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RESEARCH ARTICLE

Design Strategy for Compact Ridge-Waveguide Power Dividers With Arbitrary Split Ratios

SOFIE LENDERS^{ID}, (Graduate Student Member, IEEE),
SAMUEL RIMBAUT^{ID}, (Graduate Student Member, IEEE),
BRAM HOFACK^{ID}, (Graduate Student Member, IEEE),
SAM LEMEY^{ID}, (Member, IEEE), AND **HENDRIK ROGIER**^{ID}, (Senior Member, IEEE)

IDLab, Electromagnetics Group, Department of Information Technology, Ghent University/imec, 9052 Ghent, Belgium

Corresponding author: Sofie Lenders (Sofie.Lenders@UGent.be)

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ABSTRACT A vertically integrable, ridge-waveguide-based power divider topology for achieving arbitrary split ratios is presented, along with its detailed design procedure. The divider's ridge-waveguide cross-section enables a compact design with minimal transmission loss, while its vertical topology supports modular deployment within multi-stage feed networks for antenna arrays. The proposed method achieves power split ratios between 0.05 (1:20) and 1 (1:1) using a co-design strategy of a dogbone aperture and a modified ridge geometry. By separating the power divider behavior into two distinct regimes, precise control of the split ratio is achieved while the design space remains compact, requiring no more than four geometrical parameters. Design rules are formulated accordingly, relating the targeted split ratio to each of the relevant parameters. To validate the design methodology, five power dividers were fabricated for the fifth-generation (5G) n257 frequency band between 26.5GHz and 29.5GHz, with representative split ratios of 0.05, 0.25, 0.5, 0.9 and 1. Their average measured split ratios equal 0.05, 0.24, 0.51, 0.91 and 0.99, while all are matched consistently with their reflection coefficient below -10dB within the n257 band, thereby confirming the effectiveness of the proposed design procedure.

INDEX TERMS Power dividers, unequal split, ridge waveguide components, design of microwave devices, millimeter-wave passive components, analysis of microwave passive components.

I. INTRODUCTION

Growing throughput requirements in modern mobile communication networks are driving continuous advancements in the design of multi-antenna systems [1], [2], [3]. Through large-array configurations, these systems enable highly directive beam steering, thereby mitigating adverse propagation conditions in the fifth- and sixth-generation (5G/6G) frequency bands [4], [5], [6]. As array sizes scale to achieve such performance, system integration becomes a critical challenge, necessitating tight integration of radio-frequency (RF) integrated-circuits (ICs) [7], as well as effective suppression of spurious radiation through low-loss, self-shielding feed networks [8], [9].

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Full two-dimensional (2D) beam steering has been successfully demonstrated by interfacing each antenna element of such arrays with a beamforming IC [10], [11]. However, this approach can lead to rather expensive solutions and may become impractical, or even unfeasible, as operating frequencies increase and antenna element sizes decrease. Beamforming ICs do not scale proportionally, limiting the achievable inter-element spacing, which can lead to grating lobes during beam steering [12]. Moreover, for applications such as cellular backhaul [13] and automotive radar [14], azimuth-only scanning is often sufficient.

Therefore, single-axis beam steering offers a significant reduction in hardware complexity and cost compared to full-2D beam steering implementations [15]. To this end, a passive 2D antenna array is subdivided into one-dimensional (1D) subarrays, which are controlled by active beam-steering ICs [16]. Achieving low sidelobe levels (SLLs) in the

subarray radiation patterns requires feed networks that enable precise amplitude tapering [17], [18], [19]. Special care must be devoted to efficient routing and suppression of undesired radiation [20], [21], while maintaining tight spacing of the subarrays to ensure a wide grating-lobe-free steering range [22]. Realizing such compact, amplitude-tapered feed networks necessitates a power divider topology capable of achieving a broad range of unequal power-split ratios, while remaining highly compact and exhibiting minimal losses.

Although widely used at sub-6 GHz frequencies [23], [24], [25], [26], planar transmission-line-based dividers suffer from high losses, limited self-shielding, and structural constraints due to their planar nature. This restricts their integration in splitter networks for high-frequency beam-steering systems.

Air-filled rectangular waveguide (WG)-based power dividers provide a low-loss solution with strong self-shielding [27], [28], [29]. However, their wide cross-section constrains the achievable inter-subarray spacing of the antenna array deployed on top of the feed distribution network, resulting in grating lobes during beam steering. As an alternative, ridge-waveguides (RiWGs) and RiWG-based power dividers offer comparable low-loss performance while supporting a significantly reduced footprint [29], as demonstrated in [30] for an 8×8 slot-antenna array. Nevertheless, the horizontal topology of the employed power dividers still limits compact and modular power distribution in 2D array architectures.

Therefore, we propose a vertically integrable, RiWG-based power divider topology capable of achieving a wide range of split ratios, and suitable for integration into a multi-stage power divider network situated between a 2D antenna array in front and the active steering electronics behind. The proposed divider design preserves a constant cross-section along the RiWG, optimized for compactness and low-loss performance. Since no resistive components are used, the divider's isolation is limited to a value that is quantified in this paper as a function of the power split ratio. In addition, a comprehensive design flow is presented to achieve any split ratio between 0.05 (1:20 power division) and 1 (equal 1:1 power division) with only four geometrical design parameters. The methodology is supported by frequency-scalable design rules, differentiating between two regimes with distinct operating principles. Through the design and characterization of five prototypes, fully covering the proposed range of split ratios, these rules are validated experimentally for the fifth-generation (5G) n257 frequency band (26.5–29.5 GHz) [31].

In literature, the feasibility of vertically integrable power divider topologies has already been fruitfully demonstrated. A four-level stripline-based combiner is shown in [32], while RiWG-based vertical power dividers are utilized in [33] and [34]. These, however, only provide equal power division. In [35], both an unequal and equal power divider are realized in ridge-gap waveguide technology, successfully reducing the SLLs of a dual-polarized slot antenna array. Symmetric

slot shapes are used, and unequal power division is attained by changing the waveguide cross-sections. However, this necessitates multiple matching sections with different ridge heights, complicating the design and fabrication process. Furthermore, discussions about vertical-topology power dividers in literature are typically limited to a predefined set of split ratios, which are only relevant for the considered amplitude tapering scheme. Hence, a systematic design procedure for the unequal-split divider topology is still lacking. In contrast, this work develops a general methodology and theoretical framework for the design and analysis of low-loss, compact, vertical-topology power dividers, achieving a wide range of unequal split ratios.

The remainder of this paper is organized as follows: Section II introduces the proposed power divider topology, its design specifications, and the underlying theoretical framework. Section III presents the design methodology and corresponding rules, followed by numerical validation through a representative design example. Finally, prototype fabrication and measurement results are discussed in Section IV, validating the proposed procedure.

II. POWER DIVIDER TOPOLOGY FOR ARBITRARY SPLIT RATIO

The proposed power divider topology is introduced in Section II-A, describing its operation principle and compiling its associated set of design criteria by analyzing a typical application scenario. In Section II-B, an analytical model is proposed to quantify isolation and provide an equivalent circuit formulation.

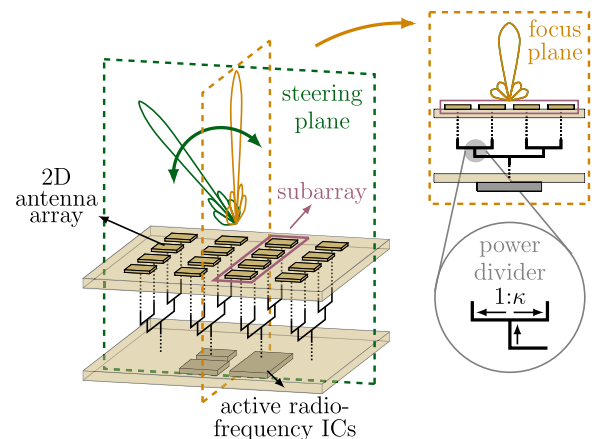


FIGURE 1. Use case for the proposed power divider: multi-antenna system realizing one-dimensional (1D) beam steering in the steering plane, with a narrow beam and low sidelobe levels (SLLs) in the focus plane, through a compact power divider network.

A. DESIGN CRITERIA AND TOPOLOGY

Fig. 1 considers a representative multi-antenna system that realizes 1D beam steering in the steering plane, supporting a wide range of steering angles without grating lobes in its field of view. Simultaneously, to compensate for

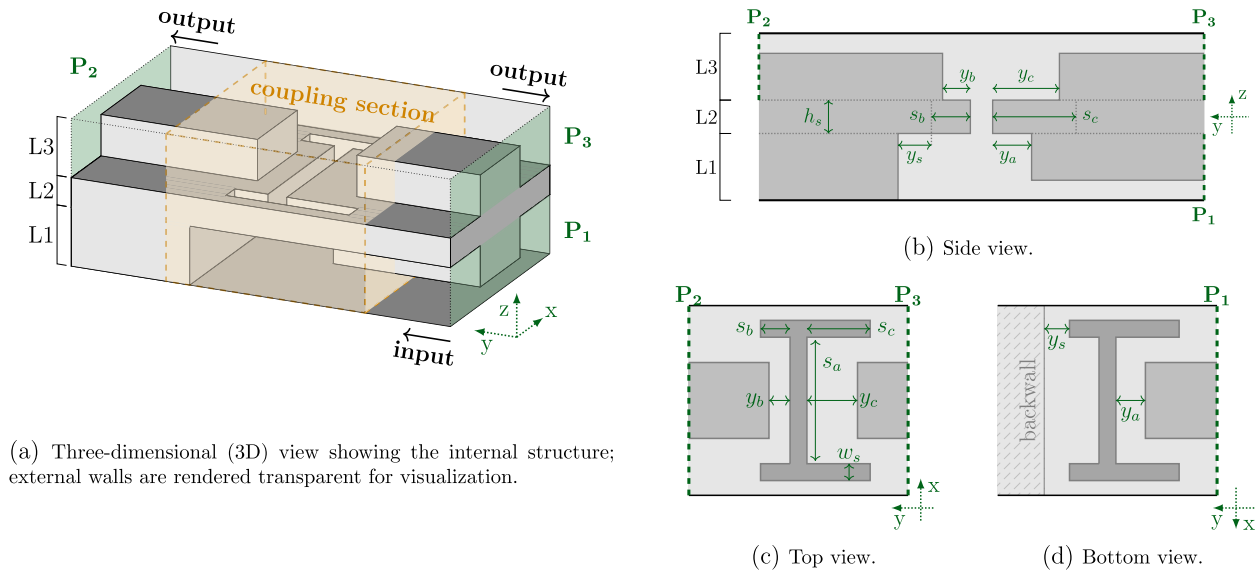


FIGURE 2. Compact vertically-integrated ridge-waveguide (RiWG) power divider topology: (a) 3D view, (b) side view, (c) top view, and (d) bottom view. The relevant parameters of the arbitrary-split coupling section are indicated: the slot width w_s , the slot lengths s_a , s_b , and s_c , the ridge shifts y_a , y_b , and y_c , the stub backwall distance y_s , and the slot width w_s and height h_s .

the high propagation losses at millimeter-wave (mmWave) frequencies through antenna gain, the beam should typically be strongly confined in the focus plane orthogonal to the steering plane, with minimal SLLs. The system should also realize this with minimal material losses and spurious radiation.

In such a system architecture, specific constraints are imposed on the power dividers, which are the fundamental building blocks of the array's feed distribution network in the focus plane (Fig. 1). Essential for this network is the vertical topology of the power dividers, which enables seamless integration between the passive, 2D antenna array (top plane in Fig. 1) and the underlying active radio-frequency ICs (bottom plane in Fig. 1), while realizing 1D beam steering in a modular architecture [36], [37]. To implement this vertical structure with minimal transmission loss and spurious radiation, a waveguide-based topology is preferred, realized by means of milling [38], metallic additive manufacturing [39], or air-filled printed-circuit-board (PCB) fabrication [40], [41].

Furthermore, to enable a full $\pm 90^\circ$ steering range without grating lobes in the steering plane, the width of each individual power divider must remain below half a wavelength in the entire frequency band of operation. Hence, the power divider topology must be compact, while ensuring minimal conductive and radiation losses.

Finally, to form a narrow beam with low SLLs in the focus plane, appropriate amplitude tapering is required across each 1D subarray. Various tapering schemes can be applied, each offering different levels of SLL suppression. For instance, Dolph–Chebyshev tapering enables equal-level sidelobes below -50 dB by employing power dividers with unequal split ratios ranging from 1:14 to 1:1 for an eight-element antenna

array. To support such tapering schemes, the power divider topology must be capable of realizing a highly unequal power division that remains frequency-stable across the operational bandwidth. A fractional bandwidth (FBW) of 10% (defined at a return loss of $|S_{11,F}| = -10$ dB) around the center operation frequency f_o is set as a minimal design objective, since this provides full coverage of several relevant 5G and ISM frequency bands [42]. Throughout this FBW, a split-ratio range from 1:20 to 1:1 is targeted, to support multiple tapering strategies.

To enable designers to meet such a diverse set of constraints, the compact, vertical, arbitrary-split power divider topology shown in Fig. 2a is proposed. The structure is fully implemented in RiWG-technology, with waveguides of fixed cross-section. This ensures both compactness and low loss, as well as straightforward cascading of multiple power dividers in a multi-stage divider network. The power injected at input port P_1 through the bottom-layer (L1) RiWG is divided between output ports P_2 and P_3 in the top-layer (L3) RiWGs by a central coupling section. The power division in this coupling section is governed by the interplay between a generalized-dogbone-shaped slot and the waveguide's ridge configuration. The joint manipulation of both elements is crucial for achieving a broad and controllable range of split ratios: a conventional rectangular aperture cannot provide the required flexibility in terms of control of the coupling strength between the vertically stacked input and output RiWGs, whereas without adaptation of the ridges, there is insufficient control to achieve small ratios. In what follows, a detailed co-design strategy is proposed.

The generalized-dogbone-shaped slot in layer L2 features a uniform width w_s and lengths s_a , s_b and s_c (Fig. 2c). By varying the leg lengths s_b and s_c (Fig. 2b), the split ratio κ

can be tuned for values close to 1. To realize smaller κ -values, the ridge configuration must be adapted, by introducing ridge shift parameters y_a , y_b and y_c , which define the distance between the ridges and the slot (Figs. 2b and 2c). The magnitude of the P₂-side ridge shift y_b mainly influences the split ratio, whereas setting $y_a = y_c = 0$ minimizes the number of design parameters. For all κ -values, the core slot length s_a (Fig. 2c) ensures matching, together with a stub section of length y_s (Figs. 2b and 2d) in the bottom layer L1.

In Section III, it will be shown that the coupling section can produce a wide range of split ratios with no more than four variable geometrical parameters. A design procedure to optimize each of these parameters is provided, facilitating κ -values from 0.05 to 1. First, however, Section II-B discusses an analytical model introduced to quantify isolation and to develop an equivalent circuit.

B. ANALYTICAL MODEL

In an effort to limit size, fabrication complexity and cost, no lossy components are inserted in the power divider topology, resulting in a limited isolation between the two output ports P₂ and P₃. The exact level of isolation is a function of the split ratio κ and can be computed analytically by imposing on the power divider's S-parameters the fundamental properties [43] of losslessness ($\mathbf{S}^H \mathbf{S} = \mathbf{I}$), reciprocity ($\mathbf{S}^T = \mathbf{S}$), and perfect matching at input port P₁ ($|S_{11}| = 0$). Setting the arbitrary real-valued split ratio to $\kappa = |S_{21}|^2 / |S_{31}|^2 (< 1)$ and fixing the port planes such that the S-parameter phases $\angle S_{12} = \angle S_{13} = \angle S_{23} = 0$, the following real-valued S-matrix $\mathbf{S}_0$ is obtained at the frequency f_m of perfect matching:

$$\mathbf{S}_0 = \begin{pmatrix} 0 & \sqrt{\frac{\kappa}{1+\kappa}} & \sqrt{\frac{1}{1+\kappa}} \\ \sqrt{\frac{\kappa}{1+\kappa}} & -\frac{1}{1+\kappa} & \frac{\sqrt{\kappa}}{1+\kappa} \\ \sqrt{\frac{1}{1+\kappa}} & \frac{\sqrt{\kappa}}{1+\kappa} & -\frac{\kappa}{1+\kappa} \end{pmatrix}. \quad (1)$$

For arbitrary locations of the three-port's port planes, this generalizes to (2).

$$\mathbf{S} = \mathbf{D} \cdot \mathbf{S}_0 \cdot \mathbf{D} = \begin{pmatrix} 0 & \sqrt{\frac{\kappa}{1+\kappa}} e^{-j l_{e12}} & \sqrt{\frac{1}{1+\kappa}} e^{-j l_{e13}} \\ \sqrt{\frac{\kappa}{1+\kappa}} e^{-j l_{e21}} & -\frac{1}{1+\kappa} e^{-j l_{e22}} & \frac{\sqrt{\kappa}}{1+\kappa} e^{-j l_{e23}} \\ \sqrt{\frac{1}{1+\kappa}} e^{-j l_{e31}} & \frac{\sqrt{\kappa}}{1+\kappa} e^{-j l_{e32}} & -\frac{\kappa}{1+\kappa} e^{-j l_{e33}} \end{pmatrix} \quad (2)$$

Here, $\mathbf{D}$ represents an arbitrary port translation by lengths l_1 , l_2 and l_3 at ports P₁, P₂ and P₃, respectively. The corresponding electrical lengths l_{eij} equal $\beta_i l_i + \beta_j l_j$, for $i, j \in \{1, 2, 3\}$, with β_1 , β_2 and β_3 being the propagation constants in the corresponding RiWG sections ($\beta_1 = \beta_2 = \beta_3$ in our fixed-cross-section design).

Hence, the κ -dependent achievable isolation between the output ports P₂ and P₃ is given by

$$-20 \log |S_{23}| = 20 \log (\sqrt{\kappa} + 1/\sqrt{\kappa}). \quad (3)$$

This expression is monotonically decreasing for $\kappa \in [0, 1]$. Consequently, smaller split ratios will yield a larger isolation. This is valid for any κ -split power divider adhering to the fundamental properties of losslessness, reciprocity and being perfectly matched at one port.

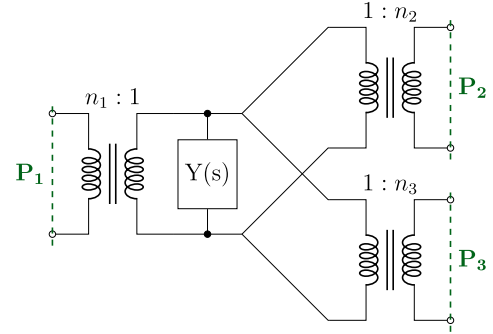


FIGURE 3. Equivalent circuit of the proposed power divider realizing a split ratio κ , consisting of three transformers with turn ratios n_1 , n_2 and n_3 , and a general admittance $Y(s)$ to model the frequency dependence.

In conventional unequal power divider design, a κ -split power division as described by (1) is achieved by altering the impedance levels of the input and output transmission-line sections, which may be implemented by altering the waveguide cross-section [44]. As mentioned before, however, constant cross-sections are preferred for deployment in a compact multi-stage power distribution network. Hence, the proposed topology maintains a fixed RiWG cross-section, and instead relies on judicious slot design to implement the required impedance transformation. This is modeled by an equivalent circuit (Fig. 3) with three transformers having turn ratios n_1 , n_2 , and n_3 satisfying

$$\begin{cases} n_2 = \sqrt{\kappa} n_3 \\ n_1 = \sqrt{1 + \kappa} n_3. \end{cases} \quad (4)$$

In the circuit, a general admittance $Y(s)$ is introduced to account for the frequency-dependent behavior of the power divider. At the frequency f_m of perfect matching, we have that $Y(j2\pi f_m) = 0$, yielding the S-parameters given in (1).

In Section III-B, it is shown that a first-order resonant RLC-tank effectively matches the proposed design in the 10% FBW around the center operation frequency f_o . Note that, in general, the latter frequency may differ from the frequency of perfect matching f_m .

III. DESIGN METHODOLOGY

Starting from the topology proposed in section II, we now outline a comprehensive design methodology that enables designers to realize split ratios between 1:20 and 1:1 in a gradual manner, at any specified center frequency, by continuously varying only four well-chosen geometrical parameters. First, a general design flow is presented for dimensioning the RiWG cross-section and the coupling section. Then, detailed design rules are put forward, relating each relevant geometrical parameter to the intended split ratio. Finally, the methodology is verified for a power divider

operating in the 5G n257 frequency band between 26.5 GHz and 29.5 GHz, confirming the validity of the design procedure by demonstrating frequency-stable power-dividing behavior.

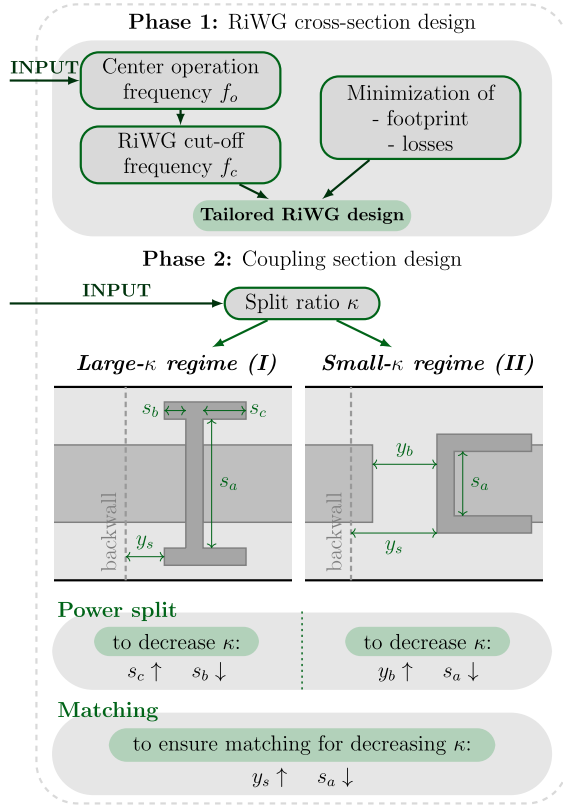


FIGURE 4. Power divider design flow for arbitrary split ratio κ and center operation frequency f_o , consisting of two phases, and putting forward qualitative conditions on the geometrical design parameters (being s_a , y_s , s_b , s_c for the *large- κ regime*, and s_a , y_s , y_b for the *small- κ regime*).

A. DESIGN FLOW AND DESIGN RULES

Fig. 4 visualizes the design process, subdivided into two main phases. In phase one, the RiWG cross-section (Fig. 5) is optimized to achieve the best trade-off between size (w) and conductor loss within the frequency band from $0.95f_o (= f_{\min})$ to $1.05f_o (= f_{\max})$, covering a 10% FBW around the center operation frequency f_o .

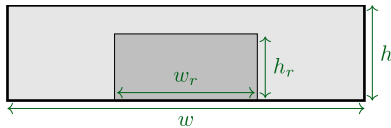


FIGURE 5. RiWG cross-section design, with the relevant parameters indicated: ridge width w_r and height h_r , and waveguide width w and height h .

To ensure linear-phase performance within this frequency band, the RiWG's first-order mode cut-off frequency f_c is set to $0.75f_o$, positioning it well below $f_{\min} (= 0.95f_o)$. Previous analyses have demonstrated that a ridge width (w_r) and height (h_r) corresponding to 40% and 70% of the waveguide width

(w) and height (h), respectively, provide an optimal trade-off between structural compactness and conduction loss [45]. Under those conditions, the waveguide width w must equal $0.4\lambda_o$ to realize the required cut-off frequency, where λ_o represents the free-space wavelength at the center operation frequency f_o . For vertical compactness, the waveguide height h is set to $0.1\lambda_o$. Hence, the following design rules are proposed:

$$\begin{cases} w = 0.4\lambda_o \\ h = 0.1\lambda_o \end{cases} \quad \begin{cases} w_r = 0.4w = 0.16\lambda_o \\ h_r = 0.7h = 0.07\lambda_o \end{cases} \quad (5)$$

In phase two of the design procedure, the arbitrary-split coupling section (Fig. 2a) is designed, tuning a minimal number of geometrical design parameters to obtain the desired split ratio κ . Specifically, the width w_s of the dogbone-shaped slot in layer L2 is fixed to a constant value, balancing the manufacturability constraints with the matching requirements of the coupling section. The ridge shifts y_a (bottom ridge) and y_c (ridge at port P₃ side) are set to zero (Figs. 2b, 2c and 2d). To cover the specified range of split ratios – from highly asymmetric (1:20) to equal (1:1) – two different power dividing mechanisms are exploited, corresponding to two distinct operational regimes, denoted as *large- κ regime (I)* and *small- κ regime (II)*. For both regimes, Fig. 4 shows the corresponding top views of the output RiWGs as evolutions of the coupling section shown in Fig. 2c.

The upper limit of the *large- κ regime (I)*, being $\kappa = 1$, and hence $|S_{21}|^2 = |S_{31}|^2$, corresponds to equal power division. This is achieved through a symmetric slot geometry ($s_b = s_c$, $y_b = 0$), with the stub backwall in layer L1 touching the slot legs ($y_s = 0$) for matching, as shown on the left in Fig. 6. To obtain a lower ratio κ , a power imbalance must be created between output ports P₃ and P₂ ($|S_{31}|^2 > |S_{21}|^2$). This asymmetry is introduced by varying the dogbone geometry, without altering the ridge configuration at the P₂-side ($y_b = 0$). Specifically, the P₃-facing slot legs are elongated ($s_c \uparrow$), whereas the slot legs at the P₂-side are shortened ($s_b \downarrow$), allowing stronger field coupling towards port P₃. As a result, the slot gradually evolves from a symmetric-dogbone-shape to a U-shape, as illustrated in Fig. 6. Matching is ensured by simultaneously tuning the overall slot length and the stub section distance in layer L1 ($s_a \downarrow$, $y_s \uparrow$).

When the U-shaped slot fully encloses the ridge of the output RiWG at the P₃-side, κ reaches a value of 0.688. To reduce κ beyond this point, an alternative approach is required, marking the onset of the *small- κ regime (II)*. The slot lengths s_b and s_c now remain constant, while asymmetry is introduced by shifting the ridge of the P₂-side RiWG away from the slot ($y_b \uparrow$). The core slot length is further reduced ($s_a \downarrow$) to have a higher field concentration at the center of the RiWG at the P₃-side. This evolution is visualized in Fig. 7. The same matching mechanisms as in the *large- κ regime* are employed.

In short, both regimes rely on no more than four variable geometrical parameters: y_s and s_a ensuring matching, s_b and

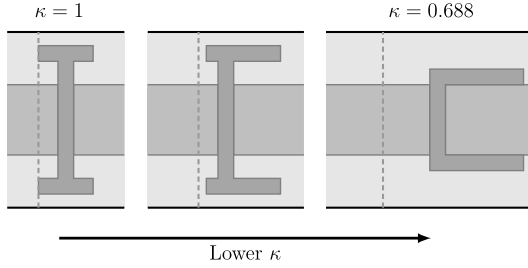


FIGURE 6. Slot-shape evolution for decreasing split ratio κ in the *large- κ regime (I)*, with geometrical design parameters s_b , s_c , s_a and y_s , as indicated in Fig. 2c.

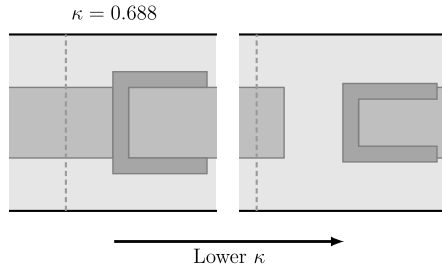


FIGURE 7. Slot-shape evolution for decreasing split ratio κ in the *small- κ regime (II)*, with geometrical design parameters y_b , s_a and y_s , as indicated in Fig. 2c.

s_c fixing the split ratio in the *large- κ regime*, and s_a and y_b fixing the split ratio in the *small- κ regime*. Supporting the general design flow, the following set of practical design rules relates the variable geometrical parameters to the desired power split ratio κ for both operating regimes, being valid for the proposed optimal RiWG cross-section of Fig. 5:

For *large- κ regime (I)* ($\kappa > 0.688$):

$$\begin{aligned} y_a / \lambda_o &= 0 \\ y_b / \lambda_o &= 0 \\ y_c / \lambda_o &= 0 \\ y_s / \lambda_o &= -0.365\kappa^3 + 0.846\kappa^2 - 0.846\kappa + 0.366 \\ s_a / \lambda_o &= -1.648\kappa^3 + 3.093\kappa^2 - 1.093\kappa - 0.0081 \\ s_b / \lambda_o &= -1.346\kappa^2 + 2.773\kappa - 1.396 \\ s_c / \lambda_o &= -8.37\kappa^3 + 23.051\kappa^2 - 21.47\kappa + 6.82 \\ w_s / \lambda_o &= 0.0373 \end{aligned} \quad (6)$$

For *small- κ regime (II)* ($\kappa \leq 0.688$):

$$\begin{aligned} y_a / \lambda_o &= 0 \\ y_b / \lambda_o &= 30.861\kappa^6 - 84.701\kappa^5 + 93.503\kappa^4 \\ &\quad - 53.412\kappa^3 + 17.097\kappa^2 - 3.195\kappa + 0.333 \\ y_c / \lambda_o &= 0 \\ y_s / \lambda_o &= -0.365\kappa^3 + 0.846\kappa^2 - 0.846\kappa + 0.366 \\ s_a / \lambda_o &= 0.1\kappa^2 + 0.0946\kappa + 0.0554 \\ s_b / \lambda_o &= 0 \\ s_c / \lambda_o &= 0.238 \\ w_s / \lambda_o &= 0.0373 \end{aligned} \quad (7)$$

These expressions have been empirically derived through judicious full-wave simulation and optimization, using the electromagnetic field simulator CST Studio Suite, combined with curve fitting. They remain valid for κ -values from 0.05 (corresponding to a 1:20 power division) to 1 (equal power division). By expressing all dimensions relative to the wavelength λ_o , the design rules are applicable for any arbitrary center operation frequency f_o , with the proposed formulas ensuring a -10 dB-FBW of at least 10% around this center frequency.

Out of these parameters, the ridge shift at the P₂-side (y_b) is the most crucial for accurately controlling the split ratio, especially for κ -values below 0.2. Hence, it exhibits the most complex behavior, best fitted by a sixth-order polynomial, while all other relations are captured with sufficient accuracy by polynomials not exceeding third order. These relationships, supporting the design flow of Fig. 4, are visualized in Fig. 8. They closely fit the data points that were obtained through direct full-wave simulation of the power divider with corresponding split ratio κ .

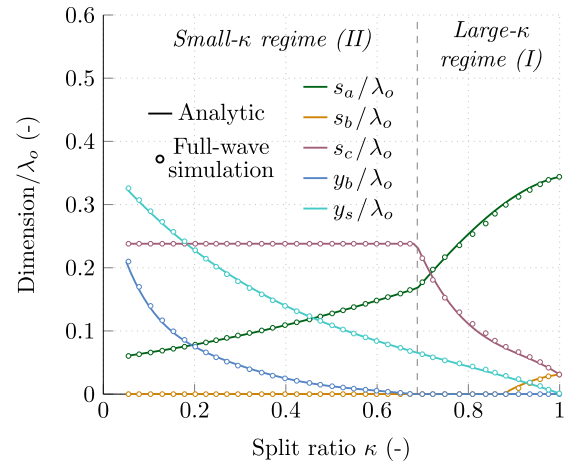


FIGURE 8. Design rules for arbitrary split ratio κ between 0.05 and 1, fitted to the simulated designs, realizing a -10 dB-FBW of at least 10% around the center operation frequency f_o , with varying geometrical design parameters s_a , y_s , s_b , s_c for the *large- κ regime (I)*, and s_a , y_s , y_b for the *small- κ regime (II)*, normalized to the corresponding center wavelength λ_o .

B. DIMENSIONING THE EQUIVALENT CIRCUIT MODEL

As described in Section II, the frequency variation of the power division realized by the proposed design rules (6)–(7) can be modeled in the equivalent circuit (Fig. 3) by the admittance $Y(s)$. The target admittance behavior is fitted onto the EM simulation of the κ -split design (Fig. 2).

First, the frequency range $[f_1, f_2]$ where the design achieves a 10% FBW is determined. The corresponding reflection coefficient level is denoted as $|S_{11,F}|$:

$$\text{FBW} = \frac{f_2 - f_1}{\frac{f_1 + f_2}{2}} = 10\%,$$

$$\text{with } |S_{11}(f_1)| = |S_{11}(f_2)| = |S_{11,F}|. \quad (8)$$

To model this frequency behavior in the equivalent circuit, we implement the admittance $Y(s)$ as a first-order resonant parallel RLC-tank expressed by $Y(s) = \frac{1}{R} + sC + \frac{1}{sL}$. Its resonance frequency f_r is chosen to lie at the midpoint between f_1 and f_2 , such that $f_r = \frac{f_1 + f_2}{2}$. This is ensured by the relation (9b) between the inductance L and the capacitance C . The reflection coefficient at f_r is set at 6 dB below $|S_{11,F}|$ (being $\frac{1}{2}|S_{11,F}|$) through the resistance value R given by (9a), where Z_0 equals the 50 Ω reference impedance. Lastly, the capacitance value C specified by (9c) fixes the reflection coefficient at the boundaries f_1 and f_2 to $|S_{11,F}|$. A full derivation of these expressions is provided in Appendix A.

$$R = \frac{\frac{2}{|S_{11,F}|} - 1}{2(1 + \kappa)n_3^2} Z_0 \quad (9a)$$

$$L = \frac{1}{(2\pi f_r)^2 C} \quad (9b)$$

$$C = \frac{420 / 41}{2\pi f_r R} \sqrt{\frac{3}{1 - |S_{11,F}|^2}} \quad (9c)$$

Fig. 9 shows the progression of $|S_{11,F}|$ and f_r as a function of the split ratio κ , obtained through EM simulation of the κ -split power divider design. When combined with the expressions above, the plotted data allow designers to reproduce the complete behavior of the proposed power divider in a first-order approximation, thereby eliminating the need to conduct full-wave simulations.

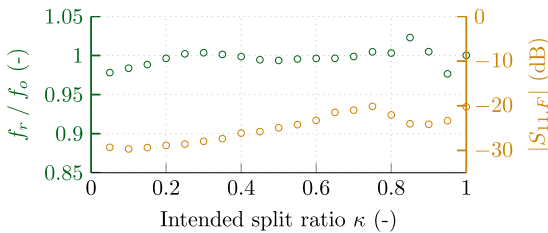


FIGURE 9. Reflection coefficient $|S_{11,F}|$ at the power divider design's 10% FBW, and the corresponding normalized midpoint frequency f_r / f_o , for split ratios between 0.05 and 1, determined from EM simulation.

To demonstrate the validity of the equivalent circuit, Fig. 10 compares its frequency behavior with that of the EM-simulated design, for κ -values of 0.2, 0.5 and 0.8. The corresponding circuit parameters, in accordance with Fig. 3, are listed in Table 1. Additionally, Fig. 11 shows the isolation characteristics for κ -values between 0.05 and 1.

These plots clearly demonstrate close agreement between the design rules presented in (6)–(7) and the proposed equivalent circuit - both in terms of power division characteristics and frequency-dependent behavior.

C. DESIGN EXAMPLE AT 28 GHZ

To showcase the design flow, representative power dividers are conceived to operate in the common 5G n257 band centered around 28 GHz, ranging from 26.5 GHz ($f_{\min}$) to 29.5 GHz ($f_{\max}$). They are designed by applying the

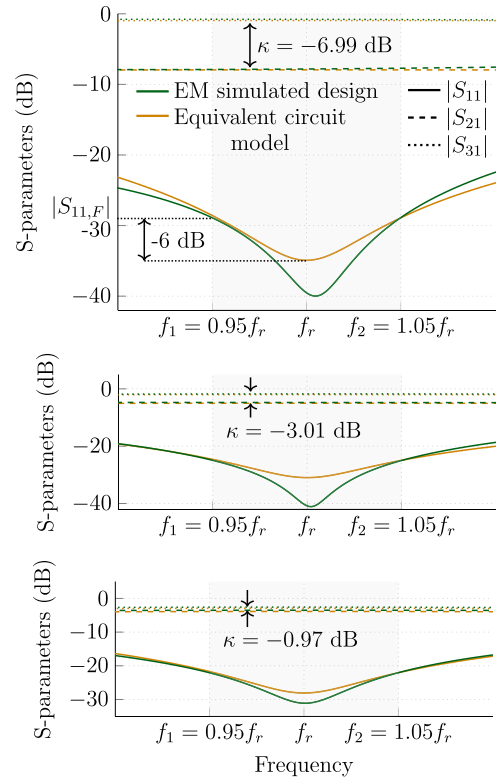


FIGURE 10. Simulated S-parameters of the power divider design (Fig. 8), and of its equivalent circuit model (Fig. 3, with $Y(s) = \frac{1}{R} + sC + \frac{1}{sL}$), for κ -values of 0.2 (−6.99 dB), 0.5 (−3.01 dB) and 0.8 (−0.97 dB).

TABLE 1. Equivalent circuit component values (Fig. 3, with $Y(s) = \frac{1}{R} + sC + \frac{1}{sL}$), for κ -values of 0.2, 0.5 and 0.8.

κ	0.2	0.5	0.8
R (k Ω)	1.13	0.57	0.34
$L \cdot f_r$ (nH-GHz)	10.2	5.13	3.02
$C \cdot f_r$ (pF-GHz)	2.49	4.94	8.38
n_1 (-)	1.095	1.225	1.342
n_2 (-)	0.447	0.707	0.894
n_3 (-)	1.0	1.0	1.0

design rules presented in (6) and (7) for split ratios ranging from 0.05 to 1, with a discrete step size of 0.025. Fig. 12 shows the corresponding simulated split ratios $\hat{\kappa}$ at the lower, center and upper frequency of the operational band, as well as the maximum in-band reflection coefficient. The effectively realized split ratios ($\hat{\kappa}$) differ only minimally from the intended values (κ), with a maximal in-band deviation of 0.0172 for a κ -value of 0.925. In terms of matching, for κ below 0.5, the reflection coefficient at port P1 equals maximally −20 dB within the frequency band of operation, and it remains well below the −10 dB objective throughout the n257 band for all κ .

In some applications, the phase behavior of the power divider may also be of importance. Especially when the dividers are used as part of an array feeding network, control over the relative phases between antenna elements is required.

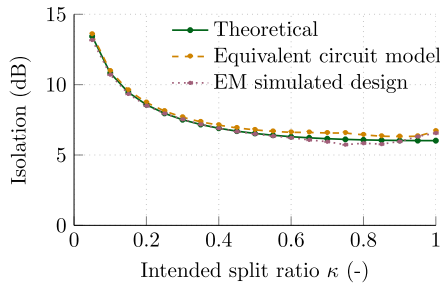


FIGURE 11. Simulated isolation $-20 \log_{10} |S_{23}|$ between the output ports of the power divider design (Fig. 8), and of its equivalent circuit model (Fig. 3), compared to the theoretically predicted isolation according to (3), for κ -values between 0.05 and 1.

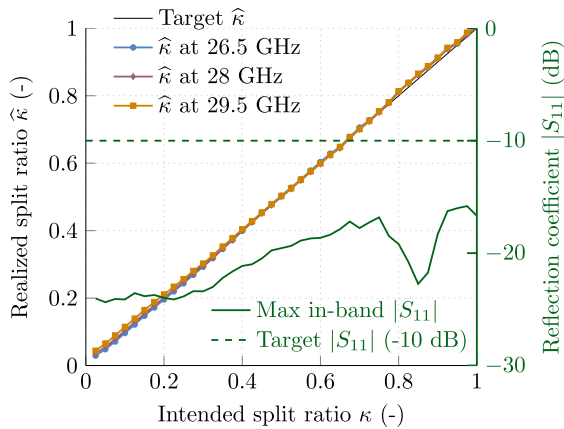


FIGURE 12. Simulated split ratio at $f_{\min}=26.5$ GHz, $f_o=28$ GHz and $f_{\max}=29.5$ GHz and matching at port P_1 , for the n257 power divider designs according to design rules (6)–(7) and with κ ranging from 0.05 to 1.

Therefore, the relative phase response of the two output ports is being considered.

When $\kappa = 1$, resulting in equal power division between the two symmetric output ports, the divider behaves similarly to the conventional E-plane waveguide T-junction [44]. Since the electric field in the dogbone-shaped coupling slot is perpendicular to the symmetry plane of the output waveguides, the electric field distribution created in these waveguides exhibits an odd symmetry. This yields equal-amplitude output signals with opposite polarities, as indicated in Fig. 13. With RiWG sections of equal length interfacing to ports P_2 and P_3 , the phase difference between these ports therefore equals 180° .

When introducing an amplitude imbalance into the coupling section ($\kappa < 1$), a different phase relation is obtained at those ports. If considered necessary in the application context, the nominal 180° phase difference can be restored at the center operation frequency f_o , by adjusting the output RiWG section lengths. In this way, a stable 180° phase difference can be ensured at f_o for all κ 's, as shown in Fig. 14 for a selection of κ -values. The corresponding length difference gives rise to a phase behavior that is slightly frequency dependent over the targeted n257 band. To further examine this phase behavior, Fig. 15 shows the simulated data points of the phase

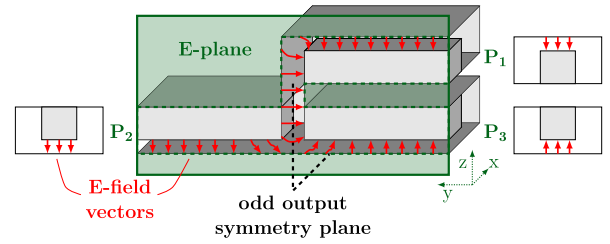


FIGURE 13. Proposed equal-split power divider ($\kappa = 1$) with odd electric field symmetry in its output waveguides leading up to ports P_2 and P_3 , resulting in a 180° output phase difference.

difference for κ between 0.1 and 1.0, at the lower, center and upper frequencies of the n257 band. For a split ratio of $\kappa = 1.0$, the phase difference remains 180° across the entire frequency band. However, as the split ratio decreases, progressively longer output RiWG sections are required to maintain a 180° difference at 28 GHz. This results in an increasing phase deviation from this nominal 180° across the frequency band, with a maximum of $\pm 9.5^\circ$ at $f_{\min}$ and $f_{\max}$, and for all split ratios.

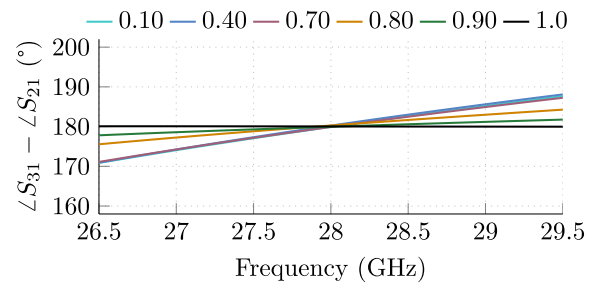


FIGURE 14. Simulated output phase difference between ports P_2 and P_3 , for the n257 power divider designs ($\kappa = 0.1, 0.4, 0.7, 0.8, 0.9$, and 1.0) according to design rules (6)–(7), after ensuring a 180° phase difference at $f_o = 28$ GHz by properly shifting the P_2 and P_3 port planes.

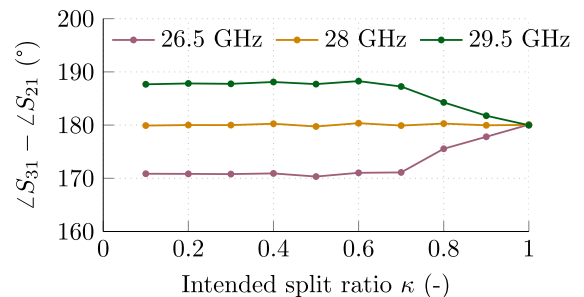


FIGURE 15. Simulated output phase difference between ports P_2 and P_3 at $f_{\min}=26.5$ GHz, $f_o=28$ GHz and $f_{\max}=29.5$ GHz, for the n257 power divider designs according to design rules (6)–(7) and with κ ranging from 0.05 to 1, after implementing a 180° phase difference at f_o by properly shifting the P_2 and P_3 port planes.

IV. PROTOTYPING AND MEASUREMENT

Based on the 28 GHz design example presented in Section III-C, five power dividers were fabricated and characterized, to experimentally validate the proposed design rules. Both the *large-* and *small-κ regime* are covered, with

carefully selected split ratios of 0.05, 0.25, 0.5, 0.9 and 1. A general overview of the prototype manufacturing and assembly is first provided, followed by a discussion of the measurement results.

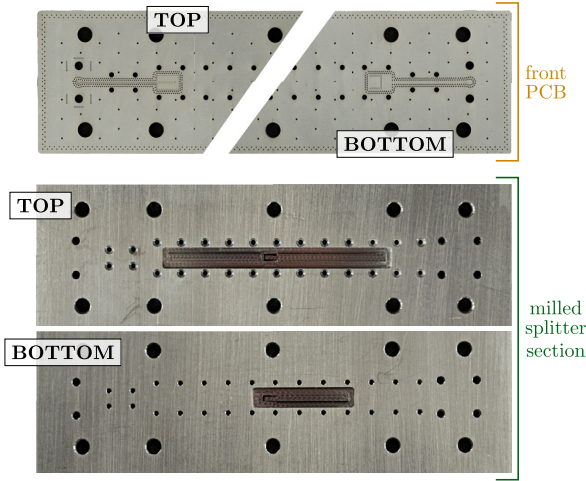


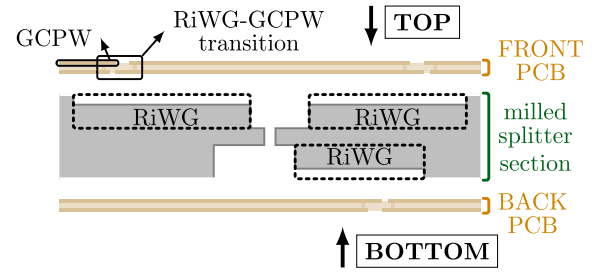
FIGURE 16. Top and bottom views of the front PCB and milled splitter section of a single power divider prototype. The back PCB is identical to the front PCB, but with only one port.

A. MANUFACTURING AND ASSEMBLY

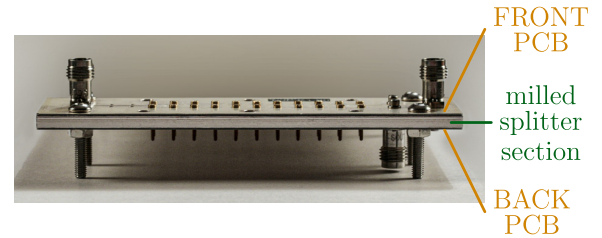
Multiple strategies exist to manufacture the presented power divider topology. In this work, a hybrid approach is adopted, which combines computer numerical control (CNC) milling and PCB technology, thereby offering a straightforward interface with the measurement setup. Furthermore, in a practical use case where the divider is part of the feed network of an antenna array (as in Fig. 1), this approach enables seamless integration of active front-end electronics, as well as radiating antenna elements.

Following the proposed hybrid scheme, the fabrication of a single power divider involves three separate parts: a milled splitter section, a front PCB, and a back PCB. The central aluminum part (EN AW-6082, $\sigma = 24 - 32$ MS/m) contains the entire coupling section, extended by 10 mm-long RiWG segments at each port. The exterior waveguide walls are not included in this part, which enables single-piece fabrication of the splitter section, as shown in Fig. 16.

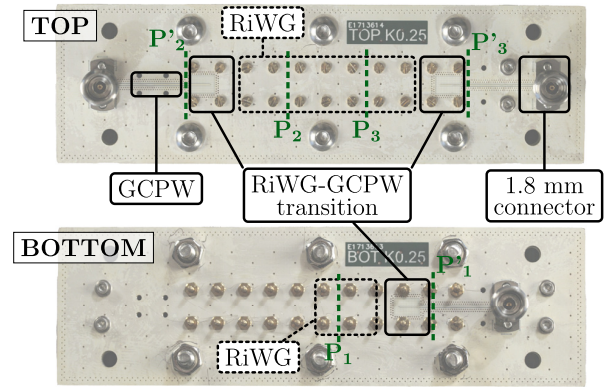
On either side of the milled splitter section, a single-layer PCB with a thickness of 324 μm is placed. This is illustrated schematically in Fig. 17a, which shows the vertical stack up of the front PCB, the milled splitter section and the back PCB. The PCB layers are implemented using a 254- μm -thick RO4350B high-frequency laminate substrate ($\epsilon_r = 3.66$ and $\tan \delta = 0.0037$), with 35- μm -thick copper layers on both sides, having a silver surface finish. These PCBs serve a dual purpose. On the one hand, they provide the external enclosures of the power divider and, on the other hand, they facilitate measurement on a Rohde & Schwarz (R&S) ZNA67 vector network analyzer (VNA). Interfaces between the RiWGs and 1.8 mm coaxial connectors are provided



(a) Schematic side view.



(b) Side-view picture of assembled prototype.



(c) Top- and bottom-view pictures of assembled prototype ($\kappa = 0.25$).

FIGURE 17. Power divider prototype assembly of the milled splitter section with a front and back printed-circuit-board (PCB): schematic side view and side-, top- and bottom-view pictures. Port planes P_1 , P_2 and P_3 indicate the divider design as such (Fig. 2a), while port planes P'_1 , P'_2 and P'_3 include the transition and a 10 mm-long RiWG segment.

through RiWG-to-grounded coplanar waveguide (GCPW) transitions, designed to cover the full n257 band, and via short GCPW sections.

The prototypes are assembled manually, by sandwiching the central aluminum layer between both PCBs and tightening the assembly with M1 metal screws. Figs. 17b and 17c show side-, top- and bottom-view pictures, respectively, of such an assembled prototype ($\kappa = 0.25$). Following the described procedure, five power divider designs were prototyped. Three *small- κ* regime prototypes realize split ratios of 0.05, 0.25 and 0.5, while two operate in the *large- κ* regime, with split ratios 0.9 and 1. Referring to Fig. 2, the prototype dimensions are listed in Table 2.

TABLE 2. Power divider prototype dimensions (mm). For all prototypes, the widths w_s , w and w_r equal 0.4 mm, 4.5 mm and 1.8 mm, respectively.

κ	0.05	0.25	0.5	0.9	1
s_a	0.65	0.9	1.35	3.35	3.7
s_b	0	0	0	0.1	0.35
s_c	2.55	2.55	2.55	0.725	0.35
y_b	2.25	0.6	0.15	0	0
y_s	3.5	2.15	1.15	0.25	0

B. MEASUREMENT RESULTS

The fabricated power dividers were characterized through S-parameter measurements, followed by calibration using a thru-reflect-line (TRL) kit. This procedure de-embeds the 1.8 mm connectors, as well as the GCPW lines leading up to the transitions, resulting in the S-parameters at port planes P'_1 , P'_2 and P'_3 (Fig. 17c). The corresponding split ratios κ of all prototypes are shown in Fig. 18. They are in close agreement with the target values of 0.05, 0.25, 0.5, 0.9 and 1, showing an average deviation of 0.48 dB, 0.34 dB, 0.23 dB, 0.10 dB and 0.08 dB, respectively. The -10 dB-impedance FBWs around 28 GHz vary between 51.3% and 56%, significantly exceeding the targeted 10% FBW for all prototypes, as illustrated in Fig. 19. For the other S-parameters, the prototype with $\kappa = 0.25$ is considered as a representative example. Fig. 20 compares its measured, simulated, and theoretical $|S_{31}|$, $|S_{21}|$, and $|S_{32}|$ parameters. According to (1), the corresponding theoretical values equal -0.97 dB, -6.99 dB, and -7.96 dB. The respective simulated values are on average 1.20 dB, 1.21 dB, and 1.10 dB lower. This difference largely originates from the GCPW-to-RiWG transitions, which each contribute approximately 0.44 dB insertion loss. The simulation also includes the finite conductivity σ and surface roughness of the PCB copper ($\sigma = 58$ MS/m, average roughness $R_a = 0.5$ μ m) to account for the silver surface finish [9] and the milled aluminum ($\sigma = 28$ MS/m, $R_a = 1$ μ m). The de-embedded measurement results align well with simulation, with the measured $|S_{31}|$, $|S_{21}|$, and $|S_{32}|$ on average 0.71 dB, 0.87 dB and 1.35 dB below simulation. This discrepancy can be attributed to the hybrid assembly method, which potentially gives rise to leakage through surface waves, as well as additional surface roughness and fabrication tolerances associated with both the milling and PCB processes. All other prototypes show similar agreement between measurement and simulation. For completeness, their $|S_{31}|$, $|S_{21}|$, and $|S_{32}|$ parameters are presented in Appendix B.

Fig. 21 presents the total transmission loss in dB, defined as $-10\log_{10}(|S_{11}|^2 + |S_{21}|^2 + |S_{31}|^2)$, measured between port planes P'_1 , P'_2 and P'_3 . Despite the fluctuations, which can partly be attributed to the non-ideality of the calibration procedure due to unavoidable fabrication differences between the TRL kit and the power divider prototypes, the measurements closely align with the simulated losses for all prototypes.

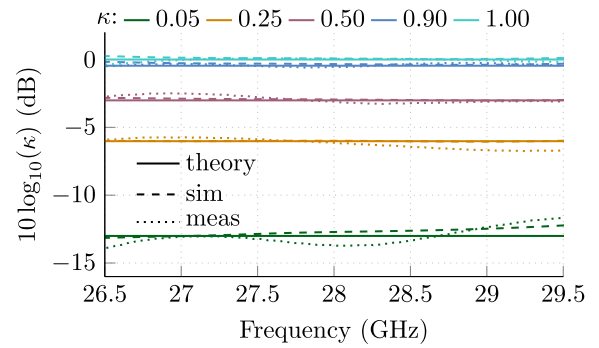


FIGURE 18. Theoretical, simulated and measured split ratios of the 28 GHz RiWG power divider prototypes, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

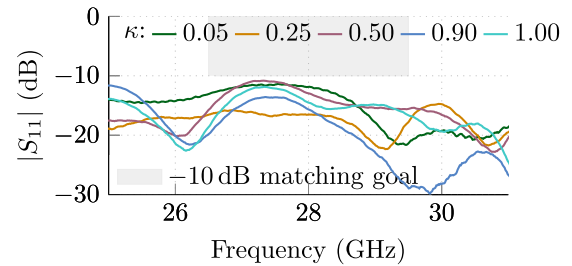


FIGURE 19. Measured reflection coefficients of the 28 GHz RiWG power divider prototypes, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

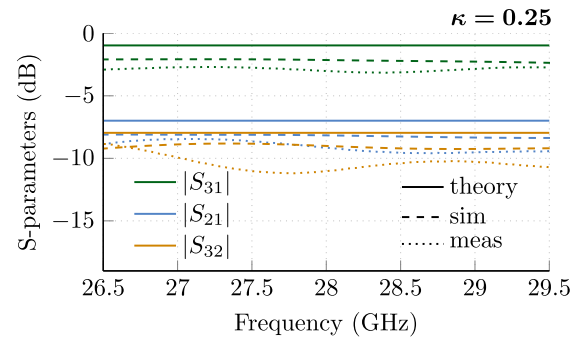


FIGURE 20. Theoretical, simulated and measured S-parameters of the 28 GHz RiWG power divider prototype with split ratio $\kappa = 0.25$, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

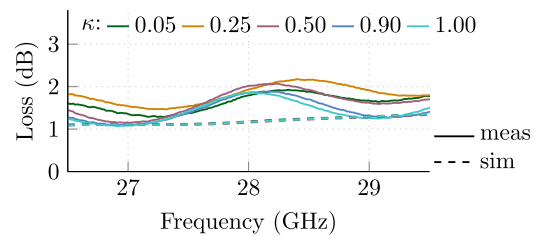


FIGURE 21. Measured and simulated transmission loss (dB) $-10\log_{10}(|S_{11}|^2 + |S_{21}|^2 + |S_{31}|^2)$ of the 28 GHz RiWG power divider prototypes, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

From these measurements, the proposed general design strategy is demonstrated to realize arbitrary split ratios while achieving -10 dB-impedance FBWs exceeding 50% around 28 GHz. For comparison with representative designs

reported in literature, equal- and unequal-split dividers will be discussed separately, beginning with the equal-split case. While several planar implementations report broader bandwidths [26], they generally operate at lower frequencies, relying on conventional lossy transmission-line technologies. The E-plane stripline divider in [32] also enables vertical integration, while offering an FBW exceeding 21.7% around 60 GHz. Rectangular waveguide solutions such as [46] achieve very low insertion loss (<0.13 dB) and provide additional isolation structures, but occupy larger footprints. More compact array feeding is possible with the RiWG design in [33], reporting a -10dB-impedance FBW of approximately 7% around 14.25 GHz.

While the aforementioned dividers focus exclusively on equal power division, arbitrary power division has been demonstrated using a planar H-plane waveguide T-junction [28], with an operating bandwidth of approximately 10% around 12.4 GHz. A vertically-integrated ridge gap waveguide topology was presented in [35] for array applications in the 23.5-24.5 GHz band, enabling both equal and unequal split ratios, with the reported V-polarized divider achieving a -10dB-impedance FBW of about 10%.

These results demonstrate that the design procedure proposed in this paper combines arbitrary power division with an FBW exceeding 50%, while remaining compatible with compact vertically integrated air-filled RiWG-based array-feeding networks. The design strategy yields a first functional prototype suitable for a wide variety of use cases, after which application-specific optimization of this divider design is possible. Beyond the set of four parameters manipulated in this paper, additional design variables may be incorporated into the procedure, to further enhance the operational bandwidth or implement more sophisticated phase-compensation mechanisms.

V. CONCLUSION

The proposed unequal-split ridge-waveguide power divider allows accurate control of split ratios between 0.05 and 1 via a systematic design methodology. The divider's vertical topology and constant waveguide cross-section facilitate modular deployment in an antenna array's feed distribution network, while ensuring compactness and low insertion losses. Moreover, the wide split-ratio range enables the implementation of diverse amplitude tapering schemes in these networks, allowing antenna designers to flexibly and significantly lower the array's side lobe levels. Although the design was demonstrated and experimentally validated for the 5G n257 frequency band, the proposed topology, together with the formulated design rules, enables scaling to any desired frequency of operation. Future research may explore the practical realization of such multi-stage divider networks, and their integration with radiating elements and active electronics. Different manufacturing methods may be applied, tailored to the specific application, and to a multitude of frequency ranges. Ultimately, the presented power divider topology shows the potential to support highly directive 1D

beam steering with suppressed side lobes, thereby effectively meeting the performance demands of future multi-antenna systems.

APPENDIX A

EQUIVALENT CIRCUIT MODEL DERIVATION

Starting from the equivalent circuit of the proposed power divider shown in Fig. 3, its S-parameters are first derived with respect to the same reference impedance Z_0 at all three ports. The general admittance $Y(s)$ is then realized as a first-order resonant parallel RLC-tank, expressed by $Y(s) = \frac{1}{R} + sC + \frac{1}{sL}$. Based on the desired power-division characteristics, the component values R , L and C are now determined.

Resonance at $\omega = \omega_r$ is achieved through the relation

$$L = \frac{1}{\omega_r^2 C}. \quad (10)$$

At resonance ($\omega = \omega_r$), the reflection coefficient equals $|S_{11,r}|$, and the RLC-circuit's admittance equals $Y_r = 1/R$. This results in the following expression for the resistance R , with Z_0 being the 50 Ω reference impedance:

$$\begin{aligned} |S_{11,r}| &= \frac{|Y_r Z_0|}{|2(\kappa + 1)n_3^2 + Y_r Z_0|} = \frac{Z_0}{2R(\kappa + 1)n_3^2 + Z_0} \\ \Rightarrow R &= \frac{\frac{1}{|S_{11,r}|} - 1}{2(1 + \kappa)n_3^2} Z_0. \end{aligned} \quad (11)$$

For $|S_{11,r}|$ lying 6 dB below $|S_{11,F}|$, this leads to (9a).

Lastly, the reflection coefficient at the frequencies ω_1 and ω_2 is fixed through the capacitance C . At $\omega = \omega_2 = \left(1 + \frac{\text{FBW}}{2}\right) \omega_r$, the RLC-circuit's admittance Y_F is given by

$$\begin{aligned} Y_F &= \frac{1}{R} + j \left(1 + \frac{\text{FBW}}{2} - \frac{1}{1 + \frac{\text{FBW}}{2}} \right) \omega_r C \\ &= \frac{1}{R} + j \alpha \omega_r C, \end{aligned} \quad (12)$$

while its reflection coefficient at ω_2 must equal $|S_{11,F}|$:

$$|S_{11,F}| = \frac{|Y_F Z_0|}{|2(\kappa + 1)n_3^2 + Y_F Z_0|}. \quad (13)$$

Substituting Y_F results into

$$\begin{aligned} |S_{11,F}| &= \frac{\sqrt{\frac{Z_0^2}{R^2} + (\alpha \omega_r C Z_0)^2}}{\sqrt{\left[2(\kappa + 1)n_3^2 + \frac{Z_0}{R} \right]^2 + (\alpha \omega_r C Z_0)^2}} \\ &= \frac{\sqrt{1 + (\alpha \omega_r R C)^2}}{\sqrt{\frac{1}{|S_{11,r}|^2} + (\alpha \omega_r R C)^2}}. \end{aligned} \quad (14)$$

Solving this equation for C yields

$$C = \frac{1}{\alpha \omega_r R} \sqrt{\frac{|S_{11,F}|^2 - 1}{1 - |S_{11,r}|^2}}. \quad (15)$$

For a 10% FBW, α equals 41 / 420, and with $|S_{11,r}|$ 6 dB below $|S_{11,F}|$, (15) results into (9c).

APPENDIX B

FULL S-PARAMETER DATA

Figs. 22–25 show the measured, simulated and theoretical $|S_{31}|$, $|S_{21}|$ and $|S_{32}|$ data for the prototypes with $\kappa = 0.05, 0.50, 0.90$ and 1.00 , complementing the $\kappa = 0.25$ results presented in Fig. 20. The solid lines indicate the frequency-independent theoretical values of the lossless power divider, calculated from (1). The dashed lines show the simulated performance of the proposed RiWG-based power dividers, including the GCPW-to-RiWG transitions and accounting for finite metal conductivity and surface roughness. This explains the gap between simulation and theory. The dotted lines represent the measurement results of the assembled prototypes, de-embedded up to these transitions. Overall, good agreement is observed between measurements and simulations for all split ratios. Remaining discrepancies can be attributed to fabrication tolerances, the hybrid assembly process, and deviations in surface roughness from the values assumed in simulation.

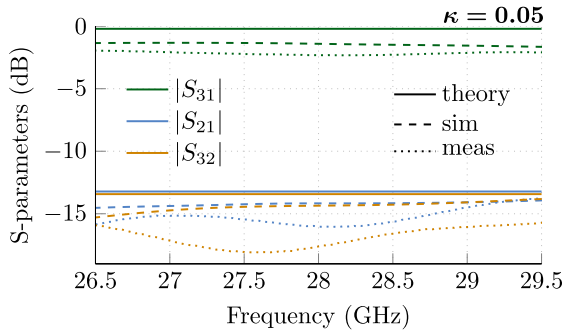


FIGURE 22. Theoretical, simulated and measured $|S_{31}|$, $|S_{21}|$ and $|S_{32}|$ of the 28 GHz RiWG power divider prototype with split ratio $\kappa = 0.05$, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

Fig. 22 presents the S-parameters for the RiWG power divider with $\kappa = 0.05$. The theoretical values equal -0.21 dB for $|S_{31}|$, -13.22 dB for $|S_{21}|$, and -13.43 dB for $|S_{32}|$. On average, the simulated values are 1.22 dB, 0.99 dB and 0.97 dB lower, respectively. The corresponding measurement results are on average 0.74 dB, 1.04 dB and 2.50 dB below the simulated values.

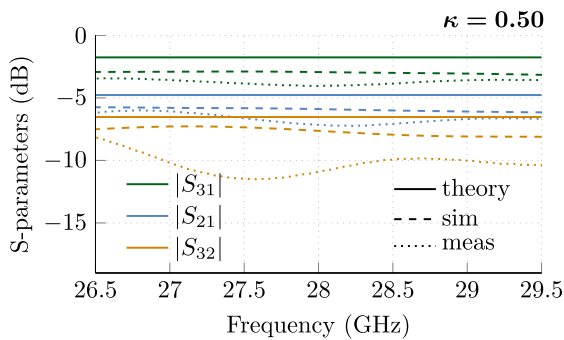


FIGURE 23. Theoretical, simulated and measured $|S_{31}|$, $|S_{21}|$ and $|S_{32}|$ of the 28 GHz RiWG power divider prototype with split ratio $\kappa = 0.50$, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

Fig. 23 presents the S-parameters for the divider design with $\kappa = 0.50$. The theoretical values equal -1.76 dB for $|S_{31}|$, -4.77 dB for $|S_{21}|$, and -6.53 dB for $|S_{32}|$. On average, the simulated values are 1.21 dB, 1.15 dB and 1.15 dB lower, respectively. The corresponding measurement results are on average 0.76 dB, 0.74 dB and 2.64 dB below the simulated values.

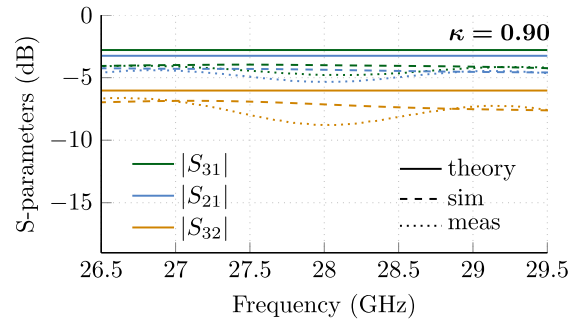


FIGURE 24. Theoretical, simulated and measured $|S_{31}|$, $|S_{21}|$ and $|S_{32}|$ of the 28 GHz RiWG power divider prototype with split ratio $\kappa = 0.90$, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

Fig. 24 presents the S-parameters for the divider design with $\kappa = 0.90$. The theoretical values equal -2.79 dB for $|S_{31}|$, -3.25 dB for $|S_{21}|$, and -6.03 dB for $|S_{32}|$. On average, the simulated values are 1.26 dB, 1.15 dB and 1.15 dB lower, respectively. The corresponding measurement results are on average 0.34 dB, 0.39 dB and 0.52 dB below the simulated values.

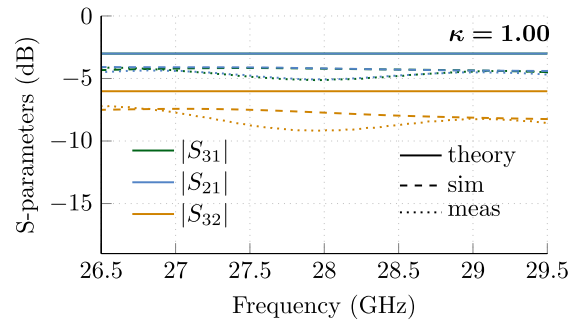


FIGURE 25. Theoretical, simulated and measured $|S_{31}|$, $|S_{21}|$ and $|S_{32}|$ of the 28 GHz RiWG power divider prototype with split ratio $\kappa = 1.0$, after TRL calibration up to port planes P'_{1-3} (Fig. 17).

Finally, Fig. 25 presents the S-parameters for the divider design with $\kappa = 1.0$. The theoretical values equal -3.01 dB for $|S_{31}|$ and $|S_{21}|$, and -6.02 dB for $|S_{32}|$. On average, the simulated values are 1.26 dB, 1.21 dB and 1.75 dB lower, respectively. The corresponding measurement results are on average 0.38 dB, 0.46 dB and 0.64 dB below the simulated values.

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SOFIE LENDERS (Graduate Student Member, IEEE) received the master's degree in electrical engineering from Ghent University, Belgium, in 2024. She is currently pursuing the Ph.D. degree in electrical engineering with IDLab, Electromagnetics Group, Department of Information Technology (INTEC), Ghent University/imec. Her research interests include millimeter-wave multi-antenna systems, highly-efficient passive waveguide components, and numerical electromagnetics.



SAMUEL RIMBAUT (Graduate Student Member, IEEE) received the M.Sc. degree in electrical engineering from Ghent University, Ghent, Belgium, in 2021. He is currently pursuing the Ph.D. degree in electrical engineering with IDLab, Electromagnetics Group, Department of Information Technology (INTEC), Ghent University/imec. His research interests include next-generation communication systems, more specifically highly efficient millimeter-wave and THz passive and active solutions.



BRAM HOF Lack (Graduate Student Member, IEEE) received the M.Sc. degree in electrical engineering from Ghent University, Ghent, Belgium, in 2022. He is currently pursuing the Ph.D. degree in electrical engineering with IDLab, Electromagnetics Group, Department of Information Technology (INTEC), Ghent University/imec. His research interests include next-generation communication systems, more specifically highly efficient PCB-based millimeter-wave passive and active solutions.



SAM LEMEY (Member, IEEE) received the M.Sc. degree in electronic engineering from Howest, University College West Flanders, Kortrijk, Belgium, in 2012, and the Ph.D. degree in electrical engineering from Ghent University, Ghent, Belgium, in 2016. From January 2018 to March 2018, he was a Visiting Scientist with the Terahertz Photonics Group, Institute of Electronics, Microelectronics and Nanotechnology (IEMN), University of Lille Nord de France, Lille, France.

He is currently a Professor with the Department of Information Technology (INTEC), Ghent University/imec. His research interests include antenna systems for wearable applications, active antenna design for the Internet of Things and (beyond) 5G applications, (optoelectronic) millimeter-wave multi-antenna systems, impulse-radio ultra-wideband antenna systems for centimeter precision localization, and full-wave/circuit co-optimization frameworks to realize (optoelectronic) active (multi-)antenna systems. He received the URSI Young Scientist Award at the 2020 URSI General Assembly and the Best Paper Award at the 2016 IEEE MTT-S Topical Conference on Wireless Sensors and Sensor Networks. He was also a co-recipient of the 2015 Best Paper Award at the 22nd IEEE Symposium on Communications and Vehicular Technology in Benelux and the 2019 ECOC Best Demo Award.



HENDRIK ROGIER (Senior Member, IEEE) received the M.Sc. and Ph.D. degrees in electrical engineering from Ghent University, Ghent, Belgium, in 1994 and 1999, respectively.

From 2003 to 2004, he was a Visiting Scientist with the Mobile Communications Group, Vienna University of Technology, Vienna, Austria. He is currently a Senior Full Professor with the Department of Information Technology, Ghent University, and a Guest Professor with the Interuniversity Microelectronics Centre, Ghent. He has authored or co-authored over 215 articles in international journals and over 250 contributions in conference proceedings. His current research interests include antenna systems, radio wave propagation, body-centric communication, numerical electromagnetics, electromagnetic compatibility, and power/signal integrity. He is a member of MTT TC-26 RFID, Wireless Sensor and IoT Committee, and AP-S TC-11 Health and Medicine, and he acts as the URSI Commission B representative for Belgium. From 2017 to 2019, he was an Associate Editor of IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES. He is an Associate Editor of *IET Electronics Letters* and *IET Microwaves, Antennas and Propagation*.

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