Reconfigurable Intelligent Surface Optimal Placement in Millimeter-Wave Networks

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This work discusses the optimal placement of a reconfigurable intelligent surface (RIS) in a millimeter wave (mmWave) point-to-point link. In particular, we present a novel system model that takes into account the relationship between the transmission beam footprint size at the RIS plane and the RIS size. Moreover, we present the theoretical framework that quantifies the RIS gain loss in the case that the transmission beam footprint in the RIS plane is smaller than the RIS size, and the beam waste for the case in which the transmission beam footprint in the RIS plane is larger than the RIS size, by extracting closed-form expressions for the received power and the end-to-end signal-to-noise-ratio (SNR) for both cases. Subsequently, building upon the expressions, we provide SNR maximization policies for both cases. Finally, we perform a comparison, in terms end-to-end SNR, between the RIS-aided link and its relay-aided counterpart.

Index Terms—Optimal placement, Reconfigurable intelligent surfaces, Relays, Signal-to-noise ratio analysis.

NOMENCLATURE

3D Three-Dimensional
AF Amplify-and-Forward
B5G Beyond the Fifth Generation
DF Decode-and-Forward
FD Full-Duplex
HD Half-Duplex
HPBW Half-Power Beamwidth
LoS Line-of-Sight
mmWave Millimeter Wave
NLoS Non-Line-of-Sight
PIN Positive-Intrinsic-Negative
RF Radio-Frequency
RIS Reconfigurable Intelligent Surface
RU Reflection Unit
RX Receiver
SNR Signal-to-Noise-Ratio
TX Transmitter

I. INTRODUCTION

Increasing data-rate demands have led current mobile-access networks relying on sub-6 GHz bands reach their limits in terms of available bandwidth. This bottleneck created the need to consider above-6 GHz bands for mobile-access networks. Currently, bands in the lower-end millimeter wave (mmWave) spectrum are used for point-to-point and point-to-multipoint line-of-sight (LoS) wireless backhaul/fronthaul and fixed-wireless access networks [1]. Such deployments span the 30-100 GHz operational frequency range. However, the expected migration of future mobile-access networks to the 30 – 100 GHz band pushes the corresponding wireless backhaul/fronthaul links towards the beyond-100 GHz bands. Due to this, backhauling/fronthauling transceiver equipment vendors have performed LoS trials in the D-band (130 – 174.8 GHz), which showcase the potential of using it in such deployments [2]. Apart from LoS, street-level deployments in dense urban scenarios necessitate devising non-LoS (NLoS) solutions since LoS links may not always be available. However, despite the fact that according to measurements [3], [4] NLoS communication through scattering and reflection from objects in the radio path is feasible in the 30 – 100 GHz range, the higher propagation loss of beyond-100 GHz bands is likely to challenge this assumption.

The conventional approach of counteracting NLoS links is by providing alternative LoS routes through relay nodes [5]. Although this is a well-established method to increase the coverage when the signal quality of the direct links is low, it is argued that it cannot constitute a viable approach for massive deployment, especially for mmWave networks. This is due to the increased power consumption of the active radio-frequency (RF) components in high frequencies that relays need to be equipped with [6]. Apart from relaying, communication through passive non-reconfigurable specular reflectors, such as dielectric mirrors, has been proposed as another alternative. Such a method for coverage enhancement has the potential to be notably more cost efficient compared with relying and has been documented at both mmWave and beyond-100 GHz bands [6], [7]. Due to the highly dynamic nature of blockage at high frequencies together with the traffic conditions which may necessitate fast rerouting of information
within a network, it would be desirable that such reflectors can change the angle of departure of the waves so that they direct the beams towards different routes. However, passive reflectors are incapable of supporting the aforementioned functionality since the conventional Snell’s law applies. Furthermore, even by enabling this functionality by means of mechanical steering of the passive reflectors, the resulting latency would substantially compromise the desired reliability. Based on the above, an intriguing question that arises is the following: Would it be possible to deploy reconfigurable reflectors that can arbitrarily steer the impinging beam based on dynamic blockage and traffic conditions and without compromising the desired latency? The answer is affirmative by considering the reconfigurable intelligent surface (RIS) paradigm.

RISs are two-dimensional structures of dielectric material, which embed tunable reflection units (RUs) [8]–[12]. They constitute a substantially different technology than relaying, due to the absence of bulky and power-hungry analog electronic components, such as power amplifiers and low-noise amplifiers. Additionally, their operation, in contrast with relaying, does not require dividers and combiners, which can incur high insertion losses. By individually tuning the phase response of each individual RIS element, the reflected signals can constructively aggregate at a particular focal point, such as the receiver. Such a tuning can be enabled by electronic phase-switching components, such as positive-intrinsic-negative (PIN) diodes, RF-microelectromechanical systems, and varactor diodes, that are introduced between adjacent elements [13]. Hence, RISs offer an alternative-to-relaying method for large-scale beamforming without the incorporation of high power consuming electronics and insertion losses involved by the additional circuitry. In practice, the RIS element phase shift can be controlled by a central controller through programmable software [13].

Recognizing the unprecedented features that RISs can bring to beyond the fifth generation (B5G) wireless systems, a great amount of research effort has been put on analyzing, designing and optimizing RIS-aided wireless systems [13]–[17], as well as comparing them with their predecessors, i.e., relaying-aided ones [11], [13]–[20]. In more detail, in [13] and [14], the authors introduced the idea of employing RIS in order to mitigate the impact of blockage and steer the transmission beam towards the desired direction. Likewise, in [15], the authors presented the optimization framework for the maximization of the reception power in a RIS-aided system, assuming that all the RIS area can be used to reflect the induced electromagnetic wave. Moreover, in [16], the authors studied the asymptotic uplink ergodic capacity performance of a RIS-aided wireless system, while, in [17], the coverage of a downlink RIS-assisted network was studied, assuming that the entire RIS area can be used, and a strategy for maximizing the cell coverage by optimizing the RIS orientation and horizontal distance was proposed.

As far as the comparison with relays is concerned, in [11] the authors derive closed-form average signal-to-noise-ratio (SNR), outage probability, and symbol-error-rate expressions for a sub-6 GHz point-to-point RIS-aided link between a single-antenna source and and destination and compare the achievable performance with an amplify-and-forward (AF) relay. According to their findings, the RIS-based link outperforms its relay-based counterpart. In [18], the authors compare the achievable rate of RIS-aided systems with both the half-duplex (HD) and full-duplex (FD) relaying ones and show that a sufficiently large RIS is needed to guarantee the same rate performance with a single-antenna relay when the former is used for anomalous reflection instead of beamforming/focusing. [19] reveals that RIS-aided systems outperform the corresponding amplify-and-forward relaying ones for about 300% in terms of energy efficiency. Finally, contrary to the findings of [19], in [20] the authors show that an RIS needs hundred of radiating elements to outperform, in terms of rate, a single-antenna HD and decode-and-forward (DF) relay. In addition, they show that the RIS can be more energy efficient than the relay only for very high data rates.

**Motivation, novelty, and contribution:** All the presented RIS-related works consider the case of the entire RIS area being illuminated by the transmitted beam. However, due to the highly directional transmissions in mmWave networks and the low manufacturing cost of RISs, which make them suitable, as it is envisioned, to cover a big portion of the facades of large structures, such as buildings, it is expected that in many cases only a part of the total RIS area is going to be illuminated. Based on this, our work is motivated by the need to answer the question of what the optimal RIS placement policy is, which can be seen as a network planning question, in the two cases of the RIS area being smaller and larger than the transmitted beam footprint. Summarizing, the technical contribution of the paper is as follows:

- We present a system model for an RIS-aided mmWave links of fixed topology, such as wireless backhaul/fronthaul links, and use electromagnetic theory to evaluate the received power for the cases in which the transmission footprint at the RIS plane is smaller and larger than the RIS. The derived expressions, apart from the TX, RIS and RX positions, take into account the antenna characteristics as well as the RIS particularities, and quantify the performance degradation due to RIS underexploitation or beam waste. Of note, the presented analysis can also find application in mobile mmWave networks.
- Based on these expressions, we evaluate the end-to-end SNR for both cases in which the transmission beam footprint at the RIS plane is smaller and larger than the RIS. The derived expressions, apart from the TX, RIS and RX positions, take into account the antenna characteristics as well as the RIS particularities, and quantify the performance degradation due to RIS underexploitation or beam waste. Of note, the presented analysis can also find application in mobile mmWave networks.
- Subsequently, we analytically extract two policies for the optimum RIS placement for both cases under investigation.
- Finally, we compute the end-to-end SNR for a HD and DF relay-aided link and compare it with the corresponding RIS-aided one. The analytical results, which are verified by means of simulations, reveal a different monotonic behavior of the SNR ratio with respect to
the transmitter (TX)-receiver (RX) distance for small and large RIS structures.

Organization: The rest of this contribution is structured as follows: In Section [II] the system model is presented, while, in Section [III] we report the end-to-end SNR. Moreover, the optimal RIS placement problem is discussed and solved in Section [IV] whereas, an insightful comparison against the corresponding relaying system is provided in Section [V]. Numerical results and discussions are given in Section [VI]. Finally, Section [VII] concludes the paper by highlighting the most important findings and remarks.

Notation: For the convenience of the readers, recurrent parameters and symbols with their meaning are presented in Table [I].

### Table I
| Parameter/Symbol | Meaning |
|------------------|---------|
| \( r_1 \) | Carrier frequency |
| \( f_t \) | Wavelength |
| \( W \) | Transmit power |
| \( \gamma_{\text{div}} \) | Signal bandwidth |
| \( N_0 \) | Noise figure |
| \( h_{\text{RIS/RX,rel}} \) | RIS/relay height with respect to the ground |
| \( h_{\text{t/rx}} \) | TX/RX height with respect to the ground |
| \( D_{\text{L},D_{\text{L},D_{\text{R},rel}}} \) | TX/RX/relay antenna diameter |
| \( \theta_{\text{s},\text{rel}} \) | TX/RX/relay antenna HPBW |
| \( \alpha, \beta \) | TX/RX/relay antenna aperture efficiency |
| \( g_i, g_{\text{rel}} \) | TX/RX/relay antenna gain |
| \( \epsilon \) | Parameter defining the steepness of the RU radiation pattern |
| \( \rho \) | Parameter defining the steepness of the RU radiation pattern |
| \( d_{x}, d_{y} \) | X-axis and y-axis length, respectively, of the RUs |
| \( \phi_i, \theta_i \) | Radii of the TX beam elliptic footprint |
| \( \beta_r \) | Eccentricity of TX beam elliptic footprint |
| \( r_1 \) | Distance between the center of the TX antenna and the center of the RX antenna |
| \( r_2 \) | Distance between the center of the TX antenna and the center of the RX antenna |
| \( r_{i,\text{rel}}, r_{j,\text{rel}} \) | Optimal RIS and relay, respectively, horizontal distance |
| \( \rho_{\text{RIS}} \) | RIS area |
| \( \gamma_{\theta} \) | Area of the TX beam elliptic footprint |
| \( \gamma_{\phi} \) | Beam width |
| \( \rho_{\text{RIS}} \) | Power received by the relay at the RIS |
| \( \rho_{\text{RIS}} \) | Power received by the relay at the RIS |
| \( \rho_{\text{rel}} \) | SNR of the relay-aided link in the 1st hop |
| \( \rho_{\text{rel}} \) | SNR of the relay-aided link in the 2nd hop |
| \( \rho_{\text{rel}} \) | SNR of the relay-aided link in the 3rd hop |
| \( \rho_{\text{rel}} \) | SNR of the relay-aided link in the 4th hop |

II. SYSTEM MODEL

As illustrated in Fig. [1] we consider a fixed-topology scenario, in which a TX communicates with a RX through an RIS. \( \theta_i \) and \( \theta_r \) are the incidence and departure angles, respectively, of the electromagnetic wave. The TX-RIS and RIS-RX LoS links are established in a mmWave band and constitute an alternative path to the direct TX-RX link that is assumed to be blocked. To countermeasure the high pathloss in this band, both the TX and RX are equipped with highly directional parabolic antennas with diameters \( D_t \) and \( D_r \), respectively. As a result, their gains can be obtained as [21]

\[
G_m = e_m \left( \frac{\pi D_m}{\lambda} \right)^2, \quad \text{with } m \in \{t, r\},
\]

where \( e_t \) and \( e_r \) respectively denote the aperture efficiencies of the TX and RX parabolic reflectors, while \( \lambda \) represents the transmission wavelength. Note that this type of antennas has been extensively used for wireless backhauls/fronthaul scenarios (see e.g., [22] and reference therein), due to their capability to support pencil-beamforming transmissions. Under such directional transmissions, almost the entire transmit energy is located within the half-power beamwidth (HPBW) \( \theta_{\text{HPBW}} \) and the three-dimensional (3D) antenna pattern can be modeled as a cone for HPBWs smaller than approximately 15° [23] Ch. 12]. The HPBW of the TX and RX antennas can be tightly approximated as in [24]

\[
\phi_m \approx \frac{1.22 \lambda}{D_m}, \quad \text{with } m \in \{t, r\}. \tag{2}
\]

By substituting (2) into (1), we can rewrite the TX and RX antenna gains as

\[
G_m \approx \left( \frac{1.22 \pi}{\phi_m} \right)^2 e_m, \quad \text{with } m \in \{t, r\}. \tag{3}
\]

Likewise, we assume that the TX and RX antennas can be mechanically steered, both in azimuth and elevation, towards the desired angle of transmission and reception, respectively, and they are pointing towards the center of illuminated RIS region.

The RIS acts as a beamformer, which by adjusting the phase response of the RUs is capable of steering the beam at \( \theta_i \) which is the RX direction. It consists of \( N_x \times N_y \) RUs of size \( d_x \times d_y \) and a controller that has perfect knowledge of the TX and RX positions. Each RU is an electrically-small low-gain element embedded on a substrate, with radiation pattern that can be expressed as in [25]

\[
G_{\text{RIS}}(\theta) = \gamma \cos^p(\theta), \quad 0 \leq \theta < \pi/2, \tag{4}
\]
where γ stands for the RU gain and p is a constant that determines the steepness of the radiation pattern and depends on the RU type. Based on [25], γ can be evaluated as

\[ \gamma = 2(2p + 1). \]  

(5)

Due to the fact that the TX-RIS and RIS-RX links are directional LoS links, they are deterministic. Moreover, it is assumed that the transmission power is \( P_t \) and that the received signal is subject to additive white Gaussian noise with power

\[ N_0 = -174 + 10 \log_{10} W + \mathcal{F}_{\text{dB}}, \]  

(6)

where \( \mathcal{F}_{\text{dB}} \) is the noise figure in dB and \( W \) is the transmission bandwidth [26].

III. SNR

In this section, we firstly derive the illuminated RIS area and, subsequently, compute the resulting end-to-end SNR in the two cases of the RIS area being smaller and larger than the transmitted beam footprint.

A. RIS’s Illuminated Area

Since the main lobe of the transmission antenna has a conical shape, its footprint in the RIS plane is an ellipse, according to the conic-section theory [27]. The following Lemma returns the two radii of the illuminated elliptic area at the RIS.

**Lemma 1:** The two radii of the illuminated elliptic area at the RIS plane can be obtained as

\[ \alpha = \frac{\sin \left( \frac{\theta_i}{2} \right)}{\cos \left( \theta_i + \frac{\theta_t}{2} \right)} r_1 \]  

(7)

and

\[ \beta = \alpha \sqrt{1 - \epsilon^2}, \]  

(8)

where

\[ \epsilon = \frac{\sin \left( \theta_i \right)}{\cos \left( \frac{\theta_t}{2} \right)}. \]  

(9)

Moreover, \( r_1 \) denotes the distance between the center of the TX and the center of the TX footprint at the RIS plane, while \( \theta_i \) is the incident angle at the RIS with respect to its broadside direction.

**Proof:** For brevity, the proof is provided in Appendix A.

(10)

Based on (7) and (8), the TX beam footprint at the RIS plane can be evaluated as

\[ S_i = \pi \alpha \beta. \]  

(10)

The power that is reflected by the RIS is the one that falls within

\[ S = \min \left( S_i, S_{\text{RIS}} \right), \]  

(11)

where \( S_{\text{RIS}} \) denotes the RIS area. If \( S_{\text{RIS}} \leq S_i \), only part of the power that falls within \( S_{\text{RIS}} \) can be reflected towards the RX; thus, beam waste occurs. On the other hand, if \( S_{\text{RIS}} > S_i \), only part of the RIS is used to reflect the incident electromagnetic wave; thus, the RIS gain is decreased.

B. End-to-end SNR

The following proposition returns a tight approximation for the received power.

**Proposition 1:** The received power can be evaluated as

\[ P_r \approx \frac{2.22 \Gamma^2 S_i^2 e \epsilon \gamma^2 \lambda^4}{256 d_2^2 d_3^2 \phi_i^2 \phi_t^2 r_1^2 r_2^2} \cos^p (\theta_i) \cos^p (\theta_r) P_t, \]  

(12)

where \( r_2 \) denotes the distance between the center of the TX footprint at the RIS plane and the center of the RX and \( \Gamma \) stands for the amplitude reflection coefficient of each RU for which it holds

\[ \Gamma_n = \Gamma e^{j \theta_n}, \]  

(13)

where \( \Gamma_n \) stands for the complex reflection coefficient of the \( n \)-th illuminated RU and \( \theta_n \) is the phase it induces.

**Proof:** For brevity, the proof is provided in Appendix B.

Based on (11) and (12), for the case in which \( S_{\text{RIS}} < S_i \), (12) can be simplified as

\[ \epsilon_w = 1 - \frac{S_{\text{RIS}}}{S_i}, \]  

(15)

or, based on (7)-(10),

\[ \epsilon_w = 1 - \frac{S_{\text{RIS}}}{\pi \sin^2 \left( \frac{\phi_t}{2} \right) r_1^2 \sqrt{1 - \frac{\sin^2 (\theta_i)}{\cos^2 \left( \frac{\theta_t}{2} \right)}}}. \]  

(16)

Finally, the end-to-end SNR can be evaluated as

\[ \rho_{\text{RIS}} = \frac{P_r}{N_0}, \]  

(17)

or, equivalently,

\[ \rho_{\text{RIS}} \approx \left( \frac{\lambda^4}{256} \right) \frac{2.22 \Gamma^2 S_i^2 e \epsilon \gamma^2}{d_2^2 d_3^2 \phi_i^2 \phi_t^2 r_1^2 r_2^2 N_0} \cos^p (\theta_i) \cos^p (\theta_r). \]  

(18)

On the other hand, for the case in which \( S_{\text{RIS}} > S_i \), from (12) and (11), we can evaluate the received power as in (19), given at the top of the following page. Consequently, the equivalent end-to-end SNR at the RX can be approximated as in (20), given at the top of the following page.

IV. OPTIMAL RIS PLACEMENT

This section is focused on identifying the optimal RIS placement in respect to the TX and RX position that maximizes the end-to-end SNR. Let us denote by \( r_{1,h}, r_{2,h}, \) and \( r_h \) the horizontal TX-RIS, RIS-RX, and TX-RX distances, respectively, and by \( h_t, h_{\text{RIS}} \) and \( h_r \) the TX, RIS, and RX
The mean value of \( h \) denoted by \( h_{\text{RIS}} \) is the optimal TX-RIS horizontal distance, and \( h_{\text{RIS}} > S_i \) is the optimal RIS placement that maximizes the end-to-end SNR.

**Proposition 2:** The optimal TX-RIS horizontal distance, denoted by \( r_{1,h}^* \), that jointly maximizes the end-to-end SNR and minimizes the beam waste can be obtained as

\[
\begin{align*}
  r_{1,h}^* &= \sqrt{r_{1,h}^2 + (h_{\text{RIS}} - h_A)^2}, \\
  \theta_t &= \tan^{-1}\left(\frac{r_{1,h}^*}{h_{\text{RIS}} - h_t}\right), \\
  \theta_r &= \tan^{-1}\left(\frac{r_{1,h} - h_{\text{RIS}} - h_A}{h_{\text{RIS}} - h_r}\right).
\end{align*}
\]

Next, for the \( S_{\text{RIS}} \leq S_i \) and \( S_{\text{RIS}} > S_i \) cases, we determine the \( r_{1,h}^* \) that maximizes the end-to-end SNR.

**S_{\text{RIS}} \leq S_i:** In this case, apart from the end-to-end SNR maximization, we additionally aim at minimizing the beam waste. In this direction, the following proposition returns the optimum TX-RIS horizontal distance that jointly maximizes the end-to-end SNR and minimizes the beam waste.

**Proposition 3:** The optimal TX-RIS placement can be obtained as the solution of

\[
\begin{align*}
  ar_{1,h}^3 + br_{1,h}^2 + cr_{1,h} + d &= 0,
\end{align*}
\]

where again we consider \( h_t = h_r = h_A \). Of note, in (26),

\[
\begin{align*}
  a &= 2 - 2p, \\
  b &= (3p - 6) r_{1,h}, \\
  c &= (2 - 2p) (h_{\text{RIS}} - h_A)^2 + (4 - p) r_{1,h}^2
\end{align*}
\]

1In practical scenarios, it is expected that \( h_t = h_r \). However, even if \( h_t \neq h_r \), holds \( r_{1,h}^* \) is approximately given by (25) by setting \( h_A \) as the mean value of \( h_t \) and \( h_r \).
In the first time slot, only the communication between the transmit and relay receive antennas is performed. The end-to-end SNR at the relay RX, which we denote by $\rho_{rel}^{(1)}$, is given by

$$\rho_{rel}^{(1)} = \left( \frac{\lambda}{4\pi} \right)^2 \frac{P_{tx} G_t G_{rel}}{r_{1,rel}^2 N_0}, \quad (32)$$

where

$$G_{rel} \approx \left( \frac{1.22 \pi}{\phi_{rel}} \right)^2 e_{rel} \quad (33)$$
is the relay antenna gain, $\phi_{rel}$ is its HPBW, $e_{rel}$ is its aperture efficiency, and $r_{1,rel}$ is the distance between the centers of the transmit and relay receive antennas. It holds that

$$r_{1,rel} = \sqrt{r_2^2 + (h_{rel} - h_t)^2}, \quad (34)$$

where $r_{1,rel}$ is the horizontal transmitter-relay distance and $h_{rel}$ is the height of the relay with respect to the ground.

In the second time slot, only the communication between the relay transmit and receive antennas is performed. The SNR at the relay RX, which we denote by $\rho_{rel}^{(2)}$, is given by

$$\rho_{rel}^{(2)} = \left( \frac{\lambda}{4\pi} \right)^2 \frac{P_{tx} G_t G_{rel}}{r_{2,rel}^4 N_0}, \quad (35)$$

where $r_{2,rel}$ is the distance between the centers of the relay transmit and receive antennas. It holds that

$$r_{2,rel} = \sqrt{(r_h - r_{1,rel})^2 + (h_{rel} - h_r)^2}. \quad (36)$$

Based on the DF protocol, the end-to-end SNR, which we denote by $\rho_{rel}^{(DF)}$, is given by

$$\rho_{rel}^{(DF)} = \min \left\{ \rho_{rel}^{(1)}, \rho_{rel}^{(2)} \right\}. \quad (37)$$

It is trivial to prove that the maximum value of $\rho_{rel}^{(DF)}$ at the point $r_{1,rel}^{*}$, which we denote by $\rho_{rel}^{(DF)}(r_{1,rel}^{*})$, is achieved when the relay is placed in a position for which $\rho_{rel}^{(1)} = \rho_{rel}^{(2)}$ holds. Hence,

$$\rho_{rel}^{(DF)}(r_{1,rel}^{*}) = \rho_{rel}^{(1)} = \rho_{rel}^{(2)}, \quad (38)$$
or equivalently

$$\frac{G_t}{r_{1,rel}^2} = \frac{G_r}{r_{2,rel}^2}. \quad (39)$$

Remark 2: By assuming that $G_t = G_r$, which is considered a realistic assumption, since the relay usually employs the same antenna for reception and transmission, from (39), we observe that the relay position that maximizes $\rho_{rel}^{(DF)}(r_{1,rel}^{*})$ is the one that satisfies the condition

$$r_{1,rel}^{*} = r_{2,rel}. \quad (40)$$

Moreover, if $h_t = h_r$, from (34) and (36), it becomes apparent that (40) can be simplified as

$$r_{1,rel}^{*} = \frac{r_h}{2}. \quad (41)$$

This indicates that the relay should be placed in the middle of the TX-RX horizontal distance.

### B. SNR comparison

For obtaining expressions that allow us to acquire important insights and without loss of generality, for the comparison of the RIS and relay-aided links we assume that $\phi_t = \phi_r = \phi_A$, $h_t = h_r = h_A$, and $h_{RIS} = h_{rel} = h_R$. As far as the comparison of these two link types is concerned, for the RIS case we consider the two examined cases of $S_{RIS} \leq S_i$ and $S_{RIS} > S_i$. In addition, we assume that $r_h^2 >> 4(h_R - h_A)^2$ and $d_{x} = d_{y} = \frac{1}{2}$ since this commonly considered adjacent-element distance value substantially reduces mutual coupling among adjacent elements. Hence, it enables an almost independent impedance adjustment of each RU from the corresponding one of its adjacent ones.

**Remark 2:** Due to the fact that according to Proposition 3 the optimal RIS placement in the $S_{RIS} > S_i$ case is relatively close to the RX and the fact that $r_h^2 >> 4(h_R - h_A)^2$ according to our assumptions, for the optimal value $r_{1,rel}^*$ that maximizes $\rho_i(r_{1,rel})$, it holds that $r_{1,rel}^* \approx r_h$. Based on this and the fact that $\phi_t << 1$, $\rho_i(r_{1,rel})$ in (80) can be further approximated as

$$\rho_i(r_{1,rel}^*) \approx \frac{2.22 P_{tx} \Gamma_{2,\gamma}^2 e_{A}^2 r_h^2}{544 \pi^2 (h_R - h_A)^4 N_0} \times \cos^2 \left( \tan^{-1} \left( \frac{r_h}{h_R - h_A} \right) \right). \quad (45)$$

According to (42) and (43), for the ratio of $\rho_{RIS}(r_{1,rel})$ over $\rho_{rel}(r_{1,rel})$, which we denote by $\zeta_{RIS,rel}$, it holds that

$$\zeta_{RIS,rel} = \frac{\rho_{RIS}(r_{1,rel}^*)}{\rho_{rel}(r_{1,rel}^*)} \approx \frac{\Gamma_{2,\gamma}^2 S_{RIS}^2 \cos^2 \left( \tan^{-1} \left( \frac{r_h}{h_R - h_A} \right) \right)}{16 \lambda^2 \pi^2 (h_R - h_A)^2}. \quad (44)$$

**Remark 3:** As we observe from (41), $\zeta_{RIS,rel}$ is a monotonically decreasing function of $r_h$. Hence, even if for some $r_h$ it holds that $\zeta_{RIS,rel} > 1$ as $r_h$ increases there is a threshold value of $r_h$ after which $\zeta_{RIS,rel} < 1$ would hold.

**Remark 3:** Due to the fact that according to Proposition 3 the optimal RIS placement in the $S_{RIS} > S_i$ case is relatively close to the RX and the fact that $r_h^2 >> 4(h_R - h_A)^2$ according to our assumptions, for the optimal value $r_{1,rel}^*$ that maximizes $\rho_{rel}(r_{1,rel})$, it holds that $r_{1,rel}^* \approx r_h$. Based on this and the fact that $\phi_t << 1$, $\rho_{rel}(r_{1,rel})$ in (80) can be further approximated as

$$\rho_{rel}(r_{1,rel}^*) \approx \frac{2.22 P_{tx} \Gamma_{2,\gamma}^2 e_{A}^2 r_h^2}{544 \pi^2 (h_R - h_A)^4 N_0} \times \cos^2 \left( \tan^{-1} \left( \frac{r_h}{h_R - h_A} \right) \right). \quad (45)$$

According to (42) and (43), for the ratio of $\rho_i(r_{1,rel})$ over $\rho_{rel}(r_{1,rel})$, which we denote by $\zeta_{i,rel}$, it holds that
\[ \zeta_{i,rel} = \frac{\rho_i (r_{1,h}^*)}{\rho_{rel}(r_{1,rel,h})} \]
\[ \approx \frac{1}{64(h_r - h_A)^2 \lambda^2} \left( \tan^{-1}\left( \frac{r_h}{h_r - h_A} \right) \right)^2. \]  

(46)

Remark 4: Regarding the 1st derivative of \( \zeta_{i,rel} \) with respect to \( r_h \), which we denote by \( \frac{d\zeta_{i,rel}}{dr_h} \), it holds that
\[ \frac{d\zeta_{i,rel}}{dr_h} = \frac{4(h_r - h_A)^2 r_h^2 - (p - 6) r_h^2}{64(h_r - h_A)^2 \lambda^2} \]
\[ \times \Gamma \left( \frac{2}{(h_r - h_A)^2 + 1} \right)^{-\frac{5}{2}}. \]  

(47)

From (47) we observe that \( \zeta_{i,rel} \) is a monotonically increasing function of \( r_h \) for \( p < 6 \), which means that under practical implementations \( \zeta_{i,rel} \) increases as \( r_h \) increases. Furthermore, by direct inspection of (46) we expect that \( \zeta_{i,rel} \gg 1 \) even for small values of \( r_h \), whereas the numerator scales with \( r_h^2 \), whereas the denominator with \( \lambda^2 \). Such a claim is validated in Section VII.

VI. NUMERICAL RESULTS & DISCUSSION

The aim of this section is threefold: i) To illustrate how the beam waste is affected by the positioning and the size of the RIS; ii) To validate Propositions 2 and 3 regarding the optimal placement of an RIS in the \( S_{RIS} \leq S_i \) and \( S_{RIS} > S_i \) cases, respectively; iii) To validate Remarks 3 and 4 regarding end-to-end SNR ratio of the RIS- and relay-aided links. In this way, for the involved parameters we consider the values of Table II. For the considered parabolic reflector dimension, the HPBW is approximately equal to 0.98°, which complies with the analysis that is based on the pencil-beam assumption.

| Parameter | Value |
|-----------|-------|
| \( \theta \) | 10° GHz |
| \( \rho_i \) | 1 W |
| \( \lambda \) | 2 GHz |
| \( \rho_{rel} \) | 10 dB |
| \( h_r, h_A \) | 15 m |
| \( D_i, D_{rel}, D_{rel,h} \) | 15 cm |
| \( \epsilon_{Rx}, \epsilon_{Rel}, \epsilon_{rel} \) | 0.7 |
| \( \chi \) | 0.9 |
| \( p \) | 0.5 |

A. Impact of \( r_{1,h} \) on \( \epsilon_w \)

Regarding the behavior of \( \epsilon_w \) with respect to \( r_{1,h} \), in Fig. 2 we illustrate \( \epsilon_w \) vs. \( r_{1,h} \) for \( r_h = 40 \text{ m} \) and different \( S_{RIS} \). As we observe from Fig. 2, \( \epsilon_w \), as expected, increases as the RIS is moving towards the RX, since the impinging beam footprint increases. However, the larger \( S_{RIS} \) is the larger the range of \( r_{1,h} \) is for which \( \epsilon_w = 0 \) holds. In fact, we observe that for the \( S_{RIS} = 1.5 \text{ m}^2 \) case it holds that \( r_{1,h} = 0 \) for \( r_{1,h} \leq r_h \).

\footnote{Such short \( r_h \) distances are foreseen for the mmWave backhaul networks of next-generation ultra-dense small-cell networks [28].}

\( B. \) Validation of Propositions 2 and 3

As far as the validation of Proposition 2 is concerned, in Fig. 3(a) and Fig. 3(b) we depict the \( \rho_{RIS} \) vs. \( r_{1,h} \) and \( \epsilon_w \) vs. \( r_{1,h} \) curves for \( r_h = 40 \text{ m} \) and \( S_{RIS} = 0.02 \text{ m}^2 \). Such an \( S_{RIS} \) value is small enough so that there is a beam waste throughout the examined \( r_{1,h} \) range. As we observe from Fig. 3(a), there are two symmetrical points for which \( \rho_{RIS} \) obtains the same maximum value. The 1st point, which is closer to the TX, is at \( r_{1,h} = 4 \text{ m} \) and the second point, which is closer to the RX, at \( r_{1,h} = 36 \text{ m} \). However, since \( \epsilon_w \) is a monotonically increasing function of \( r_{1,h} \) for \( |r_{1,h}| > 0 \), based on (16), and as we observe from Fig. 3(b), the selected value of \( r_{1,h} \) between these two values is \( r_{1,h} = 4 \text{ m} \). Consequently, the RIS should be deployed closer to the TX.

Now, regarding the validation of Proposition 3 in Fig. 4(a) and Fig. 4(b) we illustrate the \( \rho_i \) vs. \( r_{1,h} \) and \( \epsilon_w \) vs. \( r_{1,h} \) curves for \( r_h = 40 \text{ m} \) by considering that \( S_{RIS} \) is sufficiently large so that \( S_{RIS} > S_i \). As we observe from Fig. 4(a), there is only one value of \( r_{1,h} \) that maximizes \( \rho_i \), which corresponds to a placement closer to the RX than the TX, as stated in Proposition 3. For the particular point it holds that \( r_{1,h} \approx 45 \text{ m} \) and \( S_i \approx 2.15 \text{ m}^2 \). Hence, to obtain such a behavior of the end-to-end SNR an RIS with an area at least equal to approximately 2.15 \text{ m}^2 is required. This validates the practicality of such a deployment since such an area can be mounted on the facades of large structures, such as buildings.

In addition, in Fig. 5(a) we illustrate the \( r_{1,h} \) vs. \( S_{RIS} \) curve for \( r_h = 40 \text{ m} \) and a scanning range \( r_{1,h} \in [-20 \text{ m}, 60 \text{ m}] \). From Fig. 5(a), we observe the two regions for which \( S_{RIS} < S_i \left( r_{1,h}^* \right) \) and \( S_{RIS} > S_i \left( r_{1,h}^* \right) \) hold throughout the scanning range of \( r_{1,h} \). In particular, we observe that the \( S_{RIS} < S_i \left( r_{1,h}^* \right) \) case holds for \( S_{RIS} < 0.04 \text{ m}^2 \), whereas the \( S_{RIS} > S_i \left( r_{1,h}^* \right) \) case holds for \( S_{RIS} > 2.15 \text{ m}^2 \). In addition, we observe that for the \( S_{RIS} < S_i \left( r_{1,h}^* \right) \) and \( S_{RIS} > S_i \left( r_{1,h}^* \right) \) regions the RIS should be placed closer to the TX and closer to the RX, respectively, which further validates Propositions 2 and 3 besides the results of Fig. 4. More specifically, for
Remark 3. Furthermore, in Fig. 6-(b) we depict the which validates the analysis based, according to Section V-B, increases attributed to the fact that any increase of $S$ increases $\rho$ in Fig. 5-(b), we illustrate the $\rho$ curve for the examined range of $r$. We note that for the particular value of $S_{RIS}$ it holds that $S_{RIS} < S_i \left(r_{1,h}^*\right)$ and $S_{RIS} > S_i \left(r_{1,h}^*\right)$ regions it holds that $r_{1,h}^* = 4 \text{ m}$ and $r_{1,h}^* \approx 44.9 \text{ m}$, respectively. Finally, in Fig. 5-(b), we illustrate the $\rho \left(r_{1,h}^*\right)$ vs. $S_{RIS}$ curve, where $\rho \left(r_{1,h}^*\right) \in \{\rho_{RIS} \left(r_{1,h}^*\right), \rho_i \left(r_{1,h}^*\right)\}$. As we observe, as $S_{RIS}$ increases $\rho \left(r_{1,h}^*\right)$ increases up to the value that corresponds to $S_{RIS} \approx 2.15 \text{ m}^2$ and it subsequently saturates. This is attributed to the fact that any increase of $S_{RIS}$ beyond the point for which $S_{RIS} = S_i \left(r_{1,h}^*\right) \approx 2.15 \text{ m}^2$ does not result in an increase in $\rho \left(r_{1,h}^*\right)$ since $r_{1,h}^*$ does not change.

C. Validation of Remarks 2 and 3

As far as the validation of Remarks 2 and 3 is concerned, in Fig. 6-(a) we depict the $\zeta_{RIS,rel}$ vs. $r_h$ curve for $S_{RIS} = 0.05 \text{ m}^2$. We note that for the particular value of $S_{RIS}$ it holds that $S_{RIS} < S_i \left(r_{1,h}^*\right)$ for the examined range of $r_h$. As we observe, $\zeta_{RIS,rel}$ monotonically decreases as $r_h$ increases, which validates the analysis based, according to Section V-B Remark 3. Furthermore, in Fig 6-(b) we depict the $\zeta_{i,rel}$ vs. $r_h$ curve for $S_{RIS} = 10 \text{ m}^2$. For the particular value of $S_{RIS}$ it holds that $S_{RIS} > S_i \left(r_{1,h}^*\right)$ for the examined range of $r_h$. As we observe, $\zeta_{i,rel}$ monotonically increases as $r_h$ increases, which validates the also analysis based, according to Section V-B Remark 4. Based on the results depicted in Fig. 6, we conclude that when $S_{RIS} < S_i \left(r_{1,h}^*\right)$ a substantial performance gain with respect to a relay is expected. According to the results, the larger the TX-RX distance is, the higher the performance gain is. With respect to this outcome, we again note that the facades of large structures, such as buildings, are the most suitable mounting structures for the required RIS sizes.

VII. CONCLUSIONS

This work has been motivated by the need to answer the question of where an RIS that aids a highly directional mmWave TX-RX link under blockage should be placed. Based on this, we have considered the two cases of the transmit beam
illuminating the entire RIS area in the case of an RIS smaller than the transmit beam footprint on the RIS plane, or only a portion of the RIS area in the case of large RIS structures with areas larger than the particular footprint. For these two cases, we have computed the end-to-end SNR, which enabled us to extract the optimal placement in each of them that maximizes it. In particular, in the case of an RIS smaller than the transmit beam footprint the RIS should be placed closer to the TX so as to minimize the beam waste. On the other hand, in the case of an RIS larger than the footprint a placement closer to the RX is the optimal solution. Equivalently, this network planning optimization can be viewed in a different perspective by considering a future urban scenario in which there is an abundance of buildings coated by RISs. In such a case, depending on the size of the RISs with respect to the transmit beam footprints, the transmitters will be focusing their beams to RISs closer to them or closer to the respective receivers so as to maximize the end-to-end SNR.

Furthermore, we have compared the resulting SNR expressions with the corresponding one of a HD/DF relay aiding the communication. In particular, for the $S_{\text{RIS}} < S_i$ case it was proved that the SNR ratio of the RIS- over the relay-aided link is a monotonically decreasing function of the TX-RX distance. This reveals that there is a threshold value of the particular distance above which the relay-aided link results in a better SNR performance. On the other hand, the corresponding SNR ratio is a monotonically increasing function of the TX-RX distance for the $S_{\text{RIS}} > S_i$ case. In addition, the results show that in the particular case even for a small TX-RX distance substantial SNR gains are expected by the RIS deployment compared to the relay-based one.

Finally, we note that although this work has considered links of fixed topology, the presented outcomes are also applicable to mobile networks. For instance, for a mobile user that should be served through an RIS the TX should focus its beam to an RIS mounted on a building close to it in the case of $S_{\text{RIS}} < S_i$. On the other hand, the TX should focus its beam to an RIS mounted on a building close to the mobile user in the $S_{\text{RIS}} > S_i$ case.
APPENDICES

APPENDIX A

PROOF OF LEMMA 1

By applying the law of sines in the \( \triangle ABC \), we obtain
\[
\sin \left( \frac{\phi_1}{2} \right) = \frac{\sin \left( \angle BCA \right)}{r_1}
\]
(48)

where \( \angle BCA \) can be calculated as
\[
\angle BCA = \frac{\pi}{2} - \frac{\phi_1}{2} - \phi_i.
\]
(49)

By substituting (49) into (48), we obtain (7). The eccentricity of the elliptical footprint can be evaluated as
\[
\epsilon = \frac{\sin (\phi_i)}{\sin \left( \frac{\pi}{2} - \frac{\phi_1}{2} \right)},
\]
(50)
or equivalently as in (9). Finally, \( \beta \) can be obtained as in (8). This concludes the proof.

APPENDIX B

PROOF OF PROPOSITION 1

The incident electric field on the \( n \)-th RU of the RIS illuminated area can be obtained as
\[
E_{i,n} = E_{i,n} e^{-j \frac{2 \pi}{\lambda} r_{t,n} \cdot n_o},
\]
(51)

where \( r_{t,n} \) is the distance between the TX and the \( n \)-th RU of the illuminated RIS area, while \( E_{i,n} \) is the amplitude of the incident wave, and \( n_o \) is a unitary vector that is perpendicular to the 2D plane that the electric field lies on [29] [Example 11-3]. Of note, since the RIS is placed in the far-field, \( E_{i,n} = E_i \) for each \( i = 1, \cdots, M \), where \( M \) represents the number of illuminated RUs in the RIS. It holds
\[
M = \int_{x} dx dy.
\]
(52)

Moreover, \( r_{t,n} \approx r_1 \). As a consequence, the power density at the \( n \)-th RU of the RIS can be expressed as
\[
P_i = \frac{E_i^2}{2 \eta}
\]
(53)
or, with the aid of (51), as
\[
P_i = \frac{P_i G_i}{4 \pi r_1^2},
\]
(54)

where \( \eta \) is the free-space impedance. Thus, the incident power at the \( n \)-th RU can be evaluated as
\[
P_{i,n} = P_i A,
\]
(55)

where \( A \) stands for the effective aperture of the \( n \)-th illuminated RU and can be obtained as
\[
A = \frac{\lambda^2}{4 \pi} G_{RIS}(\theta_i).
\]
(56)

By substituting (54) and (56) into (55), we obtain
\[
P_{i,n} = \left( \frac{\lambda}{4 \pi} \right)^2 P_i G_i G_{RIS}(\theta_i) \frac{1}{r_1^2}.
\]
(57)

As a result, and due to the energy conservation law, the reflected power density by the \( n \)-th RU, which is captured by the RX antenna, can be expressed as
\[
P_{r,n} = \frac{P_{i,n} |G_{n}|^2 G_{RIS}(\theta_r)}{4 \pi r_2^2}.
\]
(58)

By substituting (57) and (13) into (58), we obtain
\[
P_{r,n} = \frac{\lambda^2}{(4 \pi)^3} P_i G_i G_{RIS}(\theta_i) G_{RIS}(\theta_r) \frac{1}{r_1^2 r_2^2}.
\]
(59)

Moreover, the corresponding electric field can be evaluated as
\[
E_{r,n} = E_{r,n} e^{-j \left( \theta_n + \frac{2 \pi}{\lambda} (r_{t,n} + r_{n,r}) \right) a_o},
\]
(60)

where \( r_{n,r} \) is the distance between the \( n \)-th illuminated RU and the RX and \( a_o \) is a unitary vector perpendicular to the 2D plane that the reflected electric field lies on. Additionally, \( E_{r,n} \) can be computed as
\[
E_{r,n} = \sqrt{2 \eta P_{r,n}},
\]
(61)

which, by employing (59), can be rewritten as
\[
E_{r,n} = \sqrt{2 \eta \lambda^2 P_i G_i G_{RIS}(\theta_i) G_{RIS}(\theta_r) \frac{1}{r_1^2 r_2^2}}.
\]
(62)

From (60), the aggregated electric field at the RX can be written as
\[
E_r = \sum_{n=1}^{M} E_{r,n}.
\]
(63)

Consequently, by employing (60) and (62) it holds that
\[
E_r = \sqrt{2 \eta \lambda^2 P_i G_i G_{RIS}(\theta_i) G_{RIS}(\theta_r) \delta a_o},
\]
(64)

where
\[
\delta = \sum_{n=1}^{M} e^{-j \left( \theta_n + \frac{2 \pi}{\lambda} (r_{t,n} + r_{n,r}) \right)}.
\]
(65)

Hence, at the RX, the received power can be obtained as
\[
P_r = \left| E_r \right|^2 \frac{\lambda^2}{2 \eta} \gamma G_r,
\]
(66)

which, with the aid of (64), can be rewritten as
\[
P_r = \left( \frac{\lambda}{4 \pi} \right)^4 \Gamma^2 G_i G_r G_{RIS}(\theta_i) G_{RIS}(\theta_r) \frac{1}{r_1^2 r_2^2} |\delta|^2 P_i.
\]
(67)

By assuming that the optimal phase shift is induced by each RU so as to maximize \( P_r \), i.e.,
\[
\theta_n = - \frac{2 \pi (r_{t,n} + r_{n,r})}{\lambda},
\]
(68)

the received power can be rewritten as
\[
P_r = \left( \frac{\lambda}{4 \pi} \right)^4 \Gamma^2 G_i G_r G_{RIS}(\theta_i) G_{RIS}(\theta_r) M^2, \]
(69)

which can be equivalently written as in (12). This concludes the proof.
where, in (78), we assumed that $ho_{\text{RIS}}$ is obtained.

The three real roots of (73) can be obtained as

$$r_1^{(1)} = \frac{r_h}{2},$$

$$r_1^{(2)} = \frac{r_h + \sqrt{r_h^2 - 4(h_{\text{RIS}} - h_A)^2}}{2},$$

and

$$r_1^{(3)} = \frac{r_h - \sqrt{r_h^2 - 4(h_{\text{RIS}} - h_A)^2}}{2}.$$

Notice that $r_1^{(1)}$ minimizes (70), whereas $r_1^{(2)}$ and $r_1^{(3)}$ maximize it with the same maximum value obtained.

As far as the minimization of $\epsilon_w (r_1, h)$ is concerned, from (77) it holds that

$$\epsilon_w (r_1, h) \approx 1 - \frac{S_{\text{RIS}}}{\pi \sin^2 \left( \frac{\phi}{2} \right) \frac{r_h}{2} \sqrt{r_h^2 - 4(h_{\text{RIS}} - h_A)^2}}.$$

where, in (78), we assumed that $\frac{\phi}{2} \approx 1$ and $\tan^{-1} \left( \frac{|r_{1,h}|}{h_{\text{RIS}} - h_A} \right) \gg \frac{\phi}{2}$. From (78), it is trivial to prove that $\epsilon_w (r_1, h)$ is a monotonically increasing function of $r_1, h$. Hence, for the roots $r_1^{(2)}$ and $r_1^{(3)}$, since $r_1^{(2)} < r_1^{(3)}$, it holds that $\epsilon_w (r_1^{(2)}, h) < \epsilon_w (r_1^{(3)}, h)$. Physically, this is justified by the fact that as the RIS moves away from the TX $S_i$ increases; thus, for a fixed $S_{\text{RIS}}$, the beam waste also increases. Consequently,

$$r_1^* = r_1^{(2)}.$$

This concludes the proof.
If \( \Delta < 0 \), the cubic polynomial returns the following three real roots:

\[
\hat{r}_{1,h}^{(1)} = \frac{2}{\sqrt{3}} \sqrt{-t_1} \sin \left( \frac{1}{3} \sin^{-1} \left( \frac{3\sqrt{3}t_2}{2(\sqrt{-t_1})^3} \right) \right) - \frac{b}{3a}, \tag{89}
\]

\[
\hat{r}_{1,h}^{(2)} = \frac{2}{\sqrt{3}} \sqrt{-t_1} \sin \left( \frac{1}{3} \sin^{-1} \left( \frac{3\sqrt{3}t_2}{2(\sqrt{-t_1})^3} + \frac{\pi}{3} \right) \right) - \frac{b}{3a}, \tag{90}
\]

and

\[
\hat{r}_{1,h}^{(3)} = \frac{2}{\sqrt{3}} \sqrt{-t_1} \cos \left( \frac{1}{3} \sin^{-1} \left( \frac{3\sqrt{3}t_2}{2(\sqrt{-t_1})^3} + \frac{\pi}{6} \right) \right) - \frac{b}{3a}. \tag{91}
\]

If \( \Delta > 0 \), there is only one real root that can be obtained as

\[
\hat{r}_{1,h}^{*} = \left( \frac{-t_2}{2} + \sqrt{\Delta} \right)^{\frac{1}{3}} + \left( \frac{-t_2}{2} - \sqrt{\Delta} \right)^{\frac{1}{3}} - \frac{b}{3a}. \tag{92}
\]

Finally, if \( \Delta = 0 \), the 3 real roots are

\[
\hat{r}_{1,h}^{(5)} = -2 \left( \frac{t_2}{2} \right)^{\frac{1}{3}} - \frac{b}{3a}, \tag{93}
\]

\[
\hat{r}_{1,h}^{(6,7)} = \left( \frac{t_2}{2} \right)^{\frac{1}{3}} - \frac{b}{3a}. \tag{94}
\]

In all 3 different cases of \( \Delta \) one of the roots maximizes \( H(\hat{r}_{1,h}) \), while the others minimize it. This solution imposes the RIS to be closer to the RX than the TX, in order to exploit the maximum number of RIS’s RUs. In Fig. 7 we illustrate the value of \( r_{1,h}^{*} \), denoted by \( r_{1,h}^{*} \), that maximizes \( H(\hat{r}_{1,h}) \) as a function of \( p \). As we observe from Fig. 7 throughout the range of \( p \) the optimal value of \( r_{1,h}^{*} \) is relatively close to \( r_h \), which means that the RIS should optimally be placed closer to the RX than the TX. This concludes the proof.

**Appendix F**

**Proof of Lemma 2**

For the special case in which \( p = 1 \), (83) reduces to a second order polynomial with roots that are given by (31).

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