{2b,m})^2} \, . \end{aligned} \tag{7}$$

or with Eq. (2):

$$P_R\Big|_{reflector} \approx EIRP_T \, G_R \left(\frac{\lambda}{4\pi \; (R_{2a,m}+R_{2b,m})}\right)^2 . \tag{8}$$

The received power depends on the square of the path length $R_{2m} = R_{2a,m} + R_{2b,m}$, the absolute value of the reflection coefficient ϱ_m, and the transmitter and receiver antenna gain. In the case of a metallic reflector, $\varrho_m = |1|$, is a realistic estimate. Figure 1 also indicates that secondary reflections around the receiver provide additional multipath propagation. A metallic reflector is characterized by the following implementation-critical features:

- it is passive device,
- no power supply or data connections are needed,
- it is only mounted at buildings, towers or lamp posts,
- it is a low complexity solution,
- the reflector may have the form of a plane, may be concave or convex, with a fixed, static beam shaped to suit the area to be covered,
- the reflector improves coverage, especially in shaded areas,
- no latency is added to the signal's propagation,
- for high conductivity of the reflector $\varrho_m \approx |1|$, nearly no insertion loss is added,
- the propagation corresponds to quasi-LOS propagation conditions around corners,
- it ensures a much longer range compared with direct propagation with high shadowing pathloss,
- LOS propagation provides a strong direct component and thereby establishes a Rice channel at the receiver,
- without the reflector, the received power would be much lower and the channel would correspond to a Rayleigh channel,
- at the receiver, delay spread is generated via secondary reflections around the receiver or from other reflecting obstacles present in the deployment area, with smaller reflection coefficients.

3.2. Reflector Smaller or in the Same Order as Wavelength and RIS Arrays

A RIS array comprises many small RIS elements with spacing between them equaling approx. $\lambda/2$ of the medium carrier frequency of the frequency range in use. As in the antenna theory for array antennas with a group of antenna elements, it is assumed that the diagrams of the different elements are independent from each other ([18] pp. 254) to apply the superposition principle.

The radar cross section σ of an RIS element is modeled in this paper by a metallic sphere with a radius r as an idealized approach [19]–[22]:

$$\begin{aligned} \frac{\sigma}{\pi r^2} &= \frac{1}{k\,r} \sum_{n=1}^{\infty} (-1)^n \; (2n+1) \\ &\cdot \left[\left(\frac{k\,r J_n(k\,r) - n J_n(k\,r)}{k\,r H_{n-1}(k\,r) - n H^1_n(k\,r)} \right) - \left(\frac{J_n(k\,r)}{H^1_n(k\,r)} \right) \right] , \end{aligned} \tag{9}$$

where: r is the radius of the sphere, $k = \frac{2\pi}{\lambda}$, λ is the wavelength, J_n is the spherical Bessel function of the first kind of order n, H^1_n is the Hankel function of order n, described by $H^1_n(k\,r) = J_n(k\,r) + j\,Y_n(k\,r)$, and Y_n is spherical Bessel function of the second kind of order n.

A RIS inherent insertion loss [23] is not considered here.

Paper [24] indicates that a signal that is reflected or scattered in other directions than the incoming wave has a lower field strength than the reflection back to the source. Therefore, the radar cross section used below is an optimistic approach to reflections in other directions than that of the incoming wave. For a spacing of $\lambda/2$ between RIS elements, radius r of the metallic sphere and its normalized size equal:

$$r = \frac{\lambda}{4} \; , \quad \frac{2\pi\, r}{\lambda} = \frac{2\pi \frac{\lambda}{4}}{\lambda} = \frac{\pi}{2} > 1 \tag{10}$$

Based on Eqs. (9) and (10), the normalized radar cross section is approximately:

$$\frac{\sigma}{\pi\, r^2} = 1 \tag{11}$$

Reflections from or scattering caused by small objects with a size $< \lambda$ result in spherical waves. For a single sphere, the ratio between the received and transmitted power is given in [3], p. S1 as well as Eqs. (10) and (11) from the radar equation:

$$\begin{aligned} \left.\frac{P_R}{P_T}\right|_{scatter,sphere} &= \frac{G_T}{4\pi \; {R_{2a,s}}^2} \; \sigma \; \frac{A_e}{4\pi \; {R_{2b,s}}^2} \\ &= G_T \; G_R \; \pi \left(\frac{\lambda}{4}\right)^2 \frac{1}{(4\pi)^3} \; \frac{\lambda^2}{(R_{2a,s} \; R_{2b,s})^2} \end{aligned} \tag{12}$$

or by considering the limited $EIRP_T = P_T \; G_T$ at the base station with respect to the legal radiation limits (e.g., [2]):

$$\begin{aligned} P_R\Big|_{scatter,sphere} &= EIRP_T \; G_R \; \pi \left(\frac{\lambda}{4}\right)^2 \\ \frac{1}{(4\pi)^3} \; \frac{\lambda^2}{(R_{2a,s} \cdot R_{2b,s})^2} \; &\alpha \; \frac{1}{(R_{2a,s} \cdot R_{2b,s})^2} \quad . \end{aligned} \tag{13}$$

The pathloss is proportional to the product of the square of the two partial distances $R_{2a,s}$ and $R_{2b,s}$.

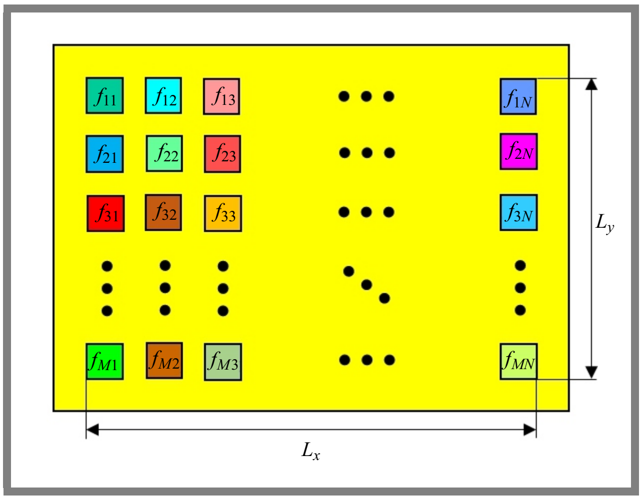

Fig. 2. RIS array made up of $M \cdot N$ RIS elements.

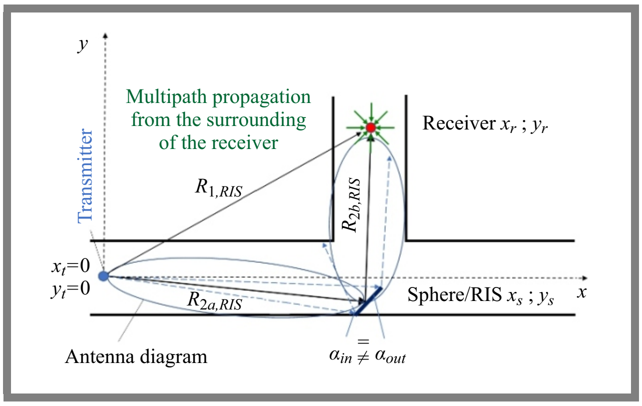

Fig. 3. Propagation scenario for indoor environments or street canyons, with the receiver not being visible from the transmitter and a sphere or an RIS array being used to ensure coverage.

A RIS array comprises many small reflectors $(M \cdot N)$ shown in Fig. 2. For simplicity, it is assumed that the different RIS elements do not influence each other. Then, the solution may be regarded as a phased array antenna which is illuminated by the base station or a user equipment antenna. The radar cross section of a single sphere is extended by the physical size of the RIS array $(M \cdot N)(\frac{\lambda}{2})^2$. A RIS array can be used for beamforming with the additional beamforming gain of $(M \cdot N)$, where compared to Subsection 2.1, the outgoing angle α_{out} is not necessarily the same as the incoming angle α_{in}. A propagation scenario similar to that shown in Fig. 1 is considered in Fig. 3. In the scenario in which phase shifts of the different RIS elements are random, the sum of the powers of the different reflected signals per RIS element is received. Then Eq. (12) is extended as in Eq. (14):

$$\left.\frac{P_R}{P_T}\right|_{RIS} = \frac{G_T}{4\pi {R_{2a,RIS}}^2} \, M \cdot N \, \sigma \frac{A_e}{4\pi {R_{2b,RIS}}^2}$$
$$= G_T \, G_R \, M \cdot N \, \pi \left(\frac{\lambda}{4}\right)^2 \frac{1}{(4\pi)^3} \frac{\lambda^2}{(R_{2a,RIS} \, R_{2b,RIS})^2} \, . \quad (14)$$

By relying on appropriate phase shifts, each element in the RIS array may be used for beamforming. The constructive superposition of each reflected signal provides a main lobe of the resulting beam from the RIS array towards the receiver antenna. Under ideal conditions, the receiver antenna is summing up the field strength of each signal. This corresponds to a group of antennas with many individual elements ([18] pp. 254). With the multiplicative approach, where the directivity of an individual antenna element is multiplied with the characteristic of the array and all antenna elements have the same characteristic and same distance $\lambda/2$ between the antenna elements like in Fig. 2, the maximum ratio between P_R and P_T follows with the array gain $M \cdot N$:

$$\left.\frac{P_R}{P_T}\right|_{RIS,max} = \frac{G_T}{4\pi \, {R_{2a,RIS}}^2} \, (M \cdot N)^2 \, \sigma \frac{A_e}{4\pi \, {R_{2b,RIS}}^2}$$
$$= G_T \, G_R \, (M \cdot N)^2 \, \pi \left(\frac{\lambda}{4}\right)^2 \frac{1}{(4\pi)^3} \frac{\lambda^2}{(R_{2a,RIS} \, R_{2b,RIS})^2} \quad (15)$$

or

$$P_R\big|_{RIS,max} = EIRP_T \, G_R \, (M \cdot N)^2 \, \pi \left(\frac{\lambda}{4}\right)^2 \frac{1}{(4\pi)^3}$$
$$\cdot \frac{\lambda^2}{(R_{2a,RIS} \, R_{2b,RIS})^2} \; \alpha \; \frac{1}{(R_{2a,RIS} \, R_{2b,RIS})^2} \quad . \quad (16)$$

The received power P_R is growing with the square of the RIS array size $M \cdot N$. With a growing number of RIS elements, the received power is increasing.

Pathloss is proportional to the product of the square of the two partial distances $R_{2a,RIS}$ and $R_{2b,RIS}$. $EIRP_T$ is limited by applicable radiation limits. P_R in Eqs. (13) or (16) corresponds to the direct reflected path. Figure 3 also indicates that secondary reflections around the receiver provide additional multipath propagation, where a reflection coefficient $|\varrho| \ll 1$ can be assumed.

A RIS array is characterized by the following technical implementation features:

- it is basically a passive device without amplification and without signal regeneration capabilities,
- by using appropriate phase shifters for different RIS elements, some beamforming is possible,
- this requires that signal processing performed by user equipment, the base station and the RIS array be centralized [25]. In this sense, it is an active system,
- a power supply is needed,
- a data connection from/to the central signal processing unit at the base station and the RIS feedback to the base station/signal processing unit is required [25],
- it is mounted on buildings, towers or lamp posts,
- with the power supply and an additional data connection, this is a very complex solution,
- no latency is added to the signal's propagation, because RIS elements only provide a phase shift,
- RIS improves coverage, especially in shaded areas,
- propagation corresponds to quasi-LOS propagation conditions around corners,
- it offers a longer range compared to direct propagation with the high shadowing pathloss,
- LOS propagation provides a strong direct component and thereby a Rice channel is established at the receiver,

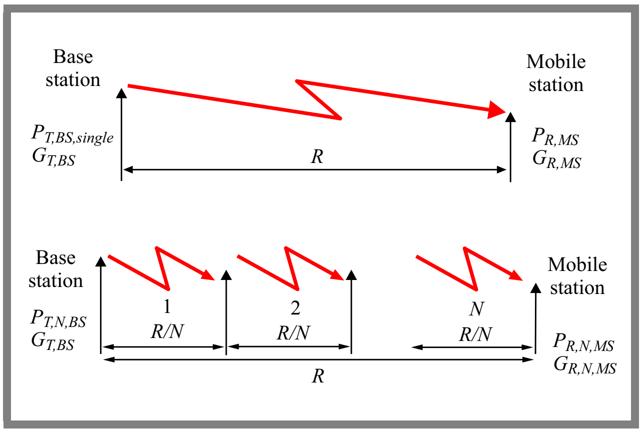

Fig. 4. Operation principle of a repeater-based or multi hop solution: a) single hop approach, b) multi hop or repeater-based approach.

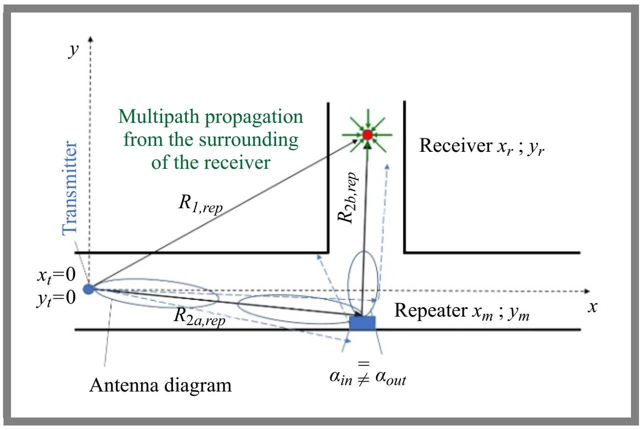

Fig. 5. Propagation scenario for indoor environments or street canyons with the receiver not visible from the transmitter's location and a repeater used to ensure coverage.

– without a RIS array, the received power would be much lower and the channel would correspond to a Rayleigh channel,
– the delay spread is generated, at the receiver, via secondary reflections around the receiver or from other reflecting obstacles present in the deployment area and characterized by smaller reflection coefficients,
– for a specific user location, the radio channel can be influenced or adjusted. However, RIS settings are strongly dependent on the receiver's location,
– if RIS is used for beamforming, it operates in a manner that is similar to a directive antenna,
– with an increased number of RIS elements $M \cdot N$ the RIS array gain and its physical size increase as well, in practice, the received power will be lower due to the inherent RIS insertion loss.

3.3. Repeater

In a solution with a repeater, the entire propagation path R is subdivided into shorter propagation paths (Fig. 4). In this simplified example, all transmit powers and antenna gains as well as the N radio path lengths are the same per hop. This means that the transmit power at the base station can be reduced for the same effective path length, as described in Subsection 3.1 and 3.2, and/or such an approach allows to transmit around corners under shadowing conditions, while achieving longer range. In the following, free space propagation per hop is assumed. In the single hop case, the received power P_R follows from the link budget in Eq. (17) and the necessary transmit power $P_{T,BS,single}$ from Eq. (18):

$$P_R = P_{T,BS,single}\, G_{T,BS}\, G_{R,MS} \left(\frac{\lambda}{4\pi R}\right)^2 , \tag{17}$$

$$P_{T,BS,single} = \frac{P_R}{G_{T,BS}\, G_{R,MS}} \left(\frac{4\pi R}{\lambda}\right)^2 . \tag{18}$$

In the multi hop or repeater-based case, it is assumed that the received power P_R at the destination is the same as in the single hop case. Then, the necessary base station and repeater transmit powers $P_{T,BS,multi\ hop}$ and $P_{T,MS,N}$ are assumed to be the same:

$$\begin{aligned} P_{T,total} &= N\, P_T = N \frac{P_R}{G_T\, G_R} \left(\frac{4\pi R}{\lambda\, N}\right)^2 \\ &= \frac{1}{N} \frac{P_R}{G_T\, G_R} \left(\frac{4\pi R}{\lambda}\right)^2 = \frac{P_{T,BS,single}}{N} . \end{aligned} \tag{19}$$

It follows from Eq. (19) that in the multi hop approach, the total needed transmit power is lower than that in the single hop scenario, and for the same distance R needs to be covered. It is more challenging, especially at very high frequencies, to generate high RF power levels. Therefore, the multi-hop solution makes millimeter wave and terahertz systems more suitable and allows to transmit around corners with lower energy consumption.
In the repeater-based solution, a propagation solution comparable with that shown in Fig. 1 and Fig. 3 is assumed (and shown Fig. 5). The repeater has an antenna looking towards the base station/access point and a second antenna pointing towards the intended coverage area which is shaded from the base station. In addition, the repeater provides an amplification from the received to the transmit signal and may perform signal regeneration and beamforming or MIMO transmission. The outgoing angle α_{out} is not necessarily the same as the incoming angle α_{in}. The link between the base station/access point and the repeater as well as the link between the repeater and the user equipment is given by the Friis formula [5], p. 18 and the repeater power amplification V, such as:

$$\begin{aligned} &\frac{P_{R,repeater}}{P_{T,base\ station}} = \\ &\quad G_{T,base\ station}\, G_{R,repeater} \left(\frac{\lambda}{4\pi\, R_{2a,rep}}\right)^2 \\ &\quad \frac{P_{R,user\ equipment}}{P_{T,repeater}} = \frac{P_{R,user\ equipment}}{V \cdot P_{R,repeater}} \\ &= G_{T,repeater}\, G_{R,user\ equipment} \left(\frac{\lambda}{4\pi\, R_{2b,rep}}\right)^2 . \end{aligned} \tag{20}$$

In the repeater-based solution, it is assumed that the base station and the repeater can generate the maximum $EIRP_T$ values allowed under applicable radiation limits. It can also be assumed that the minimum necessary received power P_R or receiver sensitivity are the same for the base station, repeater and user equipment. Due to the amplification performed in the repeater, the received power at the user equipment depends solely on the $EIRP_T$ of the repeater and the distance $R_{2b,rep}$.

The amplification V of the repeater is limited, so that the maximum allowed $EIRP_T$ based on radiation limits is complied with:

$$P_{T,repeater}\ G_{T,repeater} = V\ P_{R,repeater}\ G_{T,repeater} = EIRP_T\ , \quad (21)$$

$$V = \frac{P_{T,repeater}}{P_{R,repeater}} = \frac{EIRP_T}{P_{R,repeater}\ G_{T,repeater}}\ , \quad (22)$$

which provides the received power at the user equipment:

$$P_{R,user\ equipment} = EIRP_T\ G_{R,user\ equipment}\left(\frac{\lambda}{4\pi\ R_{2b,rep}}\right)^2 . \quad (23)$$

Equation (23) provides the received power at the user equipment. The total maximum transmit power necessary to cover the entire path length $R_{2,rep} = R_{2a,rep} + R_{2b,rep}$ is given as:

$$P_{T,base\ station} + P_{T,repeater} = \frac{EIRP_T}{G_{T,base\ station}} + \frac{EIRP_T}{G_{T,repeater}} . \quad (24)$$

Each of the two hops can cover, with the same $EIRP_T$, the same distance as achieved in the reflector-based solution from Subsection 2.1. Therefore, either twice the effective distance is feasible or the transmit power used at the base station and the repeater can be reduced.

The received power P_R from Eq. (23) corresponds to the direct reflected path. Figure 5 also indicates that secondary reflections around the receiver provide additional multipath propagation.

The repeater-based approach is characterized by the following technical implementation-related features:

- it is an active device with amplification and potentially with signal regeneration and beamforming or a MIMO transmission,
- there are basically two repeater options available.

In the first scenario an RF repeater is used only, and in the other the repeater regenerates the received signal by means of signal processing and beamforming or a MIMO transmission. In the first case RF amplification is applied only to achieve the maximum allowed $EIRP_T$, potentially with a shift to a different carrier frequency to avoid oscillation of the repeater due to positive feedback from the repeater's output, affecting its input. Repeater amplification and antenna gain can either increase the covered range compared to a reflector or a RIS, or may allow to reduce the overall RF power needed. This approach adds almost no latency, if analogue signal processing or amplification is applied only. The solution offers a medium complexity level.

In the second case, the repeater amplifies the signal to the maximum allowed $EIRP_T$. Advanced antenna systems, such as those relying on beamforming or MIMO concepts, can be used and the repeater's amplification and the antenna gain may either increase the covered range compared to a reflector or a RIS or allow to reduce the overall RF power required.

No additional feedback from the user equipment or the base station to the repeater is needed, as is the case on the RIS approach. The channel estimation capabilities of the radio interface are sufficient to implement such antenna concepts. However, signal processing for signal regeneration and antenna algorithms is adding latency. It is a solution with a medium level of complexity, since the same microelectronic components as those used in base stations or user equipment may be applied:

- only a power supply is needed,
- it is mounted on buildings, towers or lamp posts,
- overall, its complexity level is between that of a metallic reflector and a RIS system,
- a repeater is more flexible than a metallic reflector in directing energy at specific users and a RIS by using advanced antenna concepts,
- a repeater is improving coverage significantly, especially in shaded areas,
- the propagation pattern corresponds to LOS propagation conditions between the base station and the repeater and a second LOS propagation between the repeater and user equipment to direct the signal around corners,
- this allows a bigger range compared to direct propagation with a high shadowing pathloss,
- LOS propagation provides a strong direct component and thereby a Rice channel is established at the receiver,
- without a repeater, the received power would be much lower and the channel would correspond to a Rayleigh channel,
- delay spread is generated via secondary reflections around the receiver or other reflecting obstacles in the deployment area characterized by smaller reflection coefficients.

4. Comparison of reflector, RIS array and repeater

The different solutions are compared under the same conditions. All concepts are using:

- the same maximum $EIRP_T$ at the base station and also at the repeater,
- equal receiver antenna gains at user equipment,
- the same necessary received power at user equipment and at the repeater and
- the same carrier frequency or wavelength.

The size of the RIS array $M \cdot N$ is variable.

Based on the formulas for the received power at user equipment for the three system concepts involving a reflector Eq. (8), a RIS array for optimal beamforming Eq. (16) and a repeater Eq. (23) the achievable received power for the same radio path length is investigated. For comparison, the results are normalized to the reflector case. This results in the following normalized functions:

$$\frac{P_R\big|_{RIS,max}}{P_R\big|_{reflector}} = \frac{(M \cdot N)^2 \, \pi \left(\frac{\lambda}{4}\right)^2 \frac{1}{(4\pi)^3} \frac{\lambda^2}{(R_{2a,RIS} R_{2b,RIS})^2}}{\left(\frac{\lambda}{4\pi(R_{2a,m}+R_{2b,m})}\right)^2}$$
$$= (M \cdot N)^2 \frac{1}{4^3} \frac{\lambda^4}{(R_{2a,RIS} \, R_{2b,RIS})^2} \frac{(R_{2a,m}+R_{2b,m})^2}{\lambda^2} . \quad (25)$$

$$\frac{P_R\big|_{repeater}}{P_R\big|_{reflector}} = \frac{\left(\frac{\lambda}{4\pi R_{2b,rep}}\right)^2}{\left(\frac{\lambda}{4\pi(R_{2a,m}+R_{2b,m})}\right)^2}$$
$$= \frac{(R_{2a,m}+R_{2b,m})^2}{{R_{2b,rep}}^2} . \quad (26)$$

If the same geometry and distances are assumed with:

$$R_{2a,m} = R_{2a,RIS} = R_{2a,rep} = R_{2a} , \quad (27)$$

$$R_{2b,m} = R_{2b,RIS} = R_{2b,rep} = R_{2b} . \quad (28)$$

Eqs. (25) and (26) result in:

$$\frac{P_R\big|_{RIS,max}}{P_R\big|_{reflector}} = (M \cdot N)^2 \frac{1}{4^3} \frac{\lambda^4}{(R_{2a} R_{2b})^2} \frac{(R_{2a}+R_{2b})^2}{\lambda^2} , \quad (29)$$

$$\frac{P_R|_{repeater}}{P_R|_{reflector}} = \frac{(R_{2a}+R_{2b})^2}{R_{2b}^2} = \frac{R_2^2}{R_{2b}^2} . \quad (30)$$

4.1. Received Power Ratio for the RIS Array vs. Reflector Approach

With the normalized distant-dependent terms R_{2a}/λ, R_{2b}/λ and the total path length $R_2/\lambda = R_{2a}/\lambda + R_{2b}/\lambda$ it follows for Eq. (29):

$$\frac{P_R\big|_{RIS,max}}{P_R\big|_{reflector}} = (M \cdot N)^2 \frac{1}{4^3} \frac{\lambda^2}{R_2^2} \frac{1}{(\frac{R_{2b}}{R_2})^2 (1-\frac{R_{2b}}{R_2})^2} , \quad (31)$$

with $0 \leqslant R_{2b}/R_2 \leqslant 1$. As the RIS array will not be placed directly at the base station location ($R_{2b}/R_2 = 1$) or at the location of user equipment ($R_{2b}/R_2 = 0$) with poles in Eq. (31), a reasonable assumption for the RIS array location is somewhere in the middle of the overall propagation path within the following limits

$$0.2 \leqslant \frac{R_{2b}}{R_2} \leqslant 0.8 . \quad (32)$$

The minimum size of the RIS array $M \cdot N|_{min}$ for the same received power as for the reflector follows from Eq. (31):

$$\frac{P_R\big|_{RIS,max}}{P_R\big|_{reflector}} = 1 , \quad (33)$$

$$M \cdot N\big|_{min} = 8 \frac{R_2}{\lambda} \frac{R_{2b}}{R_2} \left(1 - \frac{R_{2b}}{R_2}\right) . \quad (34)$$

Due to the huge variation versus carrier frequency f (1, 3, 10, 30, 100, and 300 GHz) and entire path length R_2 (1, 10, 100, and 200 m) the evaluation of Eq. (34) is presented logarithmically as $\log(M \cdot N|_{min})$ in Fig. 6 versus the relative location R_{2b}/R_2 of the reflector or the RIS array. The minimum RIS array size increases significantly with the carrier

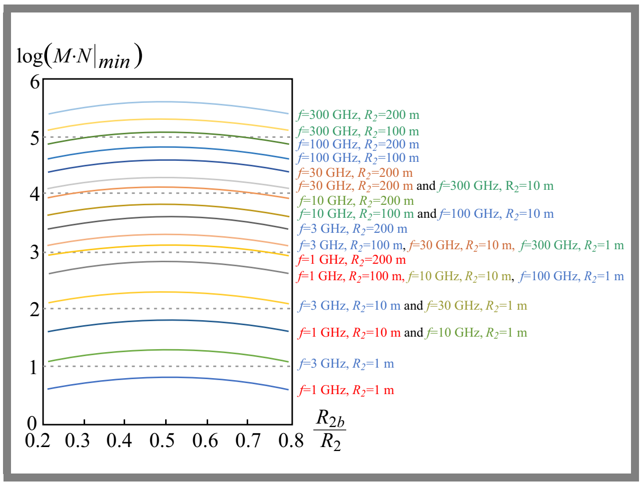

Fig. 6. Summary diagram of the minimum size of the RIS array $M \cdot N|_{min}$ Eq. (34) for optimum phase shifts per RIS element versus carrier frequency f, entire path length R_2 versus the relative location R_{2b}/R_2 of the reflector or RIS array.

frequency and the entire path length. Especially, in the millimeter or sub-Terahertz frequency range, the number of RIS elements is in the order of several tens of thousands and more. Here, it is assumed that the maximum gain $M \cdot N$ of the RIS array is available with appropriate phase shifts of the RIS elements.

If there is no RIS $gain = M \cdot N = 1$ in the case of random phase shifts per RIS element, the minimum number of RIS elements needs to grow by the square of $M \cdot N|_{min}$ in Eq. (34):

$$\frac{P_R\big|_{RIS,no\ RIS\ gain}}{P_R\big|_{reflector}} = M \cdot N \frac{1}{4^3} \frac{\lambda^2}{R_2^2} \frac{1}{(\frac{R_{2b}}{R_2})^2 \, (1-\frac{R_{2b}}{R_2})^2} , \quad (35)$$

$$M \cdot N\big|_{min,no\ RIS\ gain} = \left[8 \frac{R_2}{\lambda} \frac{R_{2b}}{R_2} \left(1 - \frac{R_{2b}}{R_2}\right) \right]^2 . \quad (36)$$

In practice, the necessary RIS size will be within the limits of Eqs. (34) and (36):

$$\log(M \cdot N|_{min}) \leqslant \log(M \cdot N)$$
$$\leqslant \log(M \cdot N|_{min,no\ RIS\ gain}) = 2\log(M \cdot N|_{min}) . \quad (37)$$

The signal processing capacity required to adjust the different RIS elements will be very complex. For smaller RIS sizes, the received power at the user equipment will be lower than for a reflector.

4.2. Received Power Ratio for the Repeater vs. Reflector Approach

In the scenario involving a repeater, the ratio of the received power between the repeater and reflector concepts depends solely on the ratio of the entire path length R_2 and the relative location of the reflector or repeater R_{2b}, as long as the path length R_{2b} is short enough to ensure sufficient received power at the repeater (Fig. 7):

$$\frac{P_R|_{repeater}}{P_R|_{reflector}} = \frac{(R_{2a}+R_{2b})^2}{R_{2b}^2} = \frac{R_2^2}{R_{2b}^2} . \quad (38)$$

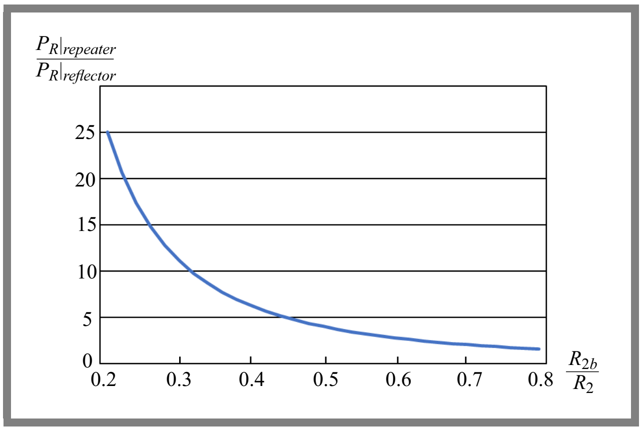

Fig. 7. Ratio of the power received by the repeater, normalized to a reflector solution, and the relative location of the reflector or repeater R_{2b}.

Equation (38) is valid under the condition that all transmitters have a radiated equivalent power $EIRP_T$. The repeater solution always provides a higher received power at the user equipment than the reflector. This gain is decreasing if the repeater is located closer to the base station. In that case, the length of the path of the second hop R_{2b} is similar to that of the entire path R_2 in the reflector scenario. If the repeater is located in the middle of the entire radio path R_2, the transmit power necessary at the base station and at the repeater can be reduced for two hops $N = 2$ by a factor of $N^2 = 4$ – see Subsection 2.1.

5. Conclusions

In the millimeter wave and sub-terahertz frequency domains, pathloss is very high compared to lower frequency ranges (Section 1). Especially, under shadowing conditions, radio communication is basically blocked and without deploying special means, reflected multipath signals excessively lower the received power, since the roughness of surfaces characterized by a low reflection coefficient (e.g. walls walls) is in the order of the wavelength. Therefore, for practical applications, shadowing needs to be overcome by technical means which are directing radio signals around corners. The following options are considered:

- reflector or metallic mirror (Subsection 3.1),
- RIS array (Subsection 3.2),
- repeater (Subsection 3.3).

These technical means create a strong component in the channel impulse response at the receiver and thereby result in the establishment of a Rice channel.

From the comparison presented in Section 4 and Tables 1–2, it can be concluded that:

- A repeater is the best overall solution with a medium degree of complexity. The same electronic components as those used in base stations/access points and user equipment may be relied upon and only a power supply is needed. The overall transmit power is lower than for the other means. Therefore, this is the most sustainable solution;
- A reflector or a metallic mirror is a cheap alternative with low complexity. It does not require a power supply or an additional data connection;
- A RIS array requires a huge number of RIS elements and is the most complex solution in terms of signal processing. It also requires an additional power supply and a data connection to ensure centralized signal processing for RIS settings and to enable fast updates for moving users.

Especially for very wideband systems and in the millimeter wave and sub-terahertz domains, such as like 6G a Rice channel provides a significantly increased coherence bandwidth or reduced frequency dependency of the radio channel and increased received power compared to Rayleigh channels under shadowing conditions. This results in reduced effective intersymbol-interference (significantly more open eye-diagram), which reduces or avoids the need for time-domain equalization of the received signal.

In the scenarios involving a repeater or a reflector, coverage at high frequency ranges required by 6G systems may be improved significantly while maintaining reasonable complexity levels.

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Tab. 1. Comparison of features

Topic	Reflector	RIS array	Repeater
Mounting at	Buildings, towers or lamp posts		
Power supply	No power supply	Power supply needed	
Data connection to the network	No additional data connection	Additional data connection, deployment difficult and expensive	No additional data connection
Active/passive device	Passive	Active due to signal processing for beamforming	Active: amplification, potential signal processing for signal regeneration and beamforming
Overall transmit power	Transmit power needed to cover full range	Transmit power needed to cover full range	Overall transmit power lower than in other solutions for same range and therefore more sustainable
Beamforming	Plane, concave or convex reflector, some fixed static beamforming feasible	Beamforming by settings phase shifters of RIS elements; RIS gain (similar to antenna array gain)	Directive and/or MIMO antennas
Update vs. receiver location	Not needed, because reflector is covering an area around the receiver	For a specific user location, the radio channel can be influenced or adjusted by RIS settings. However, RIS settings are sensitively linked to the receiver location, depending on beamwidth, and require fast updates due to receiver movements	Needed, because beamforming or MIMO antenna concepts applicable
Amplification	No amplification		Amplification
Signal regeneration	No		RF repeater only: no, for analogue repeater (amplifier); Repeater including signal processing: yes, for repeater including signal processing by means of available information from radio interface
Signal processing	No additional signal processing	Additional centralized signal processing between user equipment, base station and RIS	No additional signal processing in addition to the needs of the radio interface
Latency	No latency added to signal propagation		RF repeater only: nearly no latency added; Repeater including signal regeneration: signal processing for signal regeneration and antenna algorithms adds latency to signal propagation
Shadowing mitigation	Yes, allows to look around corners		
Coverage improvement	Yes, shaded areas can be covered even in millimeter wave and sub-THz domains		
LOS like propagation conditions	Propagation corresponds to LOS propagation between base station and reflector and a second LOS propagation between reflector and user equipment	Propagation corresponds to LOS propagation between base station and RIS array and a second LOS propagation between RIS array and user equipment	Propagation corresponds to LOS propagation between base station and repeater and a second LOS propagation between repeater and user equipment

Tab. 2. Comparison of features – continued

Topic	Reflector	RIS array	Repeater
Range extension compared to propagation under shadowing conditions	Yes, based on quasi-LOS propagation around corners	Yes, based on quasi-LOS propagation around corners, only for huge number of RIS elements better than reflector	Yes, based on amplification and directive antennas in repeater and quasi-LOS propagation around corners provides in maximum twice the range as for repeater
Strong component in channel impulse response	Yes, reflector, RIS array or repeater ensures a strong in channel impulse response compared to other secondary reflections around user equipment		
Type of radio channel at receiver	**Rice channel** due to strong component Without reflector, RIS array or repeater received power much lower and radio channel corresponds to Rayleigh channel		
Delay spread at receiver	Delay spread generated via secondary reflections around the receiver or other reflecting obstacles with smaller reflection coefficients in the deployment area. **Rice channel** – low delay spread, **Rayleigh channel** – higher delay spread		
Coherence bandwidth	Enforced Rice channel at receiver provides very high coherence bandwidth		
Intersymbol interference	With high coherence bandwidth little effective intersymbol interference, which reduces equalization effort		
Coherence time	With increasing Rice factor of radio channel, coherence time is increasing. Reduced requirements on equalizer update rate of channel estimation		
Flexibility	Static, no flexibility	Some beamforming due to RIS settings	Very flexible by MIMO and signal regeneration
Overall complexity	Low	High	Medium

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