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**Innovative High Power Combiners Design, Fabrication, and
Characterization for K/Ka Band systems**

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1 Introduction

As the operating frequency of semiconductor solid-state devices increases into the millimeter wave region, the size of the devices and their power handling capability are reduced. In order to provide the advantages of a solid-state technology for moderate power levels at millimeter-wave frequencies, solid-state components must be combined. Spatial power combining provides enhanced RF efficiency by coupling the components to beams or modes in free space rather than via transmission lines in corporate combining structures.

Waveguide spatial power combiners are extensively employed in various communication systems at both microwave and millimeter-wave frequencies. They play an important role in enabling solid-state power amplifiers to compete with microwave tube amplifiers (travelling wave tube - TWT), offering several advantages over other combiner types, such as planar divider/ combiners, especially for configurations involving numerous ports ([1] – [18]). The key benefits include lower losses and a more compact size, achieved by minimizing the path lengths. In recent years, there has been significant interest in new types of spatial power combiners.

The ongoing pursuit of higher efficiency, particularly in microwave and millimeter wave devices, has driven researchers to explore innovative structures and methods for handling power division and recombination.

Rising frequencies make heat management and achieving acceptable efficiency levels the real challenges that modern electronics must tackle. These are critical challenges, especially given that the key sectors implementing high-frequency devices range from space applications to civilian and military uses.

Today, several technologies can deliver high power values in the microwave and millimeter wave fields ([19] – [40]). The highest power levels are achieved by vacuum devices, also known as electron tubes, such as Traveling Wave Tube Amplifiers (TWTAs), Klystrons, etc. These devices are currently pushed to their power generation limits, primarily operating in pulsed power applications while also generating powers in the gigawatt range.

However, implementing this technology comes with significant drawbacks. Nearly all vacuum devices require extended warm-up times and must be powered by high voltages that are often impossible to supply in space-constrained environments. This happens because these devices rely on heated filaments, electromagnets, or bulky magnets, which also result in large dimensions, heavy weight, and poor efficiency. This is why there's an ongoing shift from vacuum devices to power devices capable of delivering at least comparable power levels to electron tubes.

Current solid-state power amplifiers (SSPAs) can provide hundreds of watts in the microwave range and several tens of watts in the lower millimeter wave range. However, to match the power levels that vacuum devices can deliver in the same frequency range, it's necessary to combine the output power of multiple solid-state devices.

Traditional planar power combining methods don't represent the most efficient technique for achieving this goal. In fact, they're often impractical even for relatively small numbers of combined power devices—more than four, for instance. This is mainly due to the inefficiency of planar architectures, which require multiple power division stages combined with high substrate losses in these planar circuits.

The state-of-the-art for microwave and millimeter wave power combining is represented by Wavefront Amplifiers (WFAs), also called Wavefront Combiners (WFCs). In literature, these devices are known as Spatial Power Amplifiers (SPAs) or Spatial Power Combiners (SPCs), though this naming can be misleading when trying to understand how these structures work.

A wavefront combiner can combine large numbers of solid-state power devices as efficiently as possible with minimal insertion losses. This is made possible by implementing electromagnetic propagation media with very low losses, such as waveguides, and also thanks to the ability to perform multiple power divisions at the same spatial point.

Several families of wavefront combiners can be found in literature and across various application fields for these devices. However, the search for innovative structures is driven by the need to push these structures' power density to the limit—expressed both in terms of weight-to-power ratio and weight-to-volume ratio.

The innovative structures presented in this thesis work aim precisely to maximize Insertion loss and efficiency values. To achieve this, different techniques and power division tools have been analyzed. Through implementing these new solutions, it will be possible to create more compact and efficient wavefront combiners.

1.1 Theoretical Fundamentals

The scattering matrix is the matrix that best represent the physical behaviour of a microwave circuit. It is important to highlight that voltages and currents previously introduced are fictitious quantities that cannot be measuread directly. This is true, instead, for amplitudes and phases of the waves travelling in a generic transmission line connecting the circuit ports. Before introducing the scattering parameters, it is necessary to introduce the following quantities, defined as incoming and outgoing waves at the reference plane Ti :

$$a_i = \frac{V_i + Z_{0i}I_i}{2\sqrt{Z_{0i}}} = \frac{V_i^+}{2\sqrt{Z_{0i}}}$$

$$b_i = \frac{V_i - Z_{0i}I_i}{2\sqrt{Z_{0i}}} = \frac{V_i^-}{2\sqrt{Z_{0i}}}$$

where Z_{0i} is the characteristic impedance along the i th feeding line; (V_{+i}, I_{+i}) and (V_{-i}, I_{-i}) are voltage and current waves travelling towards and away from the circuit. With that definitions, voltage and current can be expressed as follows:

$$V_i = (a_i + b_i)\sqrt{Z_{0i}}$$

$$I_i = (a_i - b_i)/\sqrt{Z_{0i}}$$

Thus, a_i and b_i represent the incoming (+) and reflected (-) voltage waves along the i th line at the reference plane, apart from the factor $1/2\sqrt{Z_{0i}}$. By adopting this notation is possible to express the power entering the i th port as:

$$P_i = \frac{1}{2}V_i I_i^* = |a_i|^2 - |b_i|^2 = P_i^+ - P_i^-$$

where

$$P_i^+ = \frac{1}{2}V_i^+ I_i^{+*} = |a_i|^2$$

$$P_i^- = \frac{1}{2}V_i^- I_i^{-*} = |b_i|^2$$

The scattering matrix shows the outgoing waves b_i as linear functions of the incoming waves a_i :

$$\begin{aligned}
b_1 &= s_{11}a_1 + s_{12}a_2 + \cdots + s_{1n}a_n \\
b_2 &= s_{21}a_1 + s_{22}a_2 + \cdots + s_{2n}a_n \\
&\vdots \\
b_n &= s_{n1}a_1 + s_{n2}a_2 + \cdots + s_{nn}a_n
\end{aligned}$$

or, as seen with the other two representations, in matrix from:

$$[b] = [S][a]$$

The elements of the matrix, known as scattering parameters or simply S parameters, can be defined only when a set of impedance Z_{0i} of the feeding lines has been specified. Indeed, changing the reference impedance causes a variation in the values of the scattering parameters. The S-Parameter S_{ij} is the ratio of the wave amplitude b_i leaving from port i to the wave amplitude a_j incident on port j when no other wave is incident on the remaining ports:

$$s_{ij} = \left. \frac{b_i}{a_j} \right|_{a_1=a_2=\cdots=a_N=0, a_j \neq 0}$$

It is implicitly assumed that all ports are terminated with their respective characteristic impedances Z_{0i} of the transmission lines. In other words, the lines are matched so that no wave is reflected back from the load toward the network. By setting $j = i$, the parameter s_{ii} represents the reflection coefficient at the i th port when all other ports are matched. S-parameters are related to the power travelling through the network, as following:

$$|s_{ij}|^2 = P_i^- / P_j^+$$

1.1.1 High-frequency transmission Lines: Rectangular waveguide and microstrip lines

The fundamental components of any RF circuit or system are transmission lines, which are essential for connecting devices and transmitting RF signals. At RF frequencies, passive components behave as distributed elements composed of sections of transmission lines, whose physical dimensions are typically of the same order of magnitude as the wavelength. A key factor in selecting a particular type of waveguide is related to the losses in the propagating structures, which generally increase with frequency due to conduction losses and spurious radiation from discontinuities and interconnections. Therefore, the choice of transmission line strongly depends on the operating frequency. Figure 1 summarizes the most commonly used guiding structures and transmission lines. Structures such as coaxial cables or striplines, which employ a multi-conductor configuration, support TEM mode propagation starting from DC. Conversely, waveguides characterized by a single conductor structure exhibit a cut-off frequency, below which electromagnetic propagation cannot occur.

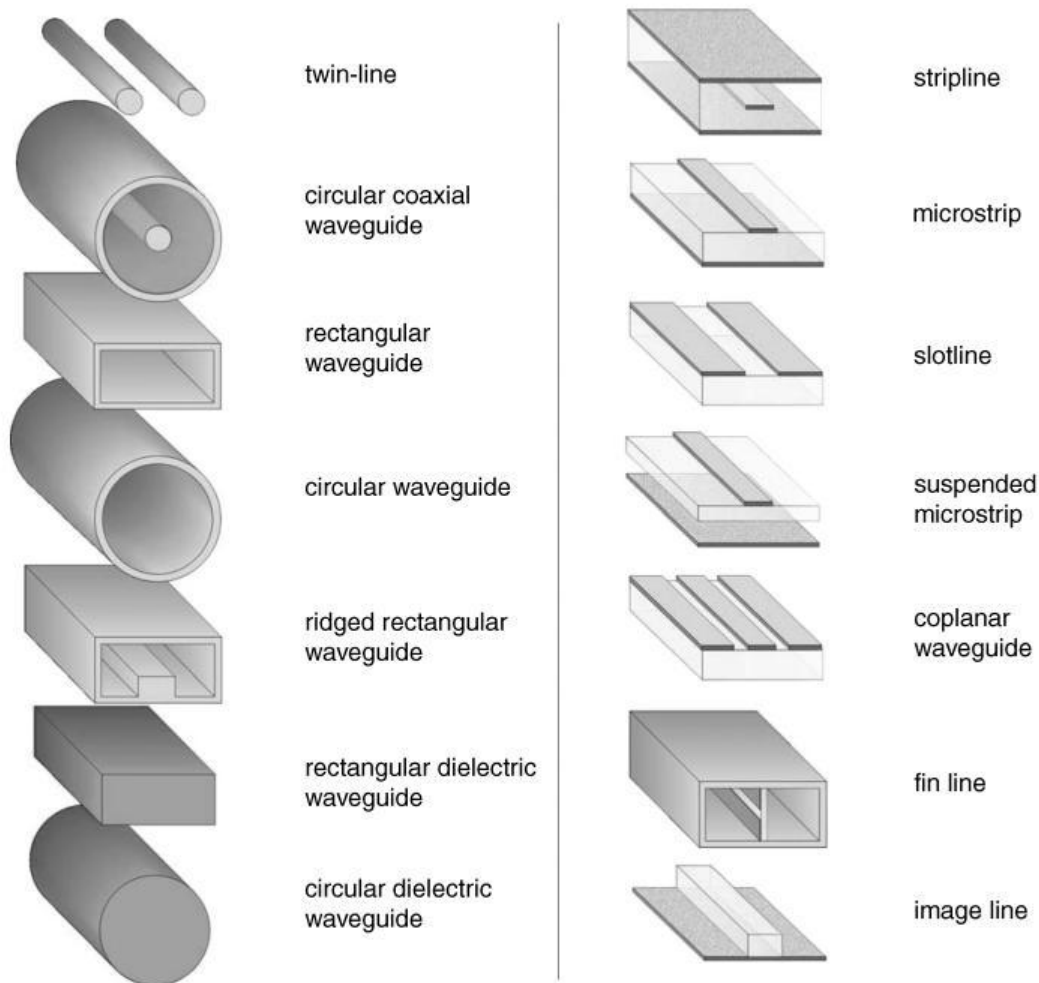


Figure 1 - Some common transmission lines for RF and microwaves.

Planar circuits, except for the stripline, are characterized by the inhomogeneity between the dielectric substrate and the air over it, thus exhibiting a dispersive behaviour. The characteristics of the propagation are appreciably different depending on the guiding structure. However, it is useful and possible to employ for all of them a common mathematical model based on waves of voltage and current propagating along the generalized transmission line. These voltages and currents are actual physical quantities only for the case of multi-conductor lines, while they just represent electric and magnetic field respectively. The following

Table lists the comparison of the performance of the mostly used transmissions lines and waveguides.

Characteristic	Coax	Waveguide	Stripline	Microstrip
Modes: Preferred	TEM	TE ₁₀	TEM	Quasi-TEM
Other	TM,TE	TM,TE	TM,TE	Hybrid TM,TE
Dispersion	None	Medium	None	Low
Bandwidth	High	Low	High	High
Loss	Medium	Low	High	High
Power capacity	Medium	High	Low	Low
Physical size	Large	Large	Medium	Small
Ease of fabrication	Medium	Medium	Easy	Easy
Integration with	Hard	Hard	Fair	Easy

Table 1 - Comparison of Common Transmission Lines and Waveguides.

1.1.2 Rectangular waveguide

The rectangular waveguide is one of the most common types of waveguides and can operate over a wide frequency range, typically from 320 MHz to 333 GHz.

Its geometry consists of a hollow metallic conductor with a rectangular cross-section, where a and b denote the dimensions of the longer and shorter sides, respectively. While the trend toward miniaturization and integration has led most modern microwave circuits to be implemented using planar transmission lines, rectangular waveguides remain essential in applications requiring high power handling, operation at millimeter-wave frequencies, satellite communications, and high-precision measurements. This type of waveguide supports only transverse electric (TE) and transverse magnetic (TM) modes, but

not the transverse electromagnetic (TEM) mode, as previously discussed. Each allowed mode has a specific cut-off frequency below which propagation cannot occur. The *cut-off* associated with the considered waveguide can be derived from the following:

$$f_c = \frac{1}{2\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}$$

At a given operating frequency f only those modes that have $f > f_c$ will propagate, while modes with $f < f_c$ will cause exponential decay of the field components away from the excitation source. Such modes are called cut-off modes or evanescent modes, as shown in Figure 2.

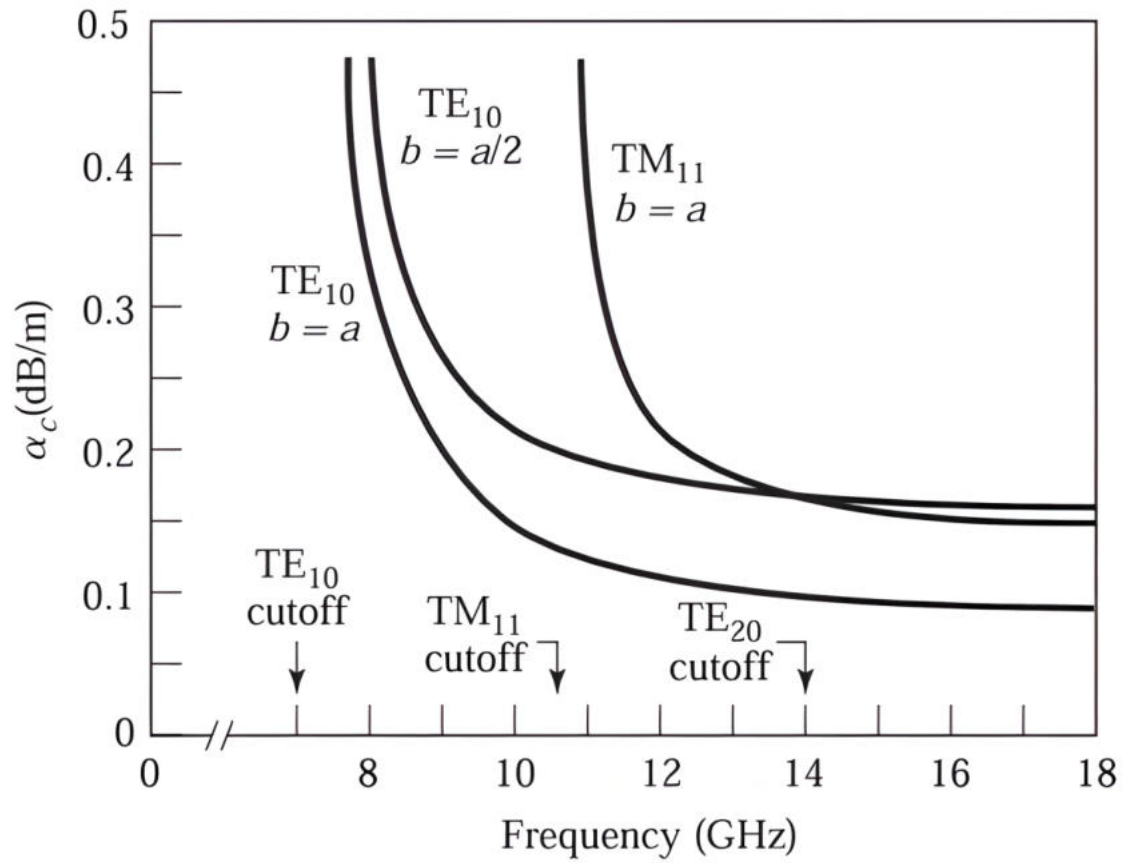


Figure 2 - Attenuation of various modes in a rectangular brass (alloy made of Copper and Zinc) waveguide with $a=2.0$

Figures 3 and 4 show both the structure of the rectangular waveguide and the EM field and electric current lines of the fundamental TE_{10} mode.

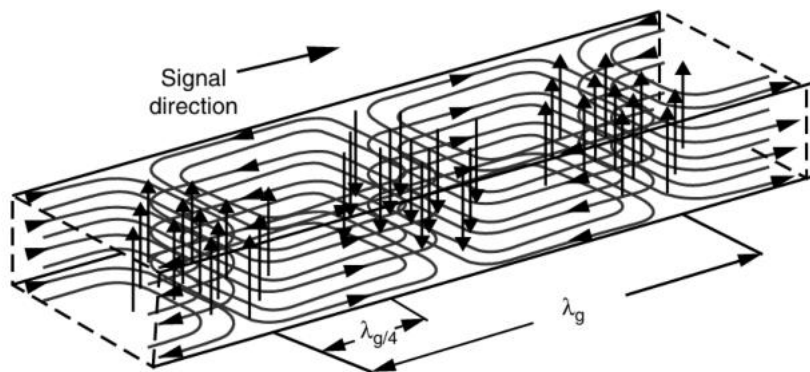


Figure 3 - Electric field lines and current density lines of the TE_{10} mode

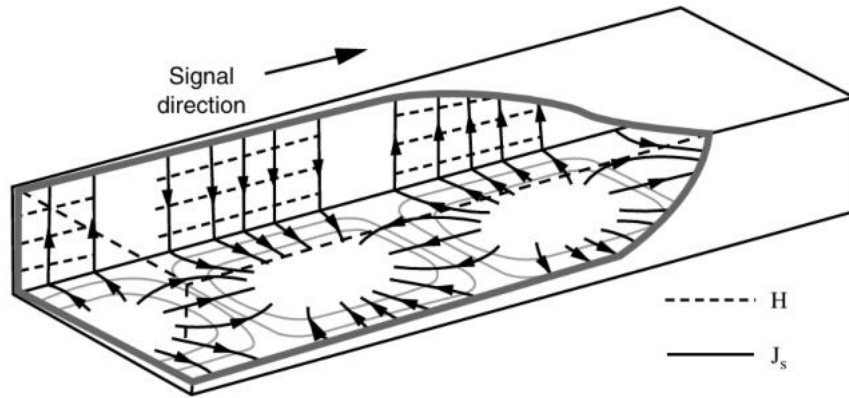


Figure 4 - Magnetic field lines and current density lines of the TE_{10} mode

1.1.3 Microstrip Lines

The microstrip line, which is illustrated in Figure 5, is the most common transmission line used in RF and microwave engineering, thanks to its large number of advantages like simplicity of manufacturing, low cost, low weight and the excellent integrability with active devices. It consists of a planar transmission line of width W printed on top of a dielectric substrate of thickness h , metalized on the opposite side. The EM field propagates in an inhomogeneous medium consisting of the dielectric substrate (ϵ_r) and the air (ϵ_0) on top of the microstrip, with two different propagation constants associated with them. This configuration prevents propagation of the TEM mode despite the presence of two different conductors.

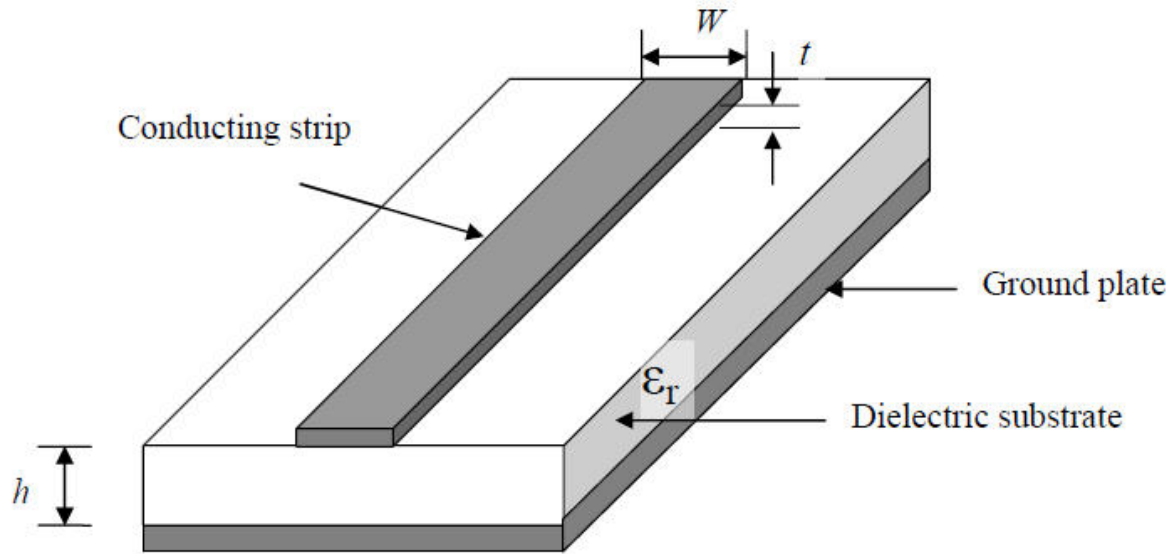


Figure 5 - Microstrip line schematic model

The phase velocity of TEM fields in the dielectric region would be $\frac{c}{\sqrt{\epsilon_r}}$, while in the air region would be c thus causing the impossibility to reach a phase matching condition at the dielectric-air interface. However, in most application the dielectric substrate is electrically very thin ($d \ll \lambda$) so the fundamental mode propagating in the structure is the quasi-TEM mode, essentially the same as those of the static case (DC).

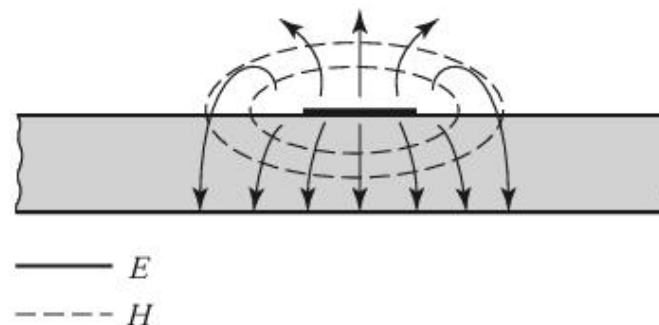


Figure 6 - Electric and magnetic field lines of the microstrip

1.2 Dividing and combining Techniques

A significant number of dividing-combining techniques have been developed over the years, driven by technological advances and by the versatility of these structures in addressing specific application requirements. The following figure provides an overview of the most widely used splitting-combining techniques:

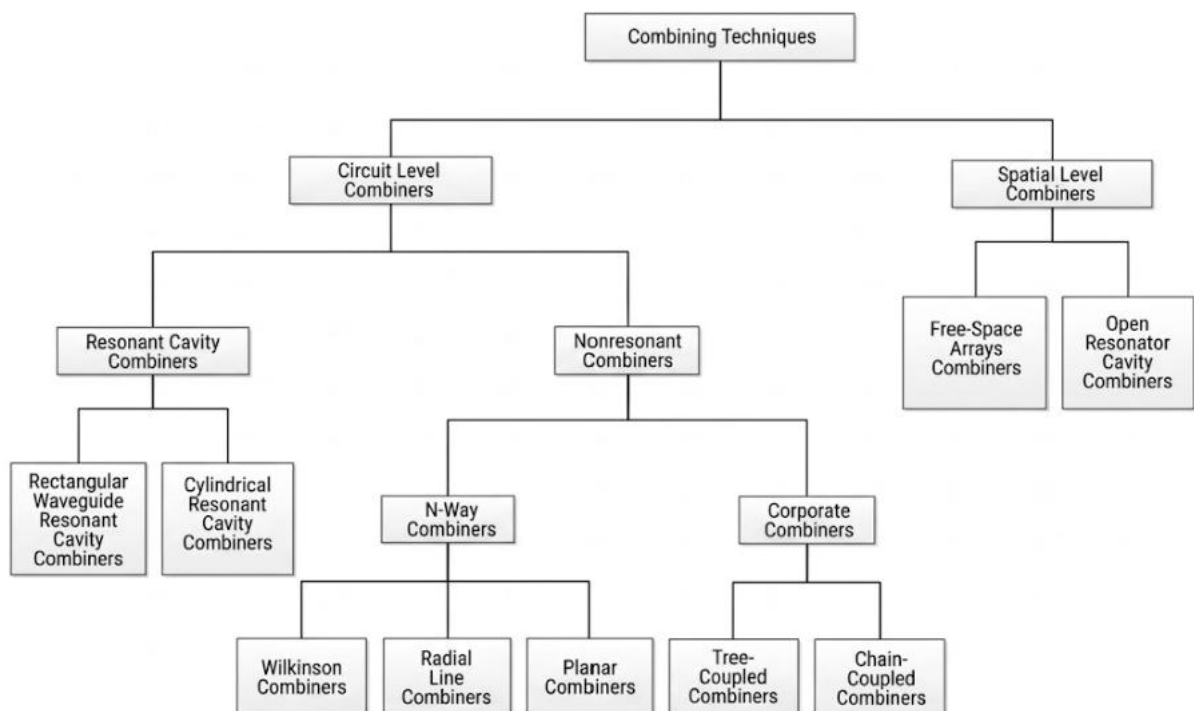


Figure 7 - Different power combining techniques

The following block diagram illustrates the splitting/combining network incorporating the SSA. The overall system is composed of multiple SSA stages operating in parallel. The input power is evenly distributed across the n branches of the divider, and each module independently amplifies the

incoming signal. The resulting output power is then collected through a combining network, with the aim of reaching the target power level.

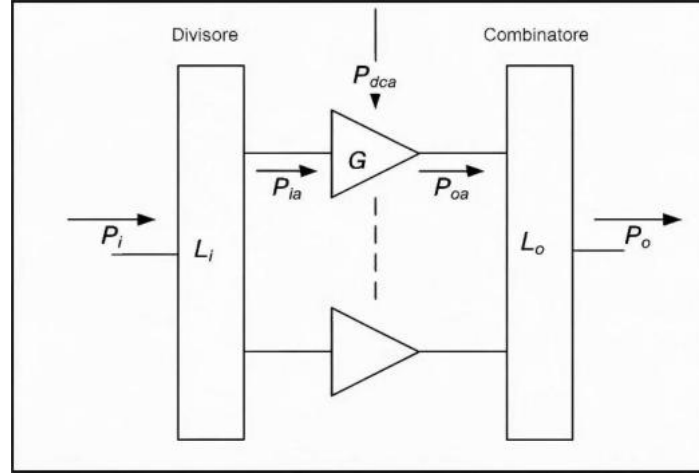


Figure 8 - Combining network schematic

In general, splitting and combining networks are designed to have identical characteristics, so that the signal behaviour in terms of amplitude and phase remains consistent across all paths [41], [42], [43], [44], [45], [46], [47], [48]. Several classes of devices can be used to split or combine RF/microwave power, and they are classified according to the number of output ports. The insertion losses introduced by both the splitting and combining stages affect the achievable power-added efficiency. The combiner efficiency can be expressed as follows:

$$\eta_c = \frac{N * P_{oa}}{P_o} = L_o$$

where N represents the number of combining paths. From the formula, it is straightforward to understand how the overall combining efficiency decreases as the number of paths increases.

The product $N \cdot P_{oa}$ takes into account the output power of each path and, consequently, the losses occurring in each line.

1.2.1 Performance parameters for power combiners

Performance of power combiners is characterized by several key parameters, which determine their suitability for specific applications. These parameters include efficiency, return loss, insertion loss, isolation, power handling, and bandwidth.

Efficiency is a critical parameter that indicates how much of the input power is delivered to the output. In an ideal lossless power combiner, the efficiency would be 100%. However, in practical devices, some power is always lost due to resistive losses in the conductors and dielectric losses in the materials. The efficiency (η) of a power combiner can be expressed as:

$$\eta = \frac{P_{out}}{\sum_{i=1}^N P_{in_i}} \times 100\%$$

where P_{out} is the output power and P_{in_i} is the input power from the i -th source, and N is the number of input ports. Higher efficiency is generally desirable to minimize power leakage and reduce heat generation.

Return Loss is a measure of how much power is reflected from the input ports of the power combiner. A high return loss indicates that most of the input power is accepted by the combiner, while a low return loss suggests that a significant portion of the power is reflected back to the source. Return loss is usually expressed in dB and is given by:

$$\text{Return Loss (dB)} = -20 \log_{10} |\Gamma_{in}|$$

where Γ_{in} is the reflection coefficient at the input port. A higher return loss (i.e., a more negative dB value) is generally preferred, as it minimizes signal reflections and improves the overall system performance.

Insertion Loss refers to the power loss that occurs when a signal passes through the power combiner. It is defined as the ratio of the output power to the total input power and is also expressed in dB:

$$\text{Insertion Loss (dB)} = -10 \log_{10} \frac{P_{out}}{\sum_{i=1}^N P_{in_i}}$$

Ideally, a power combiner should have minimal insertion loss, meaning that it should deliver as much of the input power as possible to the output. In practice, some insertion loss is unavoidable.

Isolation is the measure of power leakage between the input ports of the power combiner. High isolation is important to prevent interferences between signals from two, or more, different input ports. Isolation is also expressed in dB and a higher isolation value is desirable.

Power Handling Capacity indicates the maximum power level that the power combiner can handle without degradation in performance or structural damage. This is a crucial parameter, especially in high-power applications.

Bandwidth defines the range of frequencies over which the power combiner can operate effectively while maintaining the specified performance parameters. In summary, the selection of a power combiner depends on the specific requirements of the application, and these parameters must be carefully considered to ensure optimal performance of the overall system.

Two types of combining networks have been investigated:

- Spatial combining network

- Planar combining network

With respect to the first one, the existing literature offers several state-of-the-art configurations. Among these, the configurations selected for further consideration are the following:

1.2.2 Cylindrical cavity to microstrip coupling

A resonant cavity is a volume enclosed by perfectly conducting walls within which a field configuration is established. Several resonant cavity configurations exist; in general, the most used are those with cylindrical and rectangular geometry. As an example, Figure 9 shows the field distribution in a cylindrical cavity oscillating in the TM_{110} mode.

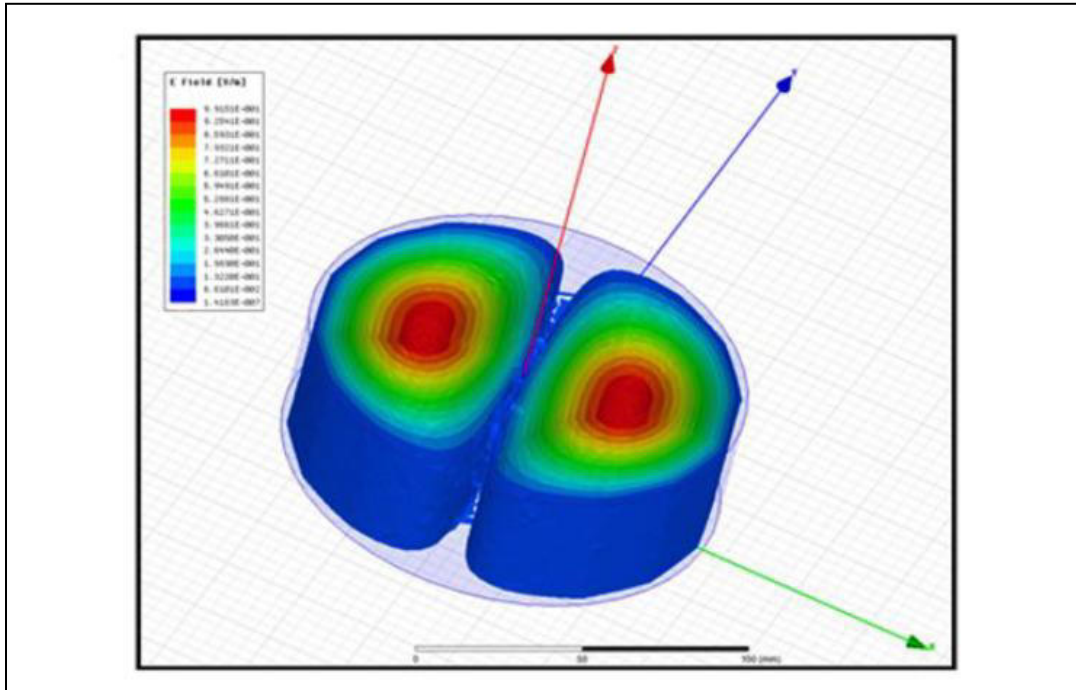


Figure 9 - TM_{110} cavity mode, shown in a cylindrical cavity; Font: A W-Band Spatial Power-Combining Amplifier using GaN MMICs

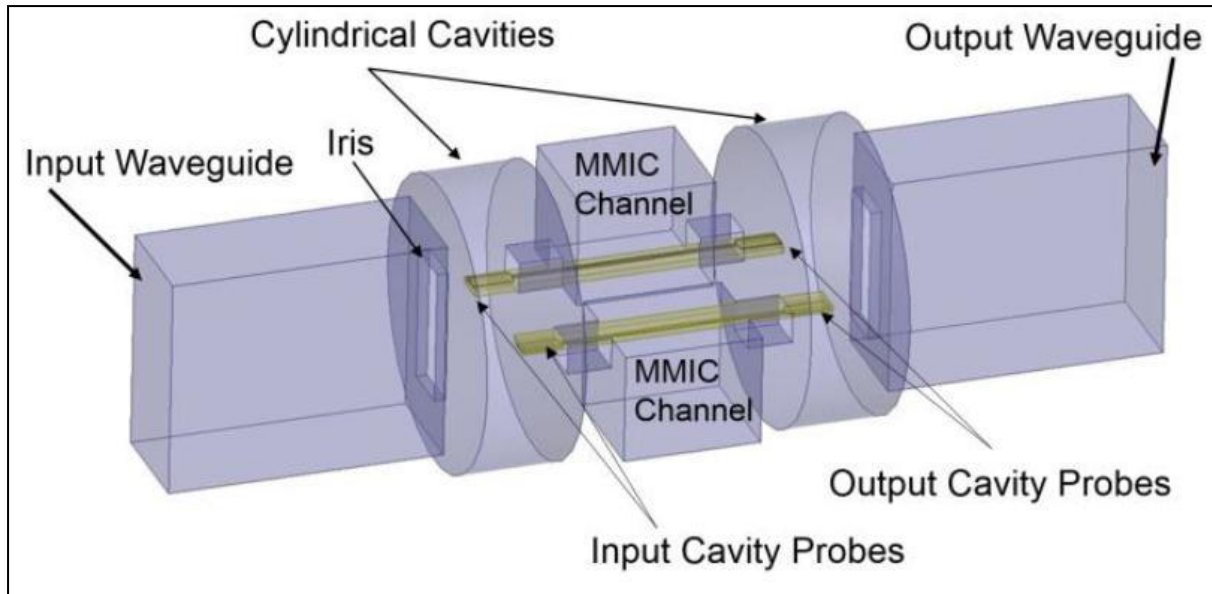


Figure 10 - Spatial power combiner using cavity modes; Font: A W-Band Spatial Power-Combining Amplifier using GaN MMICs

Regarding this type of coupling, it can be stated that the microstrip would enter the cavity exactly at the location where the electric field is at its maximum. In a cylindrical cavity, the mode considered is the TM_{110} . In this configuration, at most 2 microstrips can be combined, thus allowing for two amplifier chains operating in parallel.

1.2.3 Square cavity to microstrip coupling

Another example of SSA-cavity coupling is represented by a square geometry cavity. Figure 11 shows the block diagram of a splitting and combining network with the SSA placed at the center of the structure.

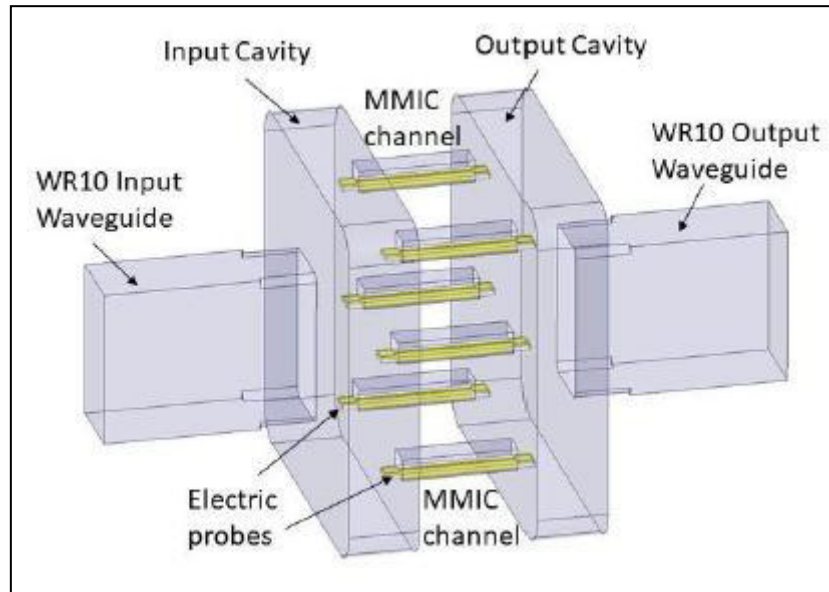


Figure 11 - Illustration of WSPCA spatial power combining amplifier using higher order mode cavities, with microstrip thru-lines indicated in the MMIC channels.

Furthermore, the electric field distribution of the TM_{320} mode in a square waveguide can also be observed.

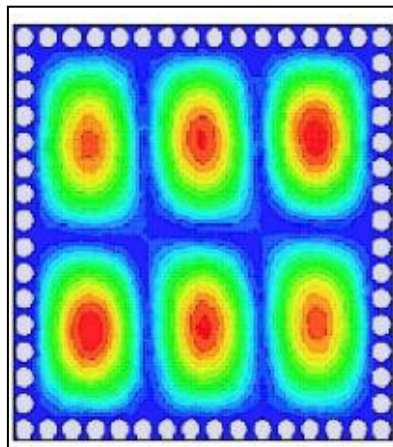


Figure 12 - E field distribution mode. TM_{320} mode

Regarding this type of coupling, it can be stated that the microstrip would enter the cavity exactly at the location where the electric field is at its maximum. In a square cavity, the mode considered is the TM_{320} . This allows for a maximum of 6 microstrips to be combined, thus enabling 6 amplifier chains operating in parallel.

1.2.4 Microstrip to Finline coupling (Configuration with Antenna)

Antennas are used to achieve impedance matching within a certain bandwidth between the transmission line carrying the signal and free space. Furthermore, they ensure that the radiated energy is concentrated within a predefined region of space, thereby conferring directivity to the radiated power. In practice, the N microstrips are allocated within the conical coaxial structure, where the power is collected and made available at the output. The antenna employed in this configuration must exhibit high directivity. The proposed antenna designs to be developed are the following:

- Vivaldi antenna
- Antipodal antenna
- Biconical antenna

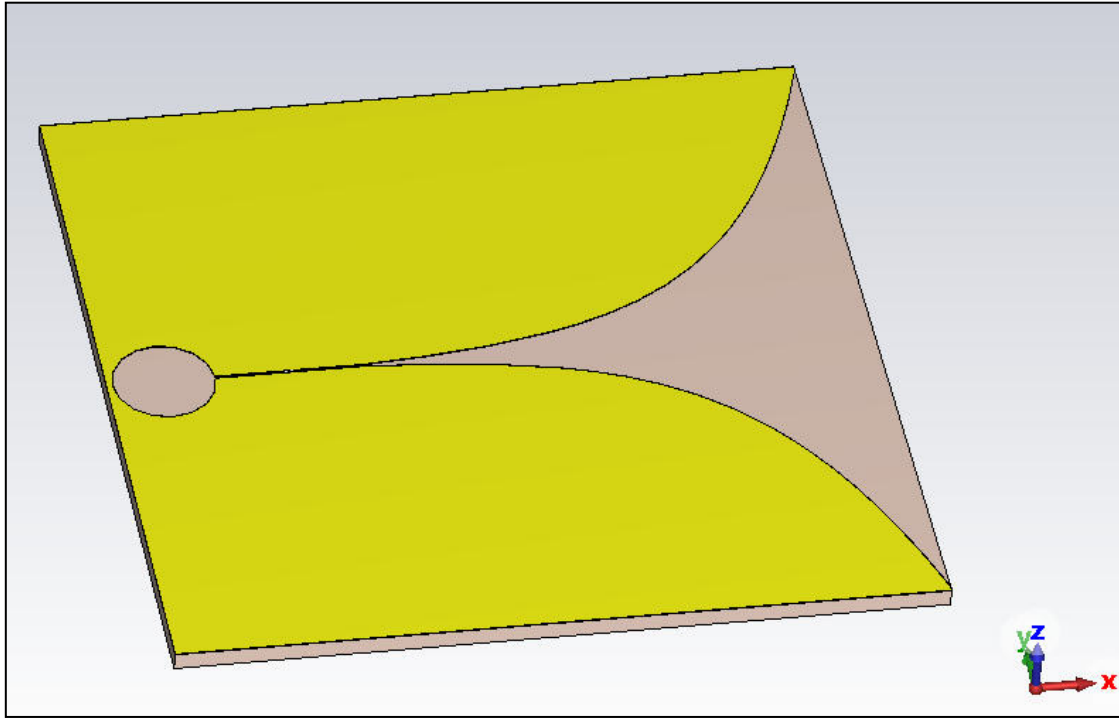


Figure 13 – Vivaldi Antenna

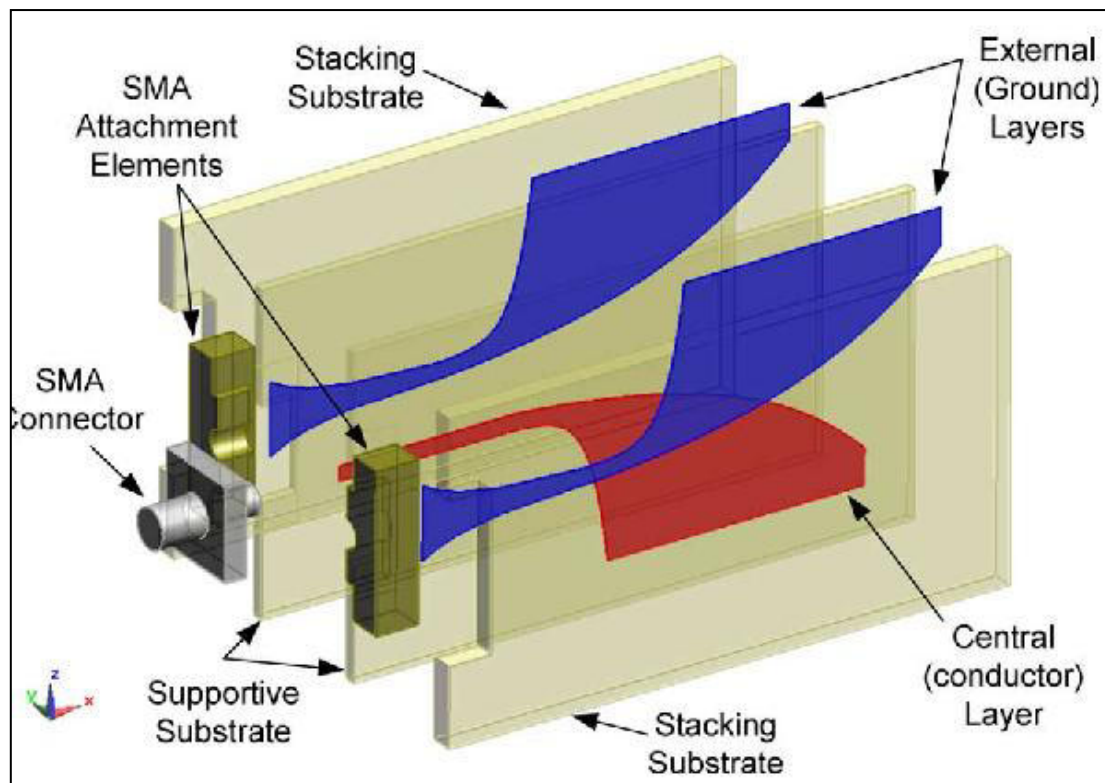


Figure 14 -Antipodal antenna

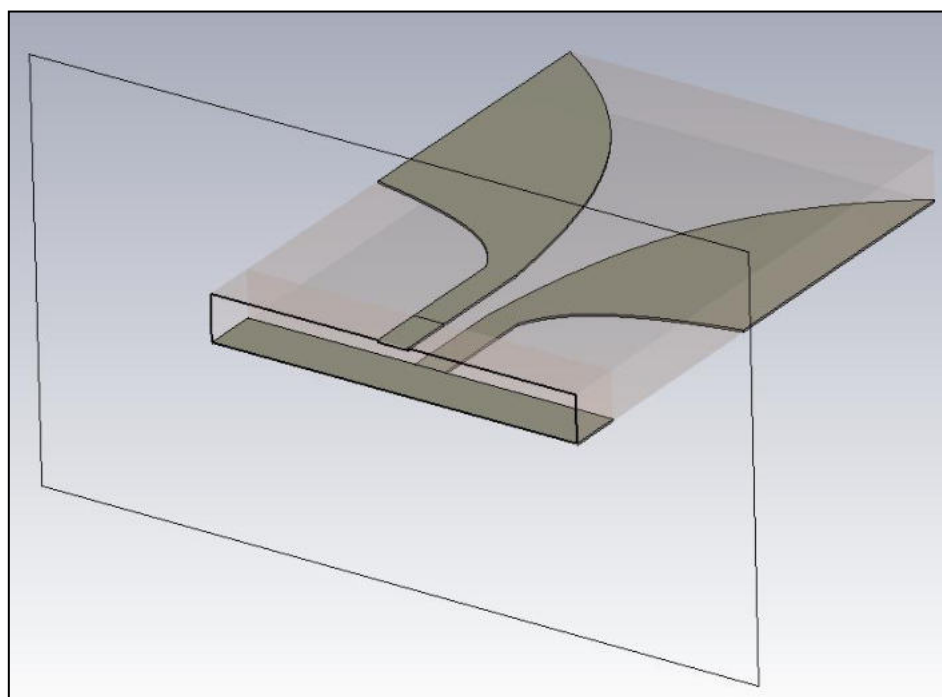


Figure 15 - Antipodal antenna over 32 - 38 GHz bandwidth

1.2.5 Microstrip planar divider

The Wilkinson divider is a specific class of power divider circuit capable of achieving isolation between the output ports while simultaneously maintaining a matched condition at all ports. The input microstrip has an impedance equal to Z_0 , while the two branches of the divider have a length of $\lambda/4$ and an impedance of $\sqrt{2} \cdot Z_0$. Furthermore, an optional shunt resistor connected between the two output branches is included to ensure high port isolation, with a value equal to $2 \cdot Z_0$, as illustrated in Figure 16.

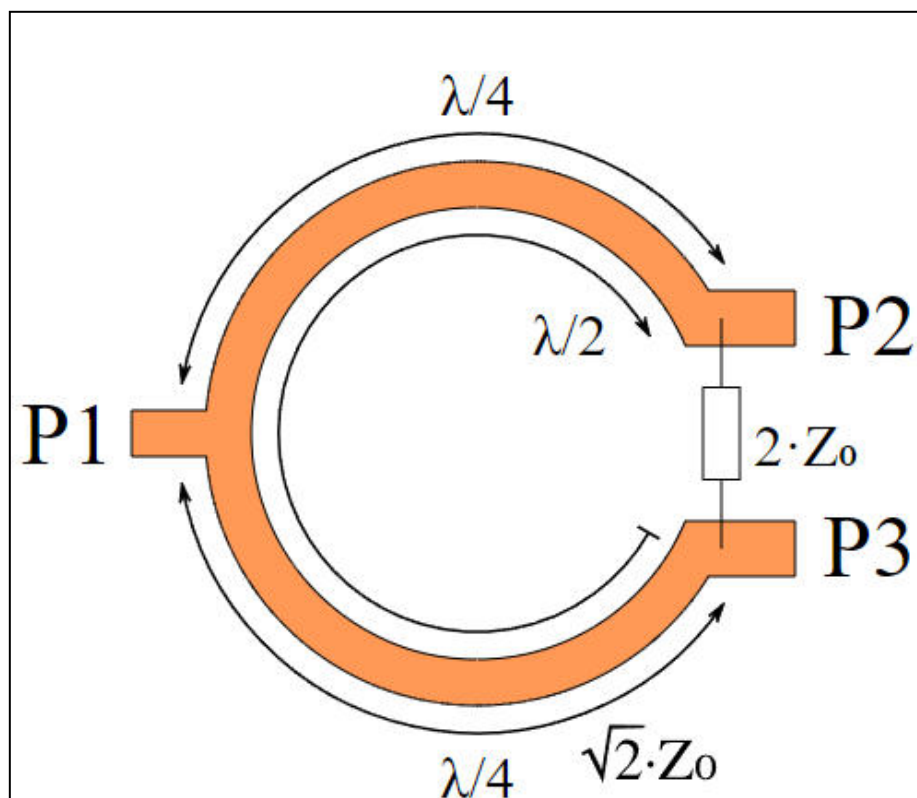


Figure 16 - Wilkinson divider schematic

1.2.6 Another Planar Divider configuration

A possible microstrip planar solution is shown in Figure 17:

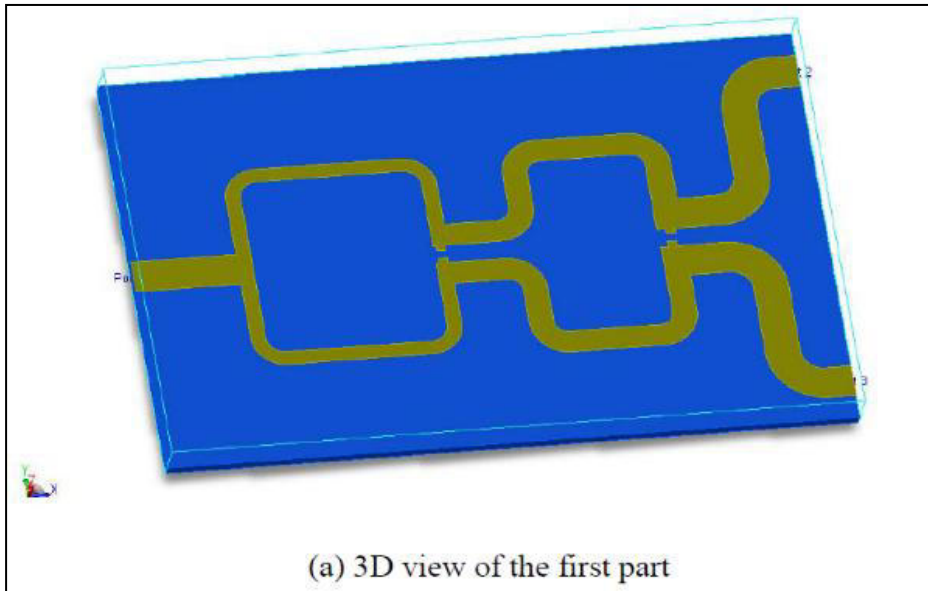


Figure 17 - Wilkinson divider

This type of planar divider has been designed to achieve a wide operating bandwidth. This characteristic is enabled by the presence of metallic trace sections of varying widths, which ensure excellent in-band matching. Furthermore, once the electrical performance of the model is sufficient for the needs, the input connector and the appropriately dimensioned output connectors are added to the model itself.

This planar divider has been designed to operate over a wide bandwidth. Such a feature is achieved by using stepped metallic trace sections of

varying widths, ensuring good impedance matching across the entire operating band.

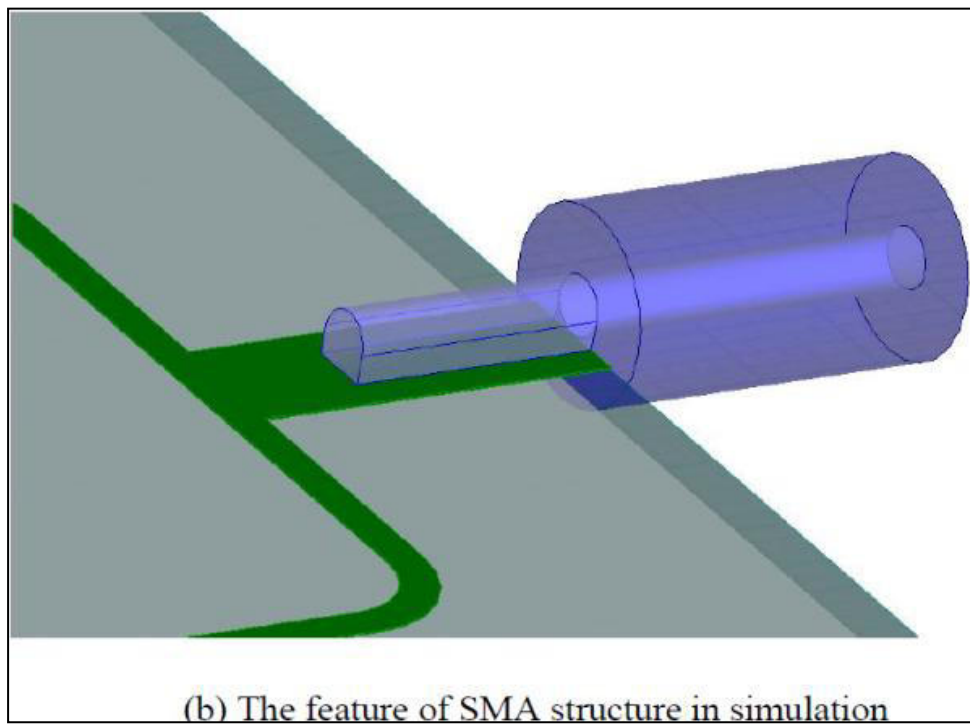
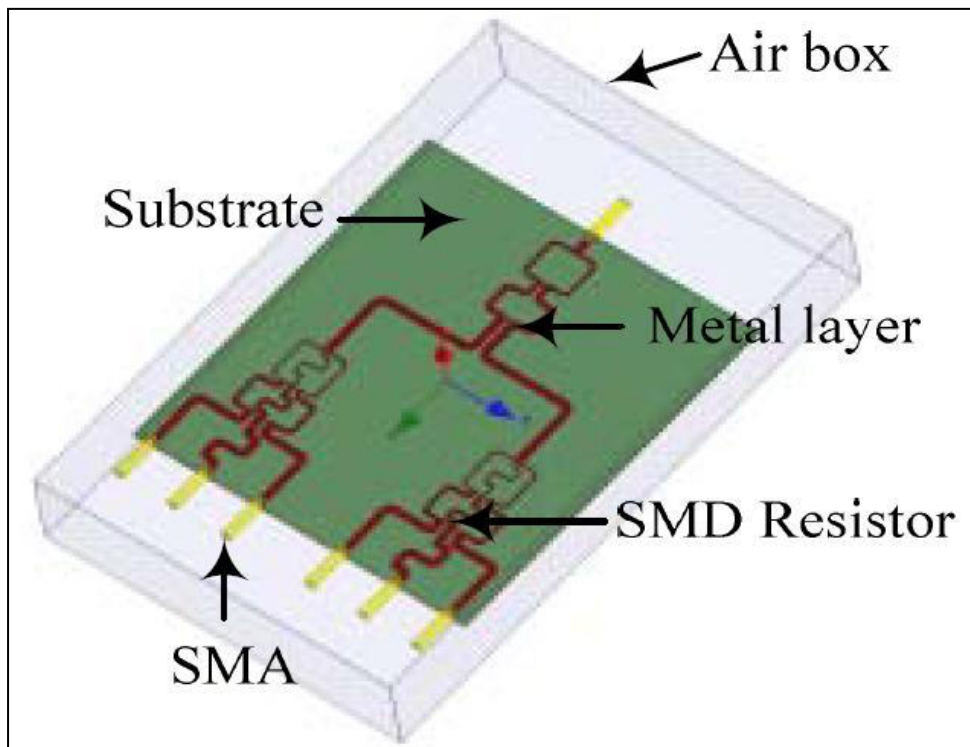


Figure 18 - multiple way planar divider

1.2.7 Wilkinson Power Divider

Wilkinson power divider is a three-port circuit, which schematics are showed in Figure 19, capable of splitting the incoming power at port 1 into two equal output signals at ports 2 and 3. A resistor placed between ports 2 and 3 is used to uncouple those ports ($s_{23} = 0$ in order to avoid reflected waves caused by an unmatched load at one port to affect the other port). The circuit consists of two $\lambda/4$ sections of impedance Z_c connected at one end at port 1 while the other ends are the dividing outputs. The device matching conditions are the following:

$$Z_c = Z_0 \sqrt{2}$$

$$R = 2 Z_0$$

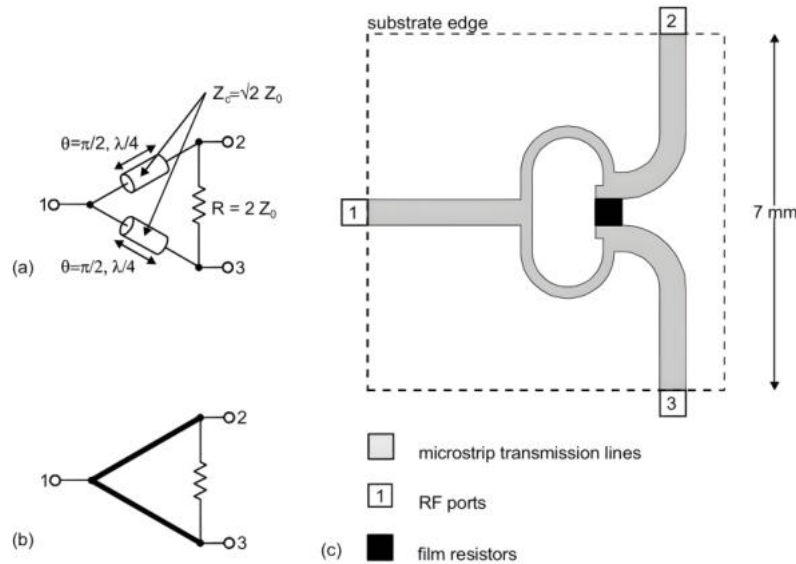


Figure 19 - Wilkinson divider: (a) schematic; (b) electrical symbol; (c) layout of a microstrip realization

The model consists of a copper ground plane with a thickness of 0.035 mm, as well as the trace, while the dielectric used is a lossy Rogers RO 3006. In Figures

below is shown the top, bottom and perspective view of the state-of-the-art planar power divider.

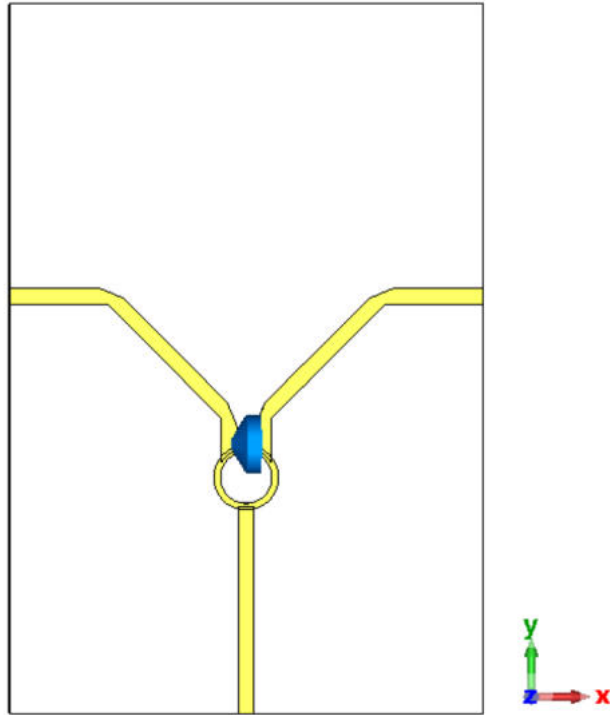


Figure 20 - Top view Wilkinson divider

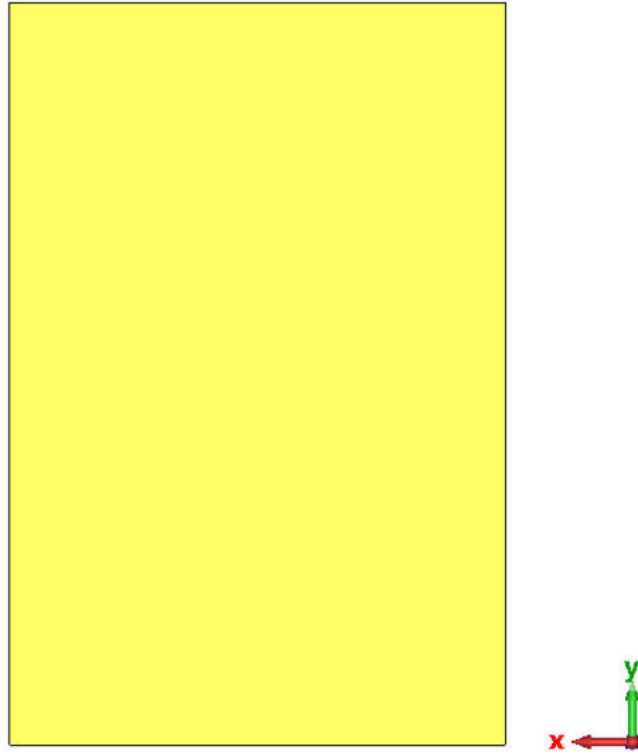


Figure 21 -- Bottom view Wilkinson divider

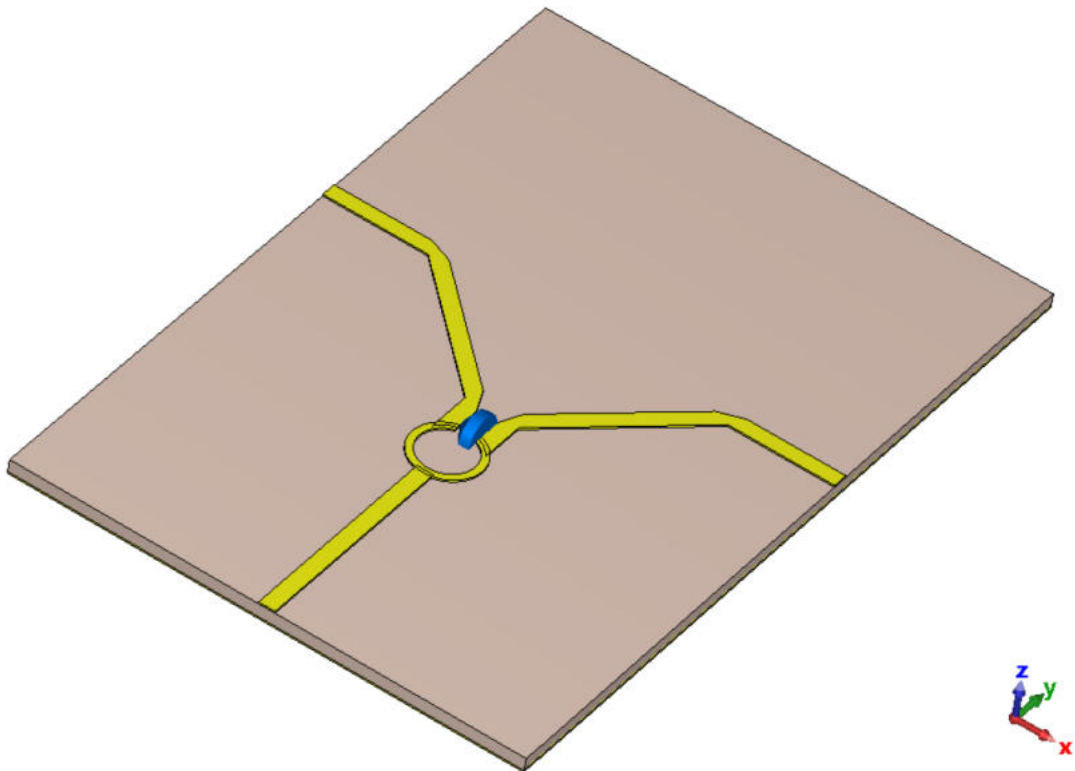


Figure 22 - Perspective view Wilkinson divider (1)

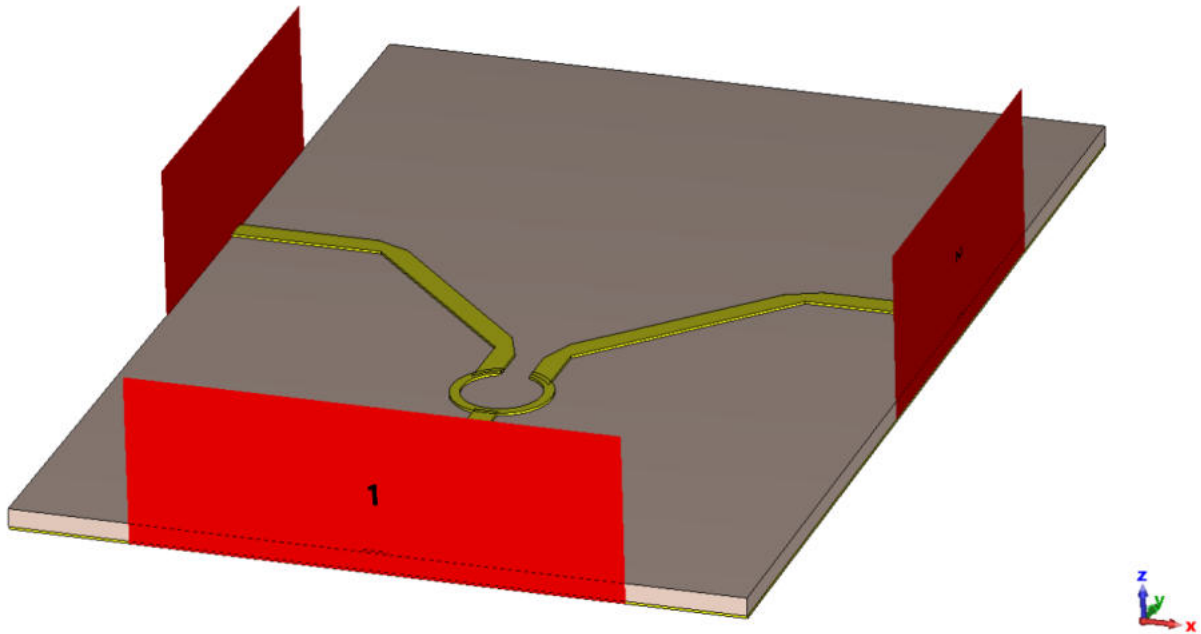


Figure 23 - Perspective view Wilkinson divider (2)

This distribution network is widely used in microwave applications. Signals get routed through critical areas where strict signal integrity must be preserved. It is particularly effective in multilayer PCBs hosting active devices with multiple Input / output ports. In Figures 24 and 25, the Return loss and power division are shown.

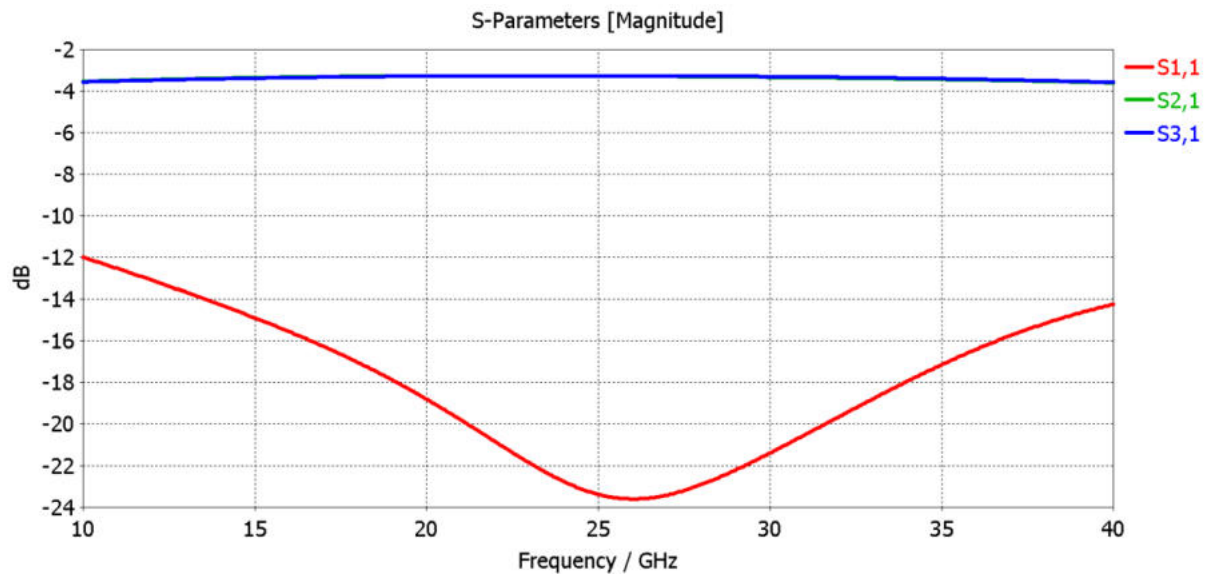


Figure 24 - Return Loss, Power Transmission - Wilkinson divider

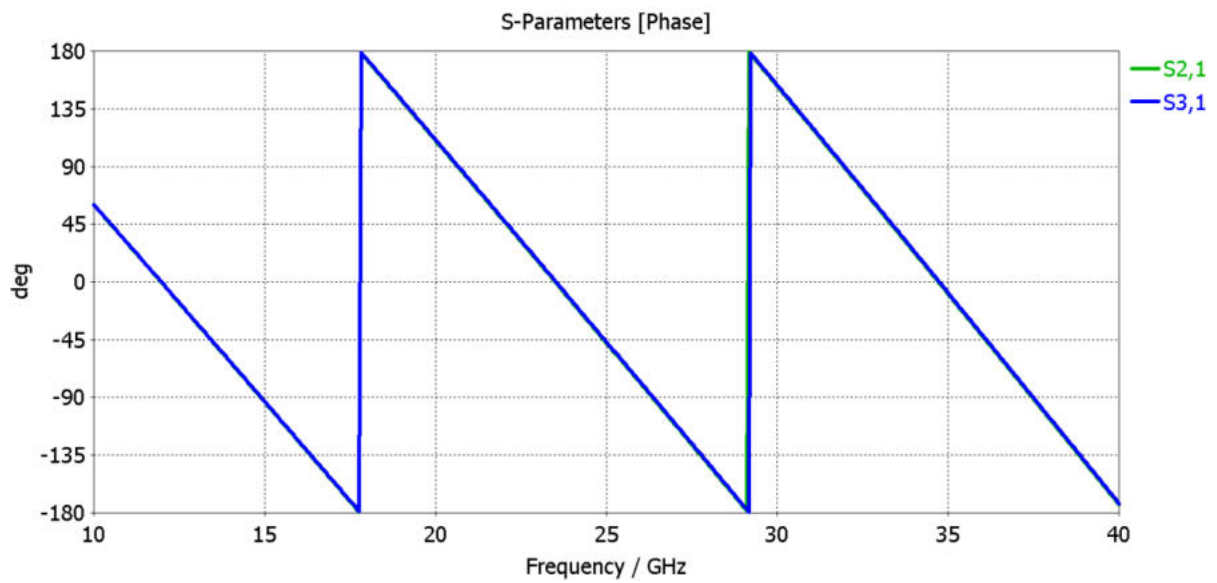


Figure 25 - Power Transmission Phase Signals - Wilkinson divider

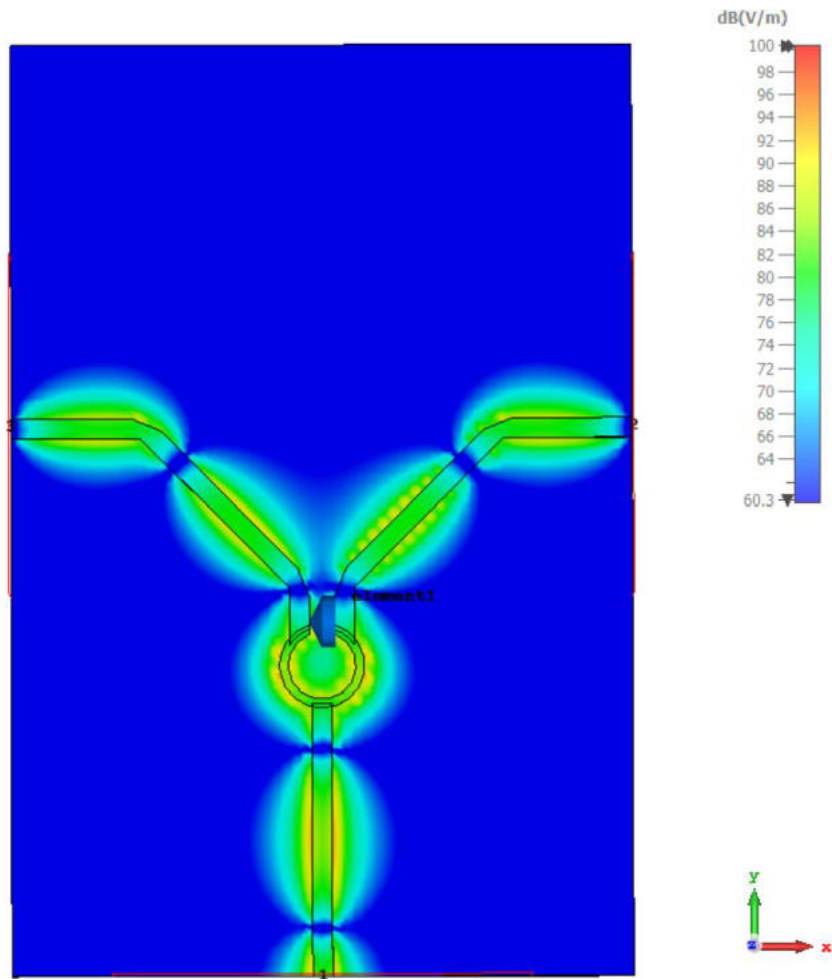


Figure 26 - E field

You may imagine a signal getting input at port 1 and splitting into ports 2 and 3. Or, equivalently, imagine two identical, in-phase signals entering ports 2 and 3 simultaneously. In this case, the mode is an even.

Because the signals are perfectly in phase, the voltage at any given point along the top arm is exactly equal to the voltage at the mirrored point on the bottom arm. The electric field lines concentrate strongly between the signal traces and the ground plane below. However, because there is zero potential difference between the ends of the two arms, no electric field lines cross the gap where the

isolation resistor is located. Meanwhile, the magnetic field, as shown in Figure 27, loops tightly around the current-carrying microstrip traces. The currents in the two arms are flowing in the same direction and at the same magnitude.

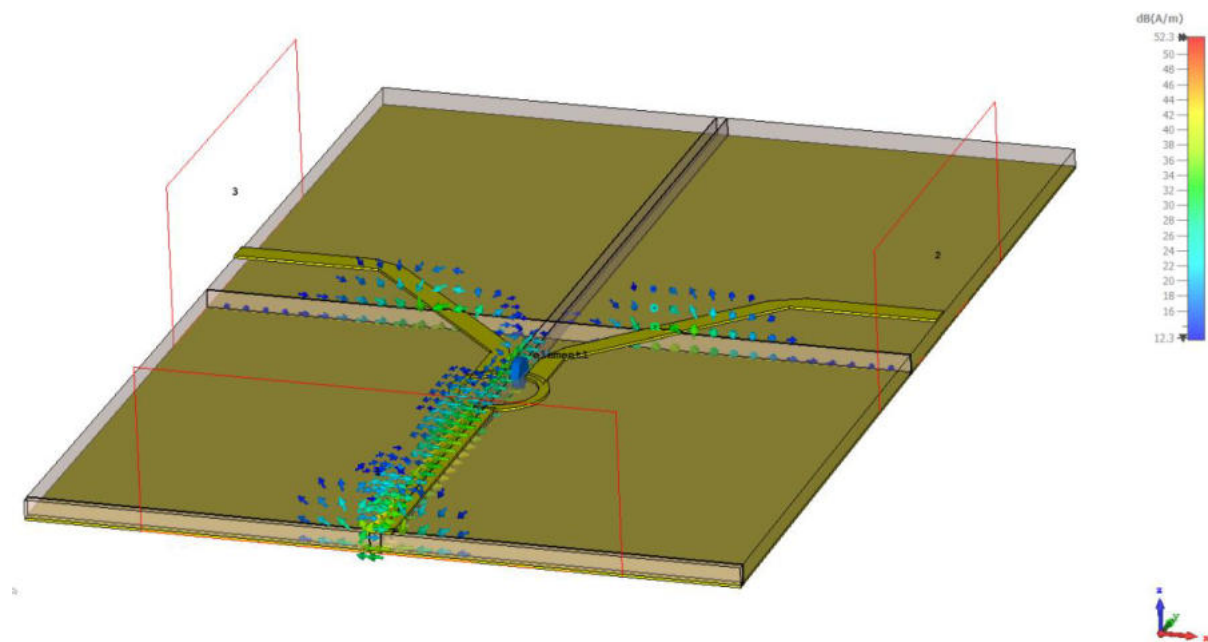


Figure 27 - H - field propagation

Now, if we input two out-of-phase signals into ports 2 and 3, an Odd Mode occurs. In this case, the symmetry is broken. Infact, one arm is at a positive voltage peak while the other is at a negative peak. A plane of anti-symmetry forms directly down the center of the device, creating a "virtual short" to the ground. A strong potential difference exists across the gap between the two output ports. Strong electric field lines bridge this gap. The currents in the two arms are now flowing in opposite directions. The magnetic fields interact, and current is forced to flow through the isolation resistor.

The odd-mode fields never make it back to the input port. This is the mechanism that provides very high isolation. The structure forces any asymmetric fields into the resistor, killing the reflected waves.

1.3 State of the Art

<i>Reference / Component</i>	<i>Component Type</i>	<i>Operating Frequency</i>	<i>Technology & Structure</i>	<i>Insertion Loss / Gain</i>	<i>Return Loss & Isolation</i>
[49]	Dual-Band Filtering Power Divider	Dual-band: • Lower: 2.45 GHz • Upper: 3.72 GHz	Microstrip structure using modified Stepped Impedance Resonators (SIRs) with 5 slots and an isolation resistor.	Minimum Insertion Loss (excl. 3dB split): • Lower band: ~1.4 dB • Upper band: ~1.5 dB	Input Return Loss: • Lower: 13.6 dB • Upper: 14.6 dB Output RL: 16.1 / 15.4 dB Isolation: 14.4 dB (lower) / 23 dB (upper)
[50]	Equal Split Wilkinson Power Divider	K-Band (18.0 GHz – 27.0 GHz)	Microstrip circuit structure consisting of two-step taper stubs designed via empirical equations.	Low Insertion Loss: • Measured: < 0.3 dB across entire range	Input Return Loss: > 22 dB Output Return Loss: > 16 dB Isolation: > 16 dB
[51]	T-Junction SIW Power Divider	K-Band (18.0 GHz – 26.5 GHz)	Substrate Integrated Waveguide (SIW) / Rectangular SIW (RSIW) with metallic rows of holes.	Transmission Coefficients: • S ₂₁ & S ₃₁ are around -3.54 dB (~0.54 dB excess loss over ideal -3 dB split)	Return Loss: • Reflection below -15 dB occupies over 26.32% of the bandwidth
[52]	Power Amplifier MMIC with Bus-bar Combiner	Ka-Band (29.0 GHz – 33.0 GHz)	0.15-μm gate length GaAs pHEMT technology; integrated compact bus-bar power combiner at output stage.	Small-Signal Gain: > 15 dB 1-dB Compression Gain (G _{1dB}): 14.84 dB	Input Return Loss: > 12 dB (< -12 dB) Output Return Loss: > 6 dB (< -6 dB)

[53]	Solid-State Power Combining Amplifier	Ka-Band (34.0 GHz – 36.0 GHz)	Waveguide magic tee corporate network splitting/combining 16 low-power amplifier modules.	High power gain; minimized output stage combining network loss.	High isolation between paths; Optimized input/output reflection coefficients via waveguide E-plane tuning.
[54]	Traveling-Wave Power Divider/Combiner	Ka-Band (28 GHz – 38 GHz)	Low-loss septum unsymmetrical E-plane T-junction series configuration; designed for modular solid-state systems.	Maximum amplitude imbalance of ± 1 dB across the band.	Input Return Loss: > 19 dB Output Reflection: > 15 dB Output Isolation: > 20 dB
[55]	Spatial Power-Combining Amplifier	W-Band (Centered at 94 GHz)	Spatial power combining with cavity modes. Integrates two 1 Watt GaN MMIC power amplifier chips in a 2-way cavity.	Provides a power gain of 9 dB.	Optimized cavity matching (details in text). High path isolation via spatial cavity distribution.
[56]	Spatial Millimeter-Wave Power Combiner	V-Band (50 GHz – 75 GHz)	16-way power combiner based on spatial combining. Employs E-plane bifurcated waveguide, impedance transformers, and resistive films.	Insertion Loss: < 6.5 dB across the entire band (including 12 dB ideal split allocation).	Return Loss: > 12 dB Port Isolation: Highly improved via resistive film integration in branches.
[57]	Planar Six-Way Power Divider	RF / Microwave Band (1710 MHz – 2500 MHz)	Two-dimensional planar structure based on wilkinson Power divider principles	Insertion Loss: 8.0 ± 0.5 dB (including the ideal 6-way split loss of 7.78 dB).	VSWR: < 1.3 for all ports Port Isolation: > 20 dB across the entire frequency range.
[58]	Spatial Power Combiner Amplifier	W-Band (Millimeter-Wave)	6-way power combiner amplifier built around a square cavity operating in the high-order TM ₃₂₀ mode. Integrates six 1 Watt GaN MMIC chips.	Compact modular input split and output combination with low excess loss.	High isolation between input probes, ensuring system resilience against individual device failures.
[59]	N-Way Hybrid Power Divider	Design-dependent (~20% operational bandwidth)	Circularly symmetric network using quarter-wavelength ($\lambda/4$) transmission lines connected to a single, common isolation node via resistors.	Equiphase and equiamplitude split; minimal excess insertion loss over the theoretical	Input VSWR: 1.2 Output VSWR: 1.6 Minimum Isolation: 27 dB between output

				10*log(N) split loss.	terminals at center frequency.
[60]	Broadband Radial Waveguide Spatial Combiner	Microwave Band (8 GHz operational bandwidth)	8-way spatial combiner utilizing coaxial probes embedded symmetrically inside a radial waveguide structure; modeled via an equivalent-circuit electromagnetic method.	Insertion Loss: ~1 dB across the full 8 GHz operating bandwidth.	Return Loss: 10 dB or better across the designated 8 GHz bandwidth window.
[61]	Solid-State Microwave Power Combining Review	Comprehensive Review (X-band, Ku-band, and mm-Wave)	Invited analytical review mapping the taxonomy of solid-state combiners: classifies device-level vs. circuit-level architectures (Kurokawa cavities, corporate networks, radial structures).	Analyzes combining efficiency limits as a function of port count (N) and individual device insertion losses.	Details typical isolation limits and gracefully degrading performance parameters for standard resonant and non-resonant structures.
[62]	Multioctave Spatial Power Combiner	Ultra-Broadband / Multioctave (3.0 GHz – 16.0 GHz; 13 GHz BW)	32-way spatial combiner featuring finline slotline arrays inside an oversized coaxial waveguide. Synthesized with tapered impedance transformers using the Spectral-Domain Method (SDM).	Combining Loss: ~1 dB estimated from small-signal evaluation, corresponding to an overall combining efficiency of ~75% for 32 MMICs.	Maintains stable input/output matching and broadband path isolation across an exceptional multi-octave range.
[63]	Planar Microstrip Three-Way Power Divider	Microwave Band (System design specific)	Planar microstrip network formed by two symmetrically interconnected branch-line directional couplers; evaluated via odd/even mode excitation and transmission line theory.	Equal three-way power splitting with low extra dielectric/conduct or insertion loss.	Demonstrates excellent impedance matching at all ports and high isolation parameters between adjacent output ports.

Table 2 - Comparison with State-of-the-Art High-Power Spatial and Radial Power Combiners.

2 Double side slotted waveguide Power Combiner for Ka band Power Amplifiers

The elementary cell is shown in Figure 28.

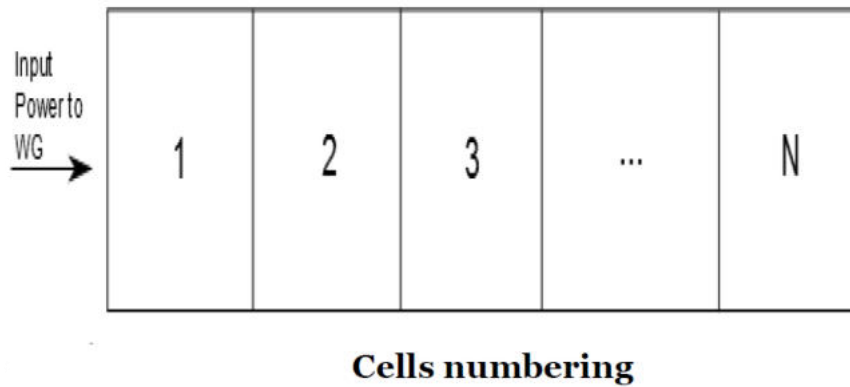


Figure 28 - Elementary cell

Each cell for a N device, each one consisting of 2 slots (top and bottom) and 2 waveguide ports, must satisfy the following S parameters, where I is the cell number.

$$|S_{11}^i|^2 = 0$$

$$|S_{21}^i|^2 = \frac{N - i}{N + 1 - i}$$

$$|S_{31}^i|^2 = |S_{41}^i|^2 = \frac{1}{2(N + 1 - i)}$$

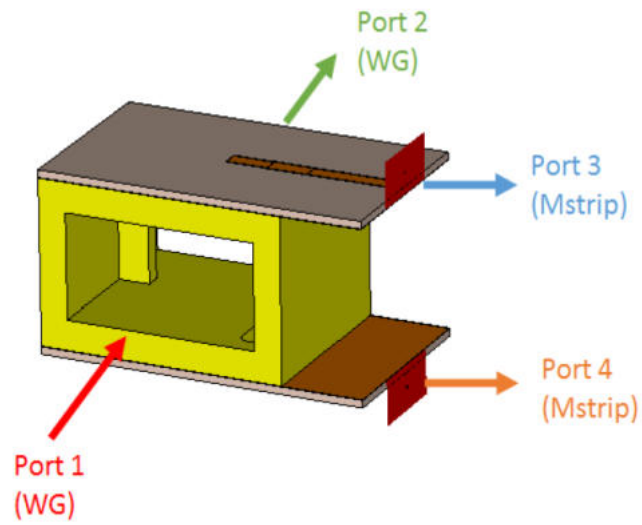


Figure 29 - Single elementary cell structure

Depending on the operating band in which the structure is intended to operate, the size of the rectangular waveguide to be used is chosen according to the Table 3 shown below.

Frequency Band	WaveguideStandard	Frequency Range (GHz)	Inside Dimensions (inches)	Inside Dimensions (mm)	Flange Designation
R Band	WR - 770	0.96 to 1.5	7.700 x 3.385	195.58 x 97.79	
	WR - 650	1.12 to 1.70	6.500 x 3.250	165.1 x 82.55	
	WR - 430	1.70 to 2.60	4.300 x 2.150	109.22 x 54.61	
D Band	WR - 340	2.20 to 3.30	3.400 x 1.700	86.36 x 43.18	
S Band	WR - 284	2.60 to 3.95	2.840 x 1.340	72.136 x 34.036	
	WR - 229	3.30 to 4.90	2.290 x 1.150	58.166 x 29.21	
G Band	WR - 187	3.95 to 5.85	1.872 x 0.872	47.5488 x 22.1488	
F Band	WR - 159	4.90 to 7.05	1.590 x 0.795	40.386 x 20.193	
C band	WR - 137	5.85 to 8.20	1.372 x 0.622	34.8488 x 15.7988	
H band	WR - 112	7.05 to 10.00	1.122 x 0.497	28.4988 x 12.6238	
	WR - 102	7.05 to 11.00	1.020 x 0.510	25.908 x 12.954	
X Band	WR - 90	8.2 to 12.4	0.900 x 0.400	22.86 x 10.16	
Xs Band	WR - 75	10.0 to 15.0	0.750 x 0.375	19.05 x 9.525	
Ku Band	WR - 62	12.4 to 18.0	0.622 x 0.311	15.7988 x 7.8994	
K Band	WR - 51	15.0 to 22.0	0.510 x 0.255	12.954 x 6.477	
	WR - 42	18.0 to 26.5	0.420 x 0.170	10.668 x 4.318	
	WR - 34	22.0 to 33.0	0.340 x 0.170	8.636 x 4.318	
Ka Band	WR - 28	26.5 to 40.0	0.280 x 0.140	7.112 x 3.556	UG-599/U
Q Band	WR - 22	33.0 to 50.0	0.224 x 0.112	5.6896 x 2.8448	G-383/U
U Band	WR - 19	40.0 to 60.0	0.188 x 0.094	4.7752 x 2.3876	UG-383/UM
V Band	WR - 15	50.0 to 75.0	0.148 x 0.074	3.7592 x 1.8796	UG-385/U
E Band	WR - 12	60.0 to 90.0	0.122 x 0.061	3.0988 x 1.5494	UG-387/U
W Band	WR - 10	75.0 to 110.0	0.100 x 0.050	2.54 x 1.27	UG-387/UM
F Band	WR - 8	90.0 to 140.0	0.080 x 0.040	2.032 x 1.016	UG-387/UM
D Band	WR - 6	110.0 to 170.0	0.0650 x 0.0325	1.651 x 0.8255	UG-387/UM
G Band	WR - 5	140.0 to 220.0	0.0510 x 0.0255	1.2954 x 0.6477	UG-387/UM
H Band	WR - 4	170.0 to 260.0	0.0430 x 0.0215	1.0922 x 0.5461	UG-387/UM
J Band	WR - 3	220.0 to 325.0	0.0340 x 0.0170	0.8636 x 0.4318	UG-387/UM
Y Band	WR - 2.8	325.0 to 500.0	0.0279 x 0.0140	0.711 x 0.356	UG-387/UM
	WR - 2.2	325.0 to 500.0	0.0220 x 0.0109	0.559 x 0.279	UG-387/UM
	WR - 1.5	500.0 to 750.0	0.0150 x 0.0075	0.381 x 0.191	UG-387/UM
	WR - 1.2	600.0 to 900.0	0.0120 x 0.0059	0.305 x 0.152	UG-387/UM
	WR - 1.0	750.0 to 1100.0	0.0100 x 0.0050	0.254 x 0.127	
	WR - 0.8	900.0 to 1400.0	0.0079 x 0.0040	0.203 x 0.102	
	WR - 0.65	1100.0 to 1700.0	0.0064 x 0.0032	0.165 x 0.083	
	WR - 0.51	1400.0 to 2200.0	0.0051 x 0.0025	0.130 x 0.065	

Table 3 – Standard Rectangular Waveguide size and frequency range

2.1 4-way Slotted Waveguide Power Divider

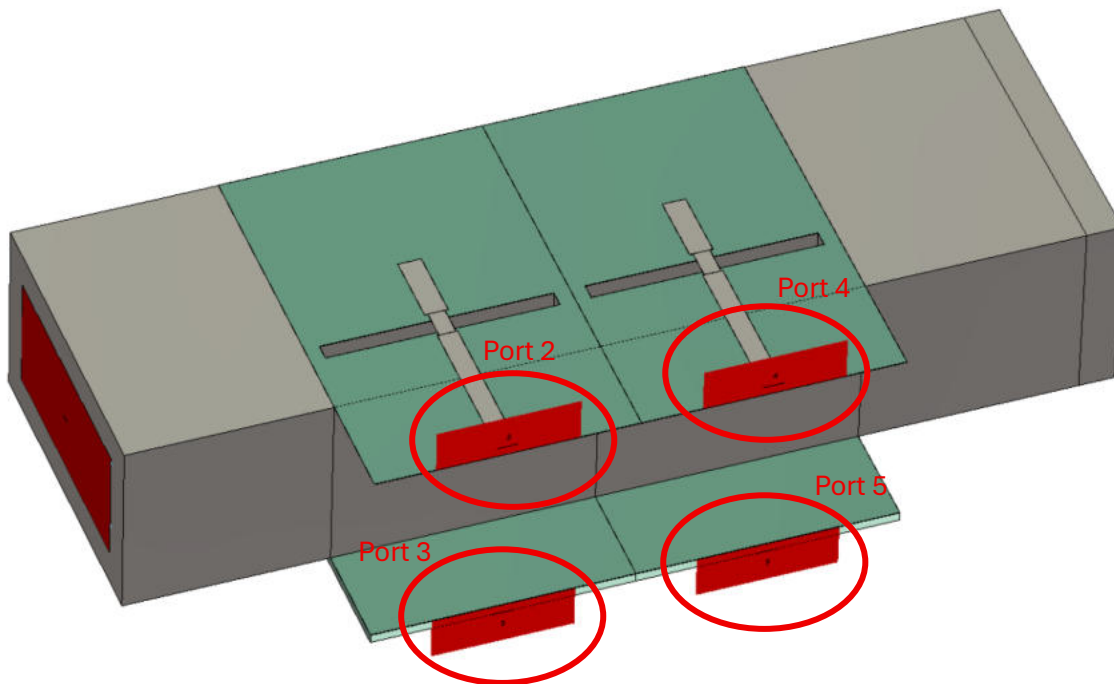


Figure 30 - 4 way Slotted Waveguide Power Divider

The proposed model employs a standard WR-28 rectangular waveguide, which is specifically suited for high-frequency applications in the Ka-band [64], [65]. This waveguide operates over a broad frequency range from 24 to 40 GHz with a central operating frequency $f_0 = 32$ GHz. The broad wall width measures 7.112 mm, while the narrow wall height is $b = 3.556$ mm. From an electromagnetic point of view, the dominant propagation mode within this structure is the transverse electric TE_{10} mode. The theoretical cutoff frequency for this fundamental mode is directly determined by the broad wall dimension and can be calculated as:

$$f_c = \frac{c}{2*a} = 21 \text{ GHz}$$

where c represents the speed of light in a vacuum ($3 * 10^8 m/s$), so that the wavelength is $\lambda_o = 10 \text{ mm}$. Any signal below 21 GHz falls into the cutoff region and is rapidly attenuated as an evanescent wave [66], [67], [68]. By setting the central frequency at 32 GHz, well above the 21 GHz cutoff, the system guarantees stable, single-mode propagation [69], [70].

The proposed transition mechanism employs a rectangular slot to facilitate highly efficient electromagnetic coupling between the transverse electric (TE) mode of the waveguide and the transverse electromagnetic (TEM) mode of the microstrip [71].

To maximize power transfer, the primary slot is located at a distance of approximately 10 mm from the terminating short circuit (backshort). This physical distance corresponds to λ_o [72]. According to transmission line theory, the superposition of incident and reflected waves in a short-circuited waveguide generates a standing wave pattern. In this pattern, the transverse magnetic field reaches its maximum value with a periodicity of an integer multiple of a half-wavelength ($\frac{\lambda}{2}$) from the short circuit. By placing the slot at a full wavelength distance, it perfectly aligns with the maximum of the magnetic field [73], [74].

Furthermore, a second rectangular slot is introduced further along the longitudinal axis to keep guaranteeing the desired coupling profile. This secondary slot is positioned at a distance of approximately 17.43 mm from the short circuit, corresponding to approximately $1.73 * \lambda_g$ (about $2 * \lambda_g$). The specific longitudinal spacing of this second slot is carefully tuned not only to sample the appropriate magnetic field intensity but also to properly manage the phase relationship and impedance matching of the coupled signals [75], [76].

Table 4 lists the overall dimensions involved in the designed structure.

Parameter Name	Variable	Value	Unit
Waveguide wall thickness	wall	1.016	mm
Virtual short line width	vshort_W	0.6617541223253	mm
Virtual short line length	vshort_L	2.0885414694455	mm
Quarter-wavelength transformer width	trans_W	0.47832291681598	mm
Quarter-wavelength transformer length	trans_L	0.86485248215421	mm
Substrate width	subs_W	10	mm
Substrate height	subs_H	0.254	mm
Slot Y offset	slot_yoffset	0.014479204685075	mm
Slot X offset	slot_xoffset	2.4679565211855	mm
Slot width	slot_W	0.47607720203816	mm
Slot length	slot_L	6.5565620233484	mm
Mstrip line offset from slot	mstrip_yoffset	0.017803929468218	mm
Microstrip width	mstrip_W	0.5692041329225	mm
Microstrip length up to PA	mstrip_L	4.6054719660635	mm
Microstrip height	mstrip_H	0.017	mm
Guided wavelength	lambdaG	11.646987318621	mm
Free-space wavelength	lambda	10	mm
VirtualShort connection	connec_short	slot_W/2	mm
Unit cell length	cell_L	7.4341238520529	mm
Backshort length air	backshort_L	6.2672334818269	mm
Waveguide height	b	3.3404321051281	mm
Waveguide width	a	7.112	mm
Bottom virtual short width	Bvshort_W	vshort_W	mm
Bottom virtual short length	Bvshort_L	vshort_L	mm
Bottom transformer width	Btrans_W	trans_W	mm
Bottom transformer length	Btrans_L	trans_L	mm
Bottom slot Y offset	Bslot_yoffset	slot_yoffset	mm
Bottom slot X offset	Bslot_xoffset	slot_xoffset	mm
Bottom slot width	Bslot_W	slot_W	mm
Bottom slot length	Bslot_L	slot_L	mm
Bottom mstrip Y offset	Bmstrip_yoffset	mstrip_yoffset	mm
Bottom mstrip width	Bmstrip_W	mstrip_W	mm
Bottom mstrip length	Bmstrip_L	mstrip_L	mm

Table 4 - Overall dimensions of the structure

2.1.1 Simulation Results

This specific power divider topology operates over a bandwidth from 28 to 32 GHz, exhibiting a good impedance matching at the operating frequency of 30 GHz [77]. At this optimal operating point, the return loss is low, reaching approximately 38 dB. Simultaneously, the power division across the four output ports is measured at -6.6 dB for ports 2 and 3, and -6.9 dB for ports 4 and 5. Moving away from the center frequency towards the band edges, it can be observed that as the impedance matching degrades (the magnitude of S_{11} increases), the power split tends to diverge. This frequency-dependent divergence introduces an amplitude mismatch, resulting in an inconsistent power distribution among the outputs [78], [79], [80]. Consequently, it's clear that this divider provides optimal, highly stable performance within a narrower frequency range of 29 to 31 GHz. Furthermore, the phase curves of the device is well-defined, with the output signals being exactly 180° out of phase in pairs [81]. Figures below show the electrical performance.

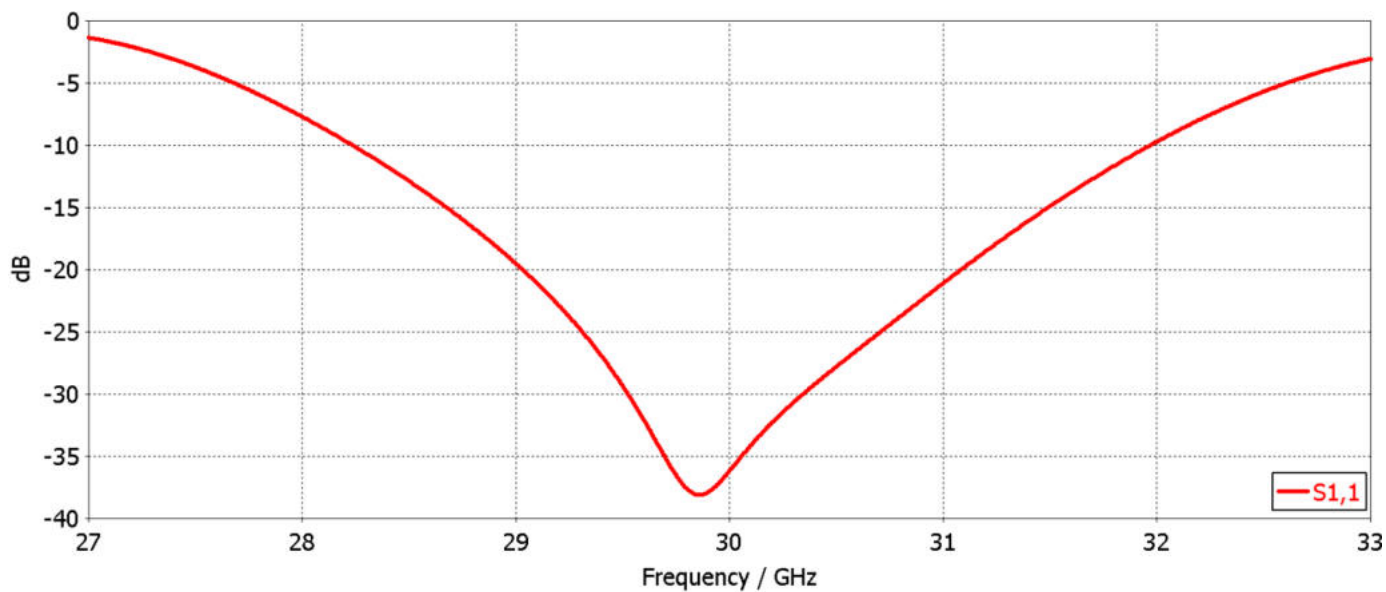


Figure 31 - Return loss over the bandwidth of interest

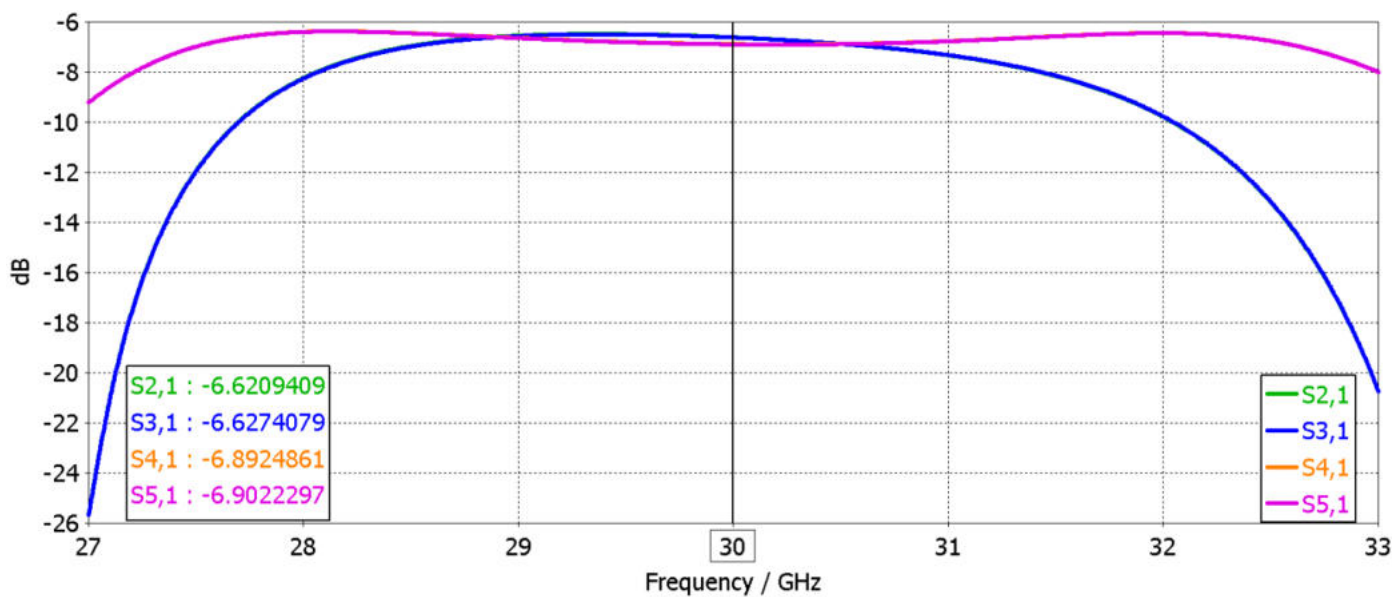


Figure 32 – Amplitude Transmission power division over the bandwidth of interest

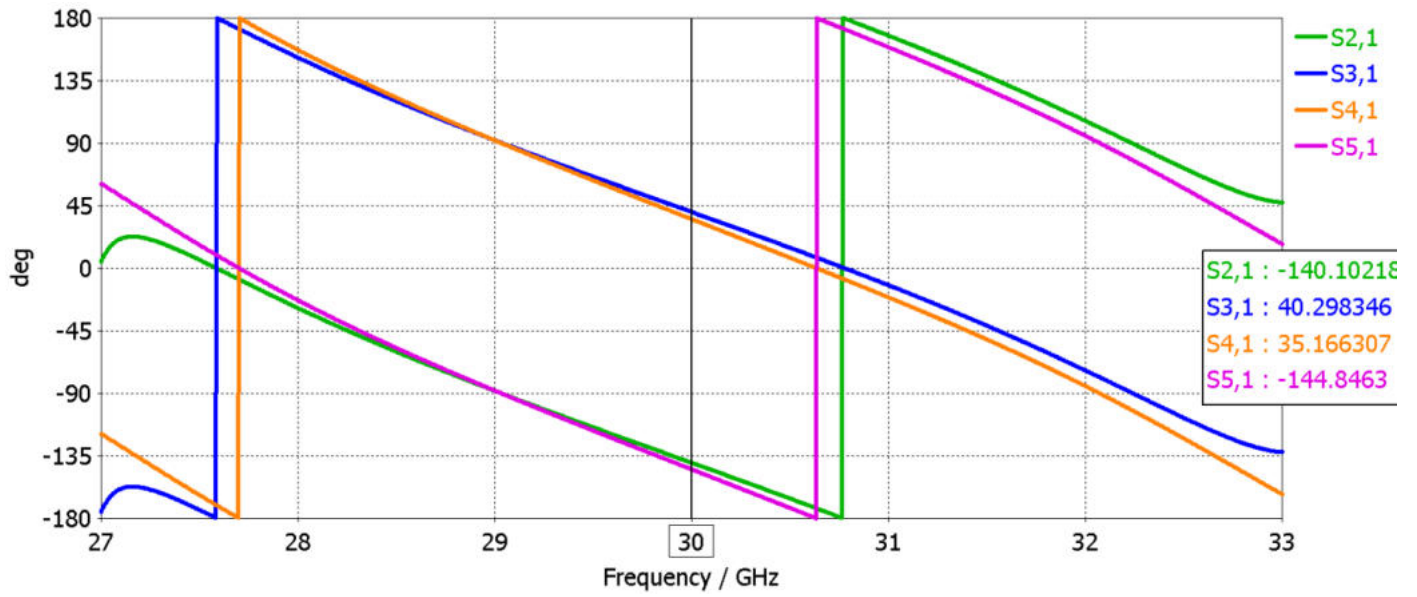


Figure 33 - Phase Transmission power division over the bandwidth of interest

2.1.2 Waveguide to microstrip Magnetic probes

Most power dividers reported in the literature rely on the electric field propagation of the fundamental TE_{10} mode in rectangular waveguide as the primary mechanism for splitting the input power. In such configurations, an electric field probe is inserted into the waveguide structure, coupling to the transverse electric field component and redistributing the incoming power among the output ports [82], [83]. This approach is widely adopted due to its conceptual simplicity and ease of implementation, as the dominant TE_{10} mode naturally presents a strong and well-defined electric field distribution across the broad wall of the waveguide [84], [85] [86].

The present section, however, is motivated by a different and less conventional perspective. Its aim is to demonstrate that a magnetic field probe can serve the very same purpose, offering an equally valid and effective means of dividing the input power [87]. This is made possible by the fundamental nature of the TE_{10} mode itself: as dictated by Maxwell's equations, the magnetic field components H_x and H_z are always oriented perpendicularly to the electric field component E_y , and are therefore spatially well-defined and accessible within the waveguide cross-section [88]. By properly designing a probe that couples to one of these magnetic field components, it is possible to extract and redistribute the input power in a manner fully analogous to the conventional electric field approach, while potentially offering additional degrees of freedom in the overall circuit topology [89], [90].

2.2 Double side slotted waveguide Power Combiner

The 4-way Power Combiner/Divider, illustrated in Figure 34 and Figure 35, is composed of two slotted standard WR-28 rectangular waveguides ($a = 7.112$ mm, $b = 3.556$ mm) and a top-bottom combining structure with microstrip lines. Figures show the designed Power Combiner by using CST Studio Suite 2025. The top-bottom coupling approach enables a more compact design by reducing the architecture length by half compared to one-sided coupling architectures, while maintaining efficient power distribution [91]. Each slot is located at a distance of 0.5 mm from the waveguide wall to achieve optimal magnetic coupling, ensuring minimal insertion loss and high combining efficiency. The microstrip lines are designed to provide a good power distribution to the output ports while maintaining an optimal impedance matching with the waveguide structure [92]. In Figure 38 and 39, the RL and IL curves are shown. Over the bandwidth of interest, the Return Loss is 10 dB above and it reaches 23 dB value at the central frequency of 30 GHz. The average Insertion Loss is around 0.35 dB.

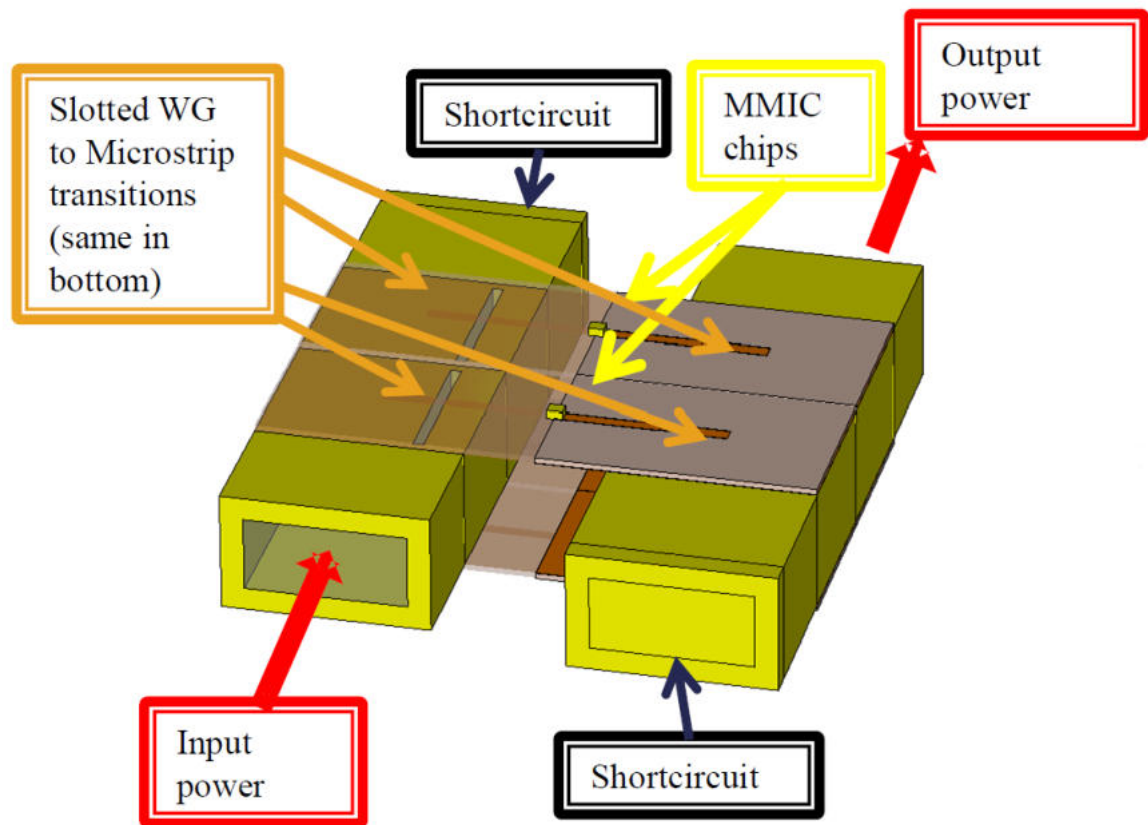


Figure 34 - Double side slotted waveguide Power Combiner in perspective view

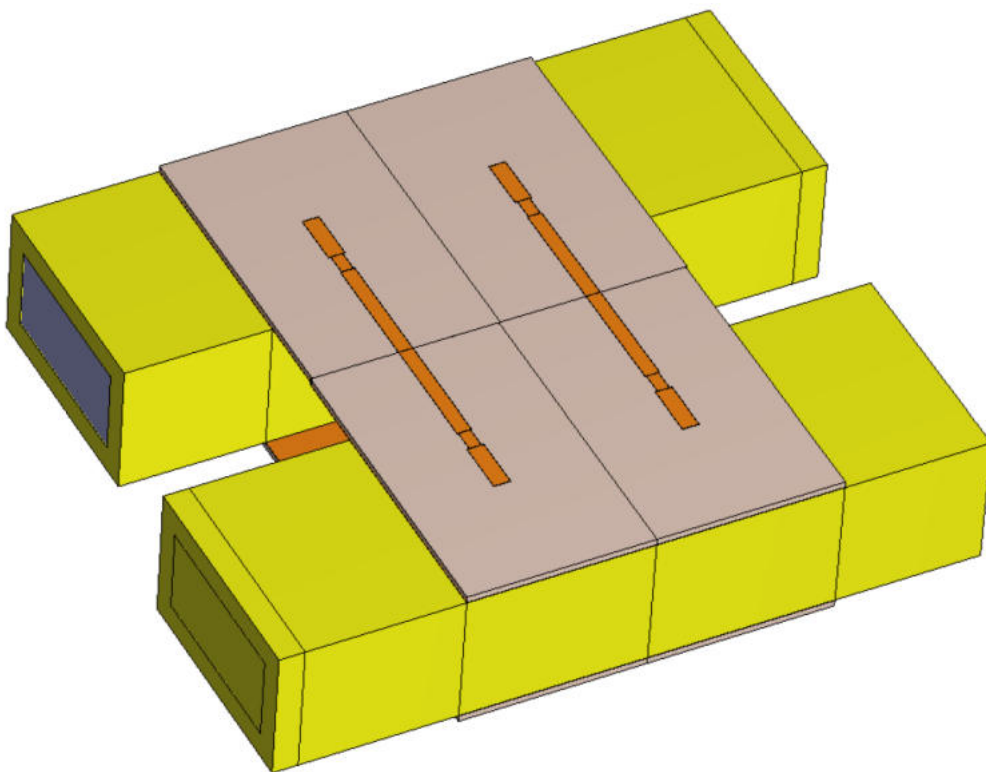


Figure 35 - Back-to-back structure 3D model

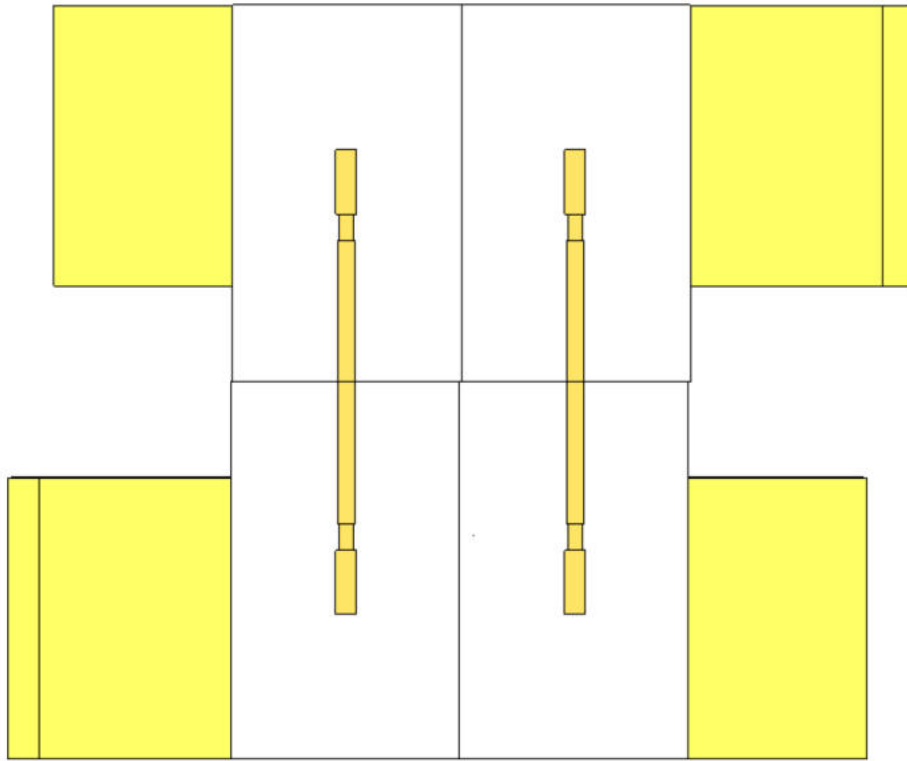


Figure 36 – Power Combiner top view

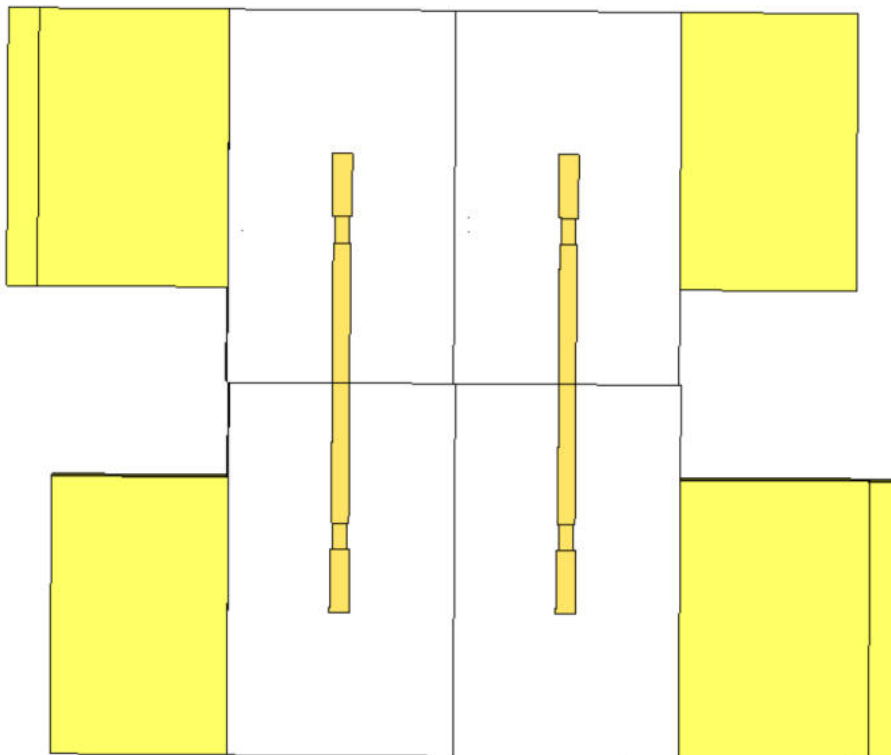


Figure 37 – Power combiner bottom view

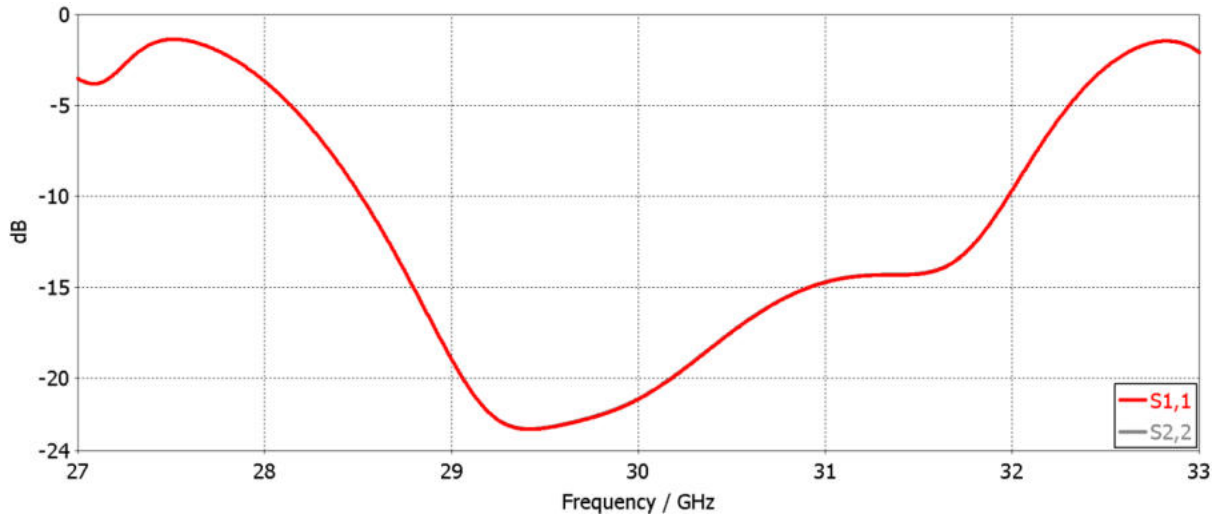


Figure 38 - Return loss

At the center frequency, the Insertion Loss (IL) measures approximately 1.27 dB, exhibiting a deviation from the ideal theoretical value of 0 dB.

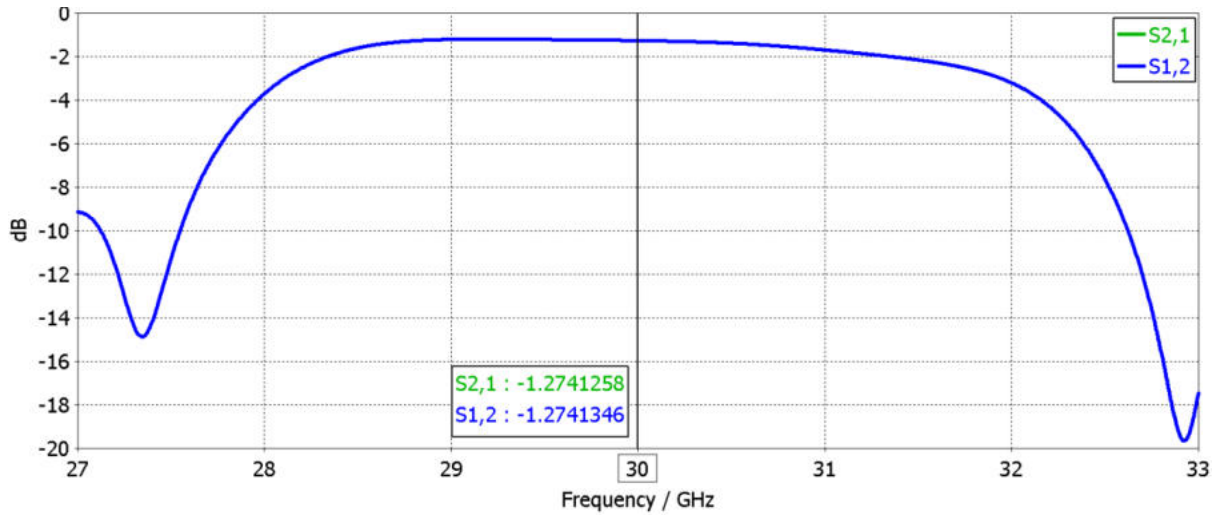


Figure 39 - Insertion loss

This discrepancy can be primarily attributed to the physical coupling mechanism [93]. Specifically, as the electromagnetic field couples with the microstrip through the rectangular slot, a mode transition occurs from the dominant TE_{10} mode within the waveguide to a quasi-TEM mode along the microstrip line. Since

the microstrip is currently surrounded by an air environment, a significant portion of the RF power is radiated into free space. To effectively improve the insertion loss and mitigate these radiation losses, the microstrip structure should be enclosed within a metallic shielding cavity, ensuring that the electromagnetic field remains strictly confined within the device [94].

2.2.1 Magnetic Coupling Mechanism:

The transition between waveguide and microstrip in the power combiner is achieved through slots placed at the broad walls of the waveguide. These slots act as discontinuities that allow the coupling of electromagnetic energy from the TE_{10} mode in the waveguide to microstrip transmission lines. The coupling strength is determined by the slot length and position related to the EM field distribution in the waveguide [95], [96]. The slots are initially designed with lengths between $\lambda_0/2$ and $\lambda_g/2$ (where λ_0 is the free-space wavelength and λ_g is the guide wavelength). In Figure 40 is shown the maximum E- field distribution along the transversal cross-section of the combining structure.

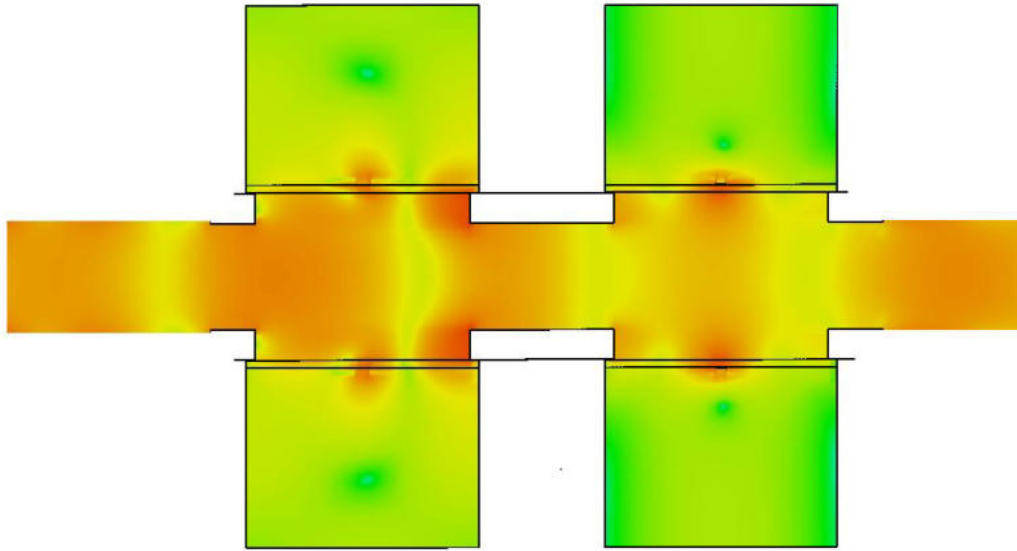


Figure 40 - E field distribution, transversal cross-section.

It's evident how the maximum electric field tends to be not perfectly homogeneous when moving from one cell to the next one. In fact, there are areas inside the cell where the Electric field is less intense compared to the other one. That's why this architecture based on waveguide to microstrip transition is not effective when the number of microstrips increases significantly. For this reason, in the following section a new architecture is presented with more than four microstrips in which good performances are guarantee.

3 Proposed novel 8-way power combiners

3.1 Advanced 8-ways Power Combiner 90-degree phase shift mode converter

In Figure 41 is shown a 90-degree phase shift mode converter.

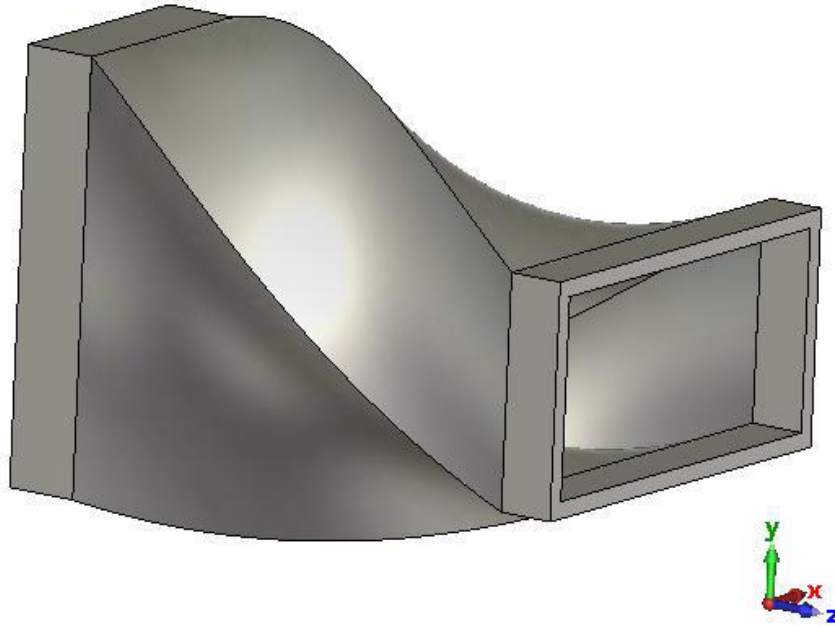


Figure 41 - Mode converter

The 90-degree phase shift mode converter ensures the connection between the input/output T-junction power divider and the two 4-way combiners. The goal of the 90-degree waveguide rotation is to allow the fundamental TE_{10} mode to undergo a 90-degree phase shift. It does not affect the amplitude characteristics of the fundamental mode. Meanwhile, it enables phase rotation thereof, as illustrated in figure below. Consequently, it does not impact system-level

beamforming or combining performance. Figures 42 and 43 show the E- field distribution of the fundamental mode in the E-plane waveguide and the H-plane waveguide, respectively. The E-field distribution in the H-plane waveguide represents the input electric field at the 4-way power combiner.

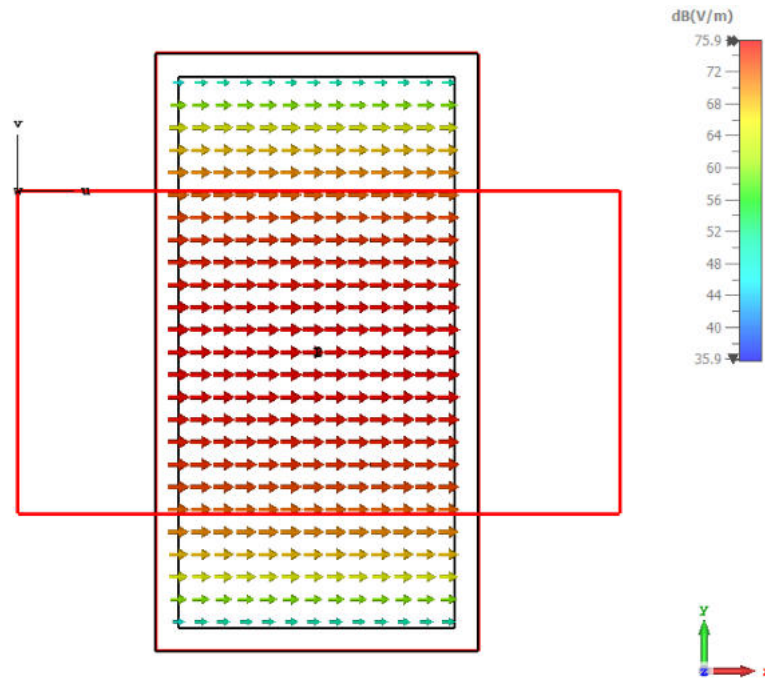


Figure 42 - E field distribution in an E-plane waveguide

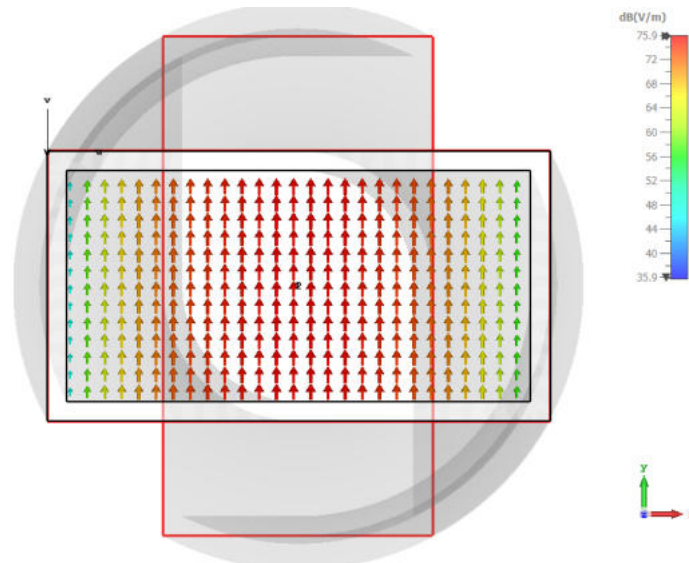


Figure 43 - E field distribution in a H-plane waveguide

An advanced and easily scalable model of a novel 8-way power combiner. In Figure 44 is shown the 3D perspective view of the 8 way advanced power combiner.

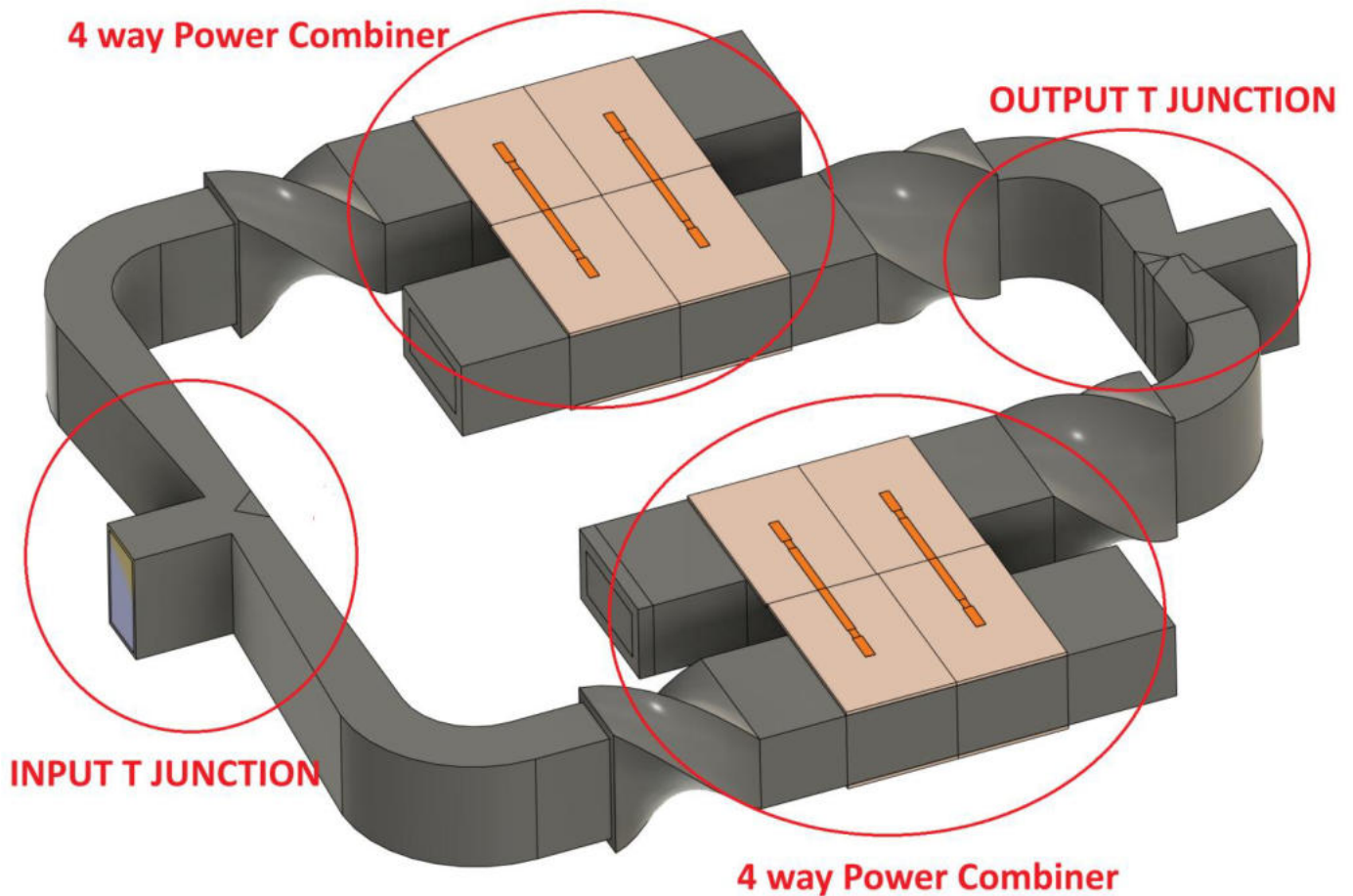


Figure 44 - 3D model of the 8-ways waveguide to microstrip Power Combiner

The architecture consists of two waveguides to microstrip transition 4 ways power combiners arranged symmetrically around a central input port. The design has two tapered waveguides and curved transitions to ensure optimal impedance matching and low losses [97]. Each branch includes a waveguide section with rectangular cross-section and internal septum for enhancing performance characteristics such as return loss and isolation. The use of smooth twists and

bends guarantees to maintain fundamental TE_{10} mode purity and reduce reflections throughout the signal path. The positioning of the microstrip elements has been carefully optimized to achieve optimal electromagnetic performance. Through comprehensive analysis of the vectorial Electric field distribution, computed at the center frequency of 30 GHz, it becomes evident that the electromagnetic coupling characteristics demonstrate exceptional efficiency, as shown in Figure 45.

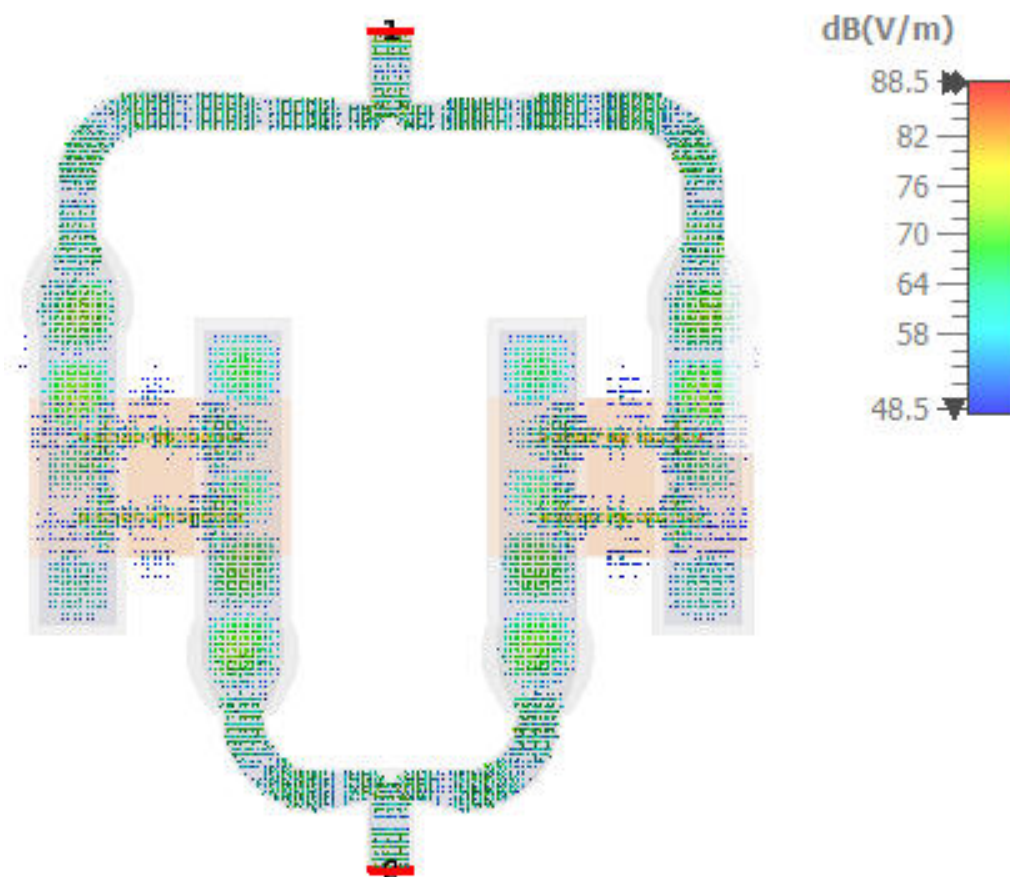


Figure 45 - E-field distribution of the 8-ways waveguide to microstrip Power Combiner with 90° phase shift mode converter

The Electric field vector exhibits high intensity, as indicated by the green color in the field distribution map, particularly in the critical regions nearby the WR-28

waveguide to microstrip transition zones. The strong E-field distribution demonstrates that minimal reflection losses and a good isolation have been achieved at the operational frequency.

3.1.1 Proposed model and simulation results

As shown in Figure 46 and Figure 47, the proposed power combiner exhibits an operating bandwidth of approximately 4.5 GHz. Within this frequency range, the device presents resonant behaviours at 30.2 GHz and 31.2 GHz, related to the points of optimal impedance matching [98]. Furthermore, the average insertion loss is exceptionally low, maintained at approximately 0.4 dB across the band. This minimized loss profile is directly attributable to the combined implementation of the rectangular waveguide and a fully enclosed metallic shielding structure located above the microstrip lines. By adopting this specific shielded topology, the microstrip is completely prevented from radiating electromagnetic energy into the surrounding environment, thereby virtually eliminating radiation losses and maximizing the overall transmission efficiency.

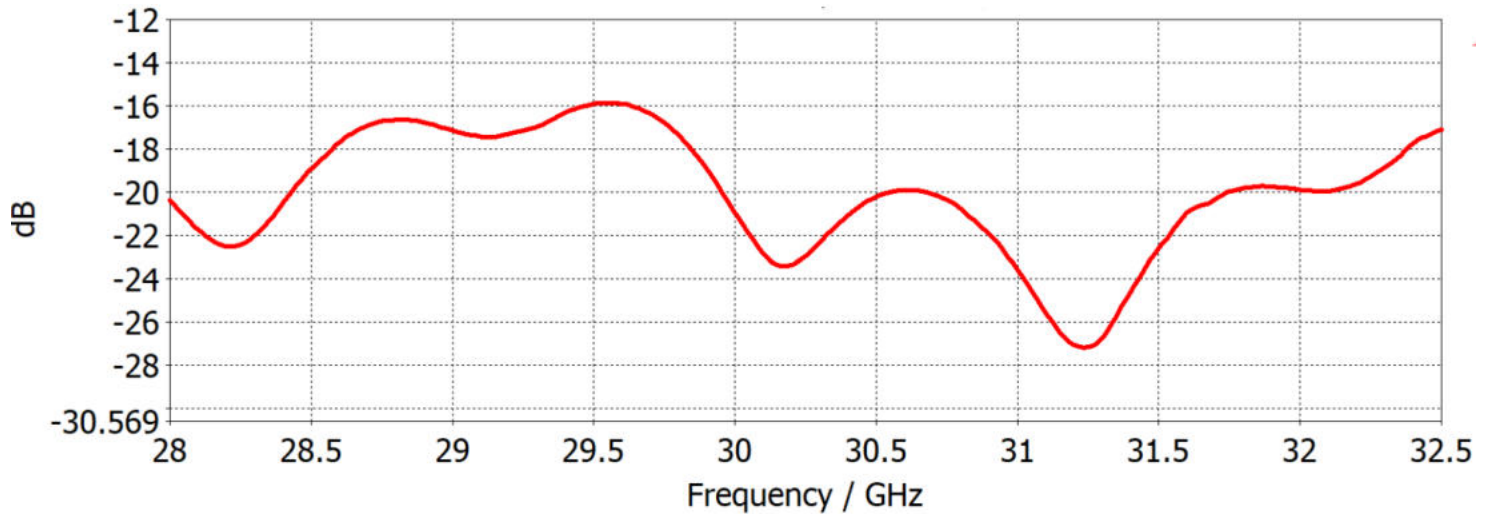


Figure 46 - Return loss

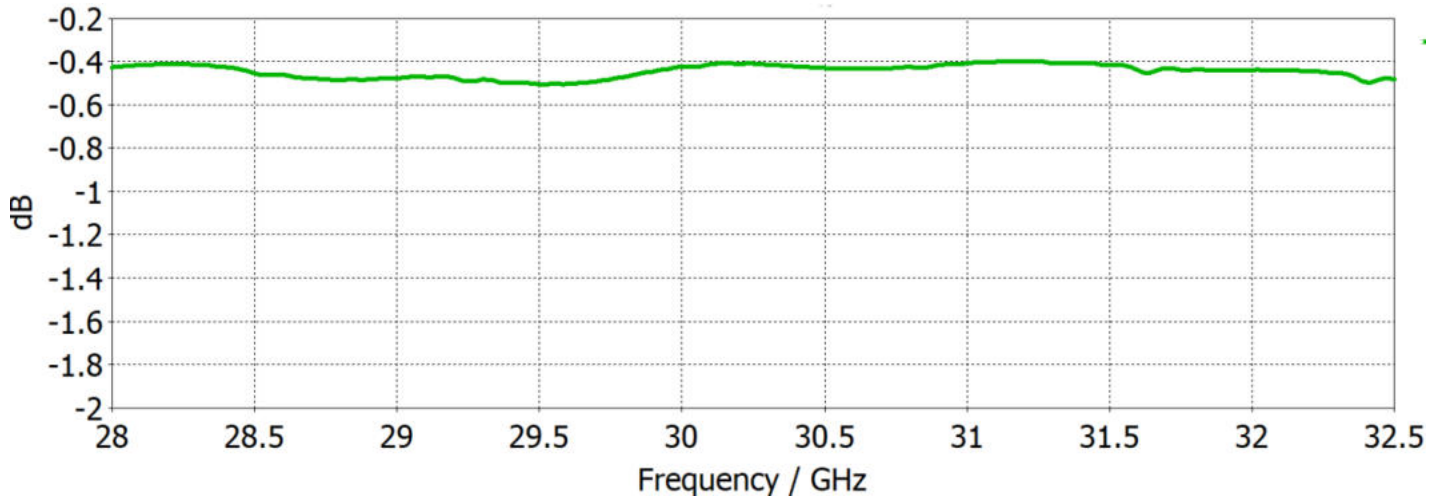


Figure 47 - Insertion loss

The following formula indicates the Power Combining Efficiency (PCE):

$$\eta \approx \frac{P_{\text{out}}}{NP_{\text{in}}} \times 10^{-\frac{IL}{10}} * 100 = 90\%$$

3.2 Coaxial to stripline 8-way Power Divider

Planar combining architectures (Wilkinson, hybrid couplers) suffer from ohmic and dielectric losses in the transmission line network, which become very high as frequency such as port number grow as well. Waveguide-based spatial combining solutions offer a good alternative, as the electromagnetic field remains fully confined within the metallic enclosure during propagation, enabling combining efficiencies exceeding 60% even for large port counts.

Simulations of the divider assume an all-air dielectric surrounding the stripline conductor, which minimizes dielectric losses and yields the best RF performance. However, an air-suspended stripline located precisely at the midplane of a waveguide section cannot be mechanically self-supporting in a real fabricated device. Two practical solutions are proposed to address this:

- **Thin metallic support rods** with small diameter placed at strategic points along the strip, ideally small enough to negligibly perturb the EM field distribution.
- **PCB-insert approach:** the stripline is realized on a thin dielectric PCB substrate (with a dielectric constant approximately around 1), which slides into the waveguide cavity as if the waveguide body acts as a pocket or housing for the board; the assembly is then sealed hermetically with a top

cover and screws, ensuring mechanical stability and environmental protection.

The PCB-insert approach is more compatible with standard manufacturing processes and allows the use of well-controlled substrate materials (e.g., Rogers laminates), though a careful re-optimization of the structure geometry will be required to account for the dielectric loading introduced by the substrate.

The structure under investigation is a radial waveguide power divider in which the output ports are arranged symmetrically around the central axis and oriented perpendicularly to the input port. The input is fed via a coaxial connector, which excites the dividing structure directly [99]. At the center of the device, a coaxial-to-stripline transition converts the input signal into a TEM mode propagating along the stripline network. This is followed by a TEM-to-TEM mode transition that distributes energy equally among the output branches. Ensuring the correct electromagnetic coupling between branches at this junction is critical, and it is achieved through a metallic cylinder (called metallic ring) whose geometry is optimized to enforce the proper boundary conditions and field continuity across the transition. Figure 48 shows the proposed model.

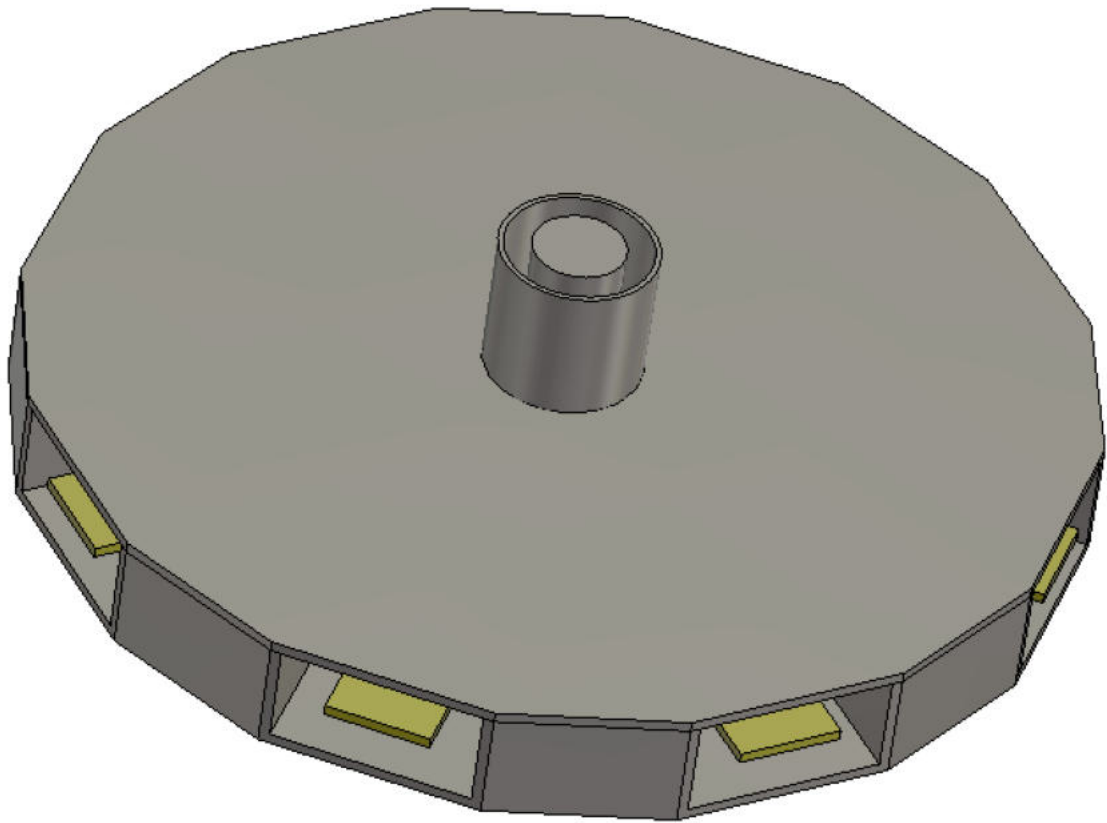


Figure 48 - Coaxial to stripline transition 8 way Power divider – Perspective view

	Value	Scale
Overall size	34.66 x 4.05	mm

Variable	Calculated Value (mm)	Original Expression	Notes
raggioesterno_cilindro	9.1	9.10	Outer cylinder radius
raggiointerno_cilindro	8	8	Inner cylinder radius
profondità_cilindro	5	5	Cylinder depth
x_max	7.4	$6.7+2*0.35$	Max X dimension
profondità_stripline	3.7	$3.35+0.35$	Stripline depth
raggio_cilindro_interno	2	2	Inner cylinder radius (small)
lunghezza_cilindro_interno	0.2	0.2	Inner cylinder length
profondità_calza_esterna	1	1	Outer shield depth
raggio_coassiale_grosso	1.5	1.5	Large coaxial radius
raggio_coassiale_centrale	1.65	1.65	Central coaxial radius
raggio_coassiale_superiore	0.965	0.965	Upper coaxial radius
lunghezza_conduttore_inferiore	5	5	Lower conductor length
raggio_protusione	2.405	$4.81/2$	Protrusion radius
profondità_protusione	0.85	0.85	Protrusion depth
raggio_interno_anello	2.45	2.45	Ring inner radius
raggio_coassiale_alto	1.5	1.5	High coaxial radius

Table 5 - overall dimensions of the 8-way power divider

Figures below show the model from different viewpoints. The yellow part represents the suspended stripline, the red part represents the metallic ring, while the grey part represents the mechanical structure.

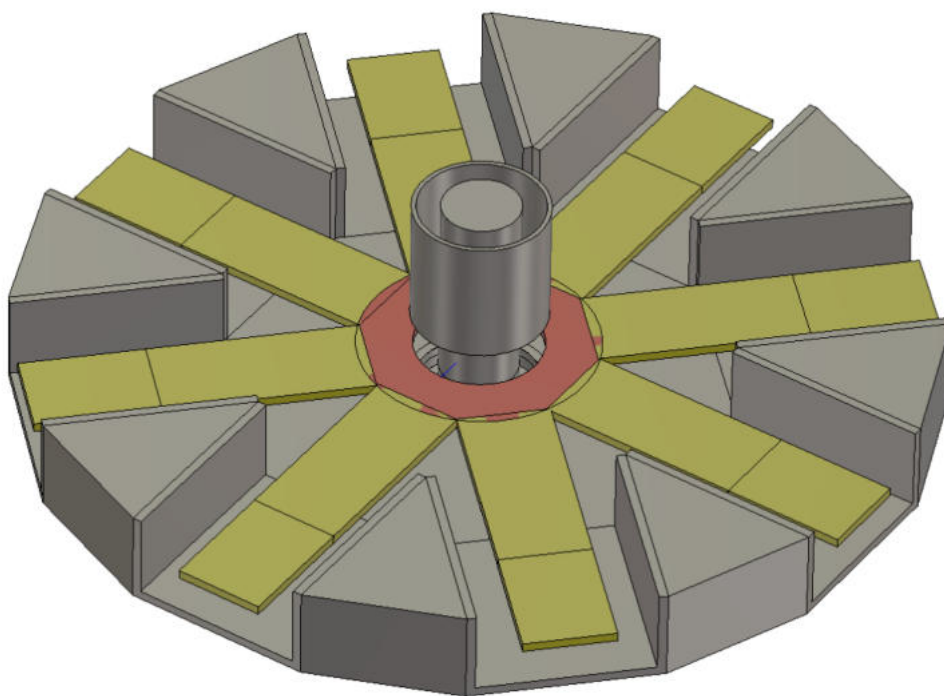


Figure 49 - Coaxial to stripline transition 8 way Power divider – Perspective view - (no cover)

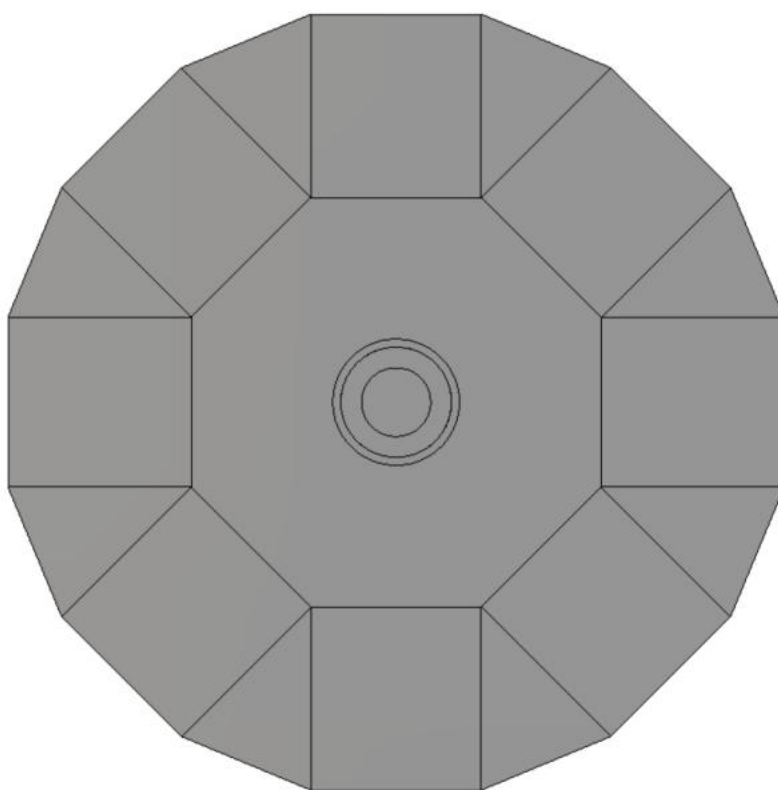


Figure 50 - Coaxial to stripline transition 8-way Power divider – Bottom view

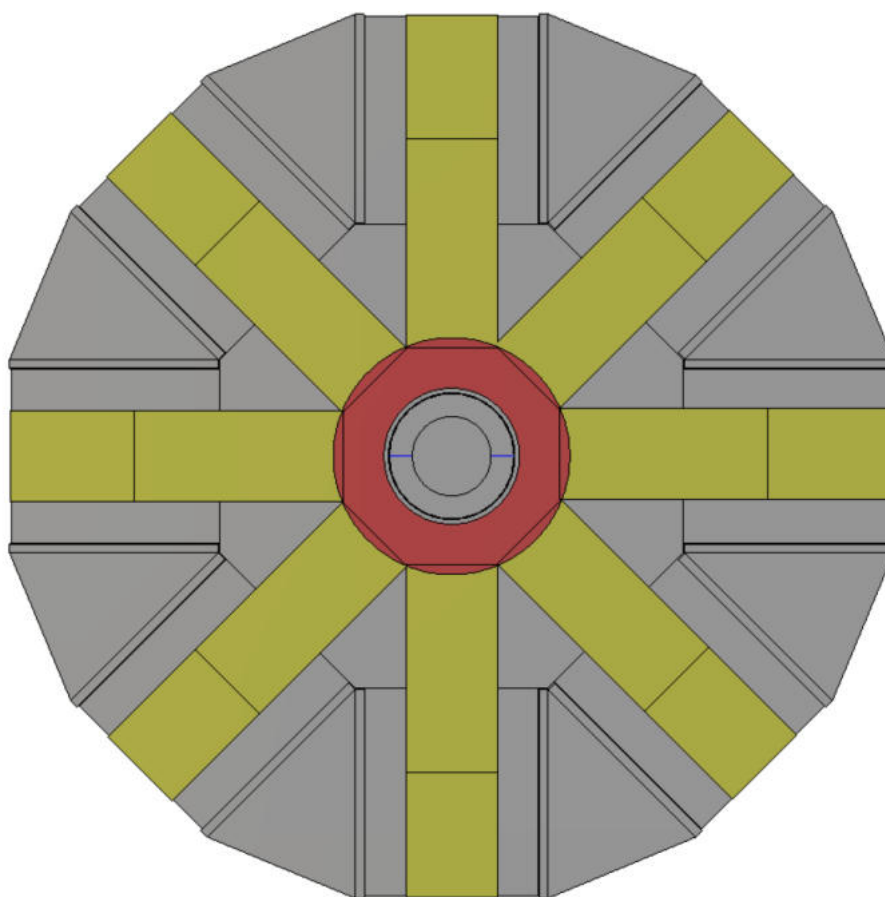


Figure 51 - Coaxial to stripline transition 8-way Power divider – Top view

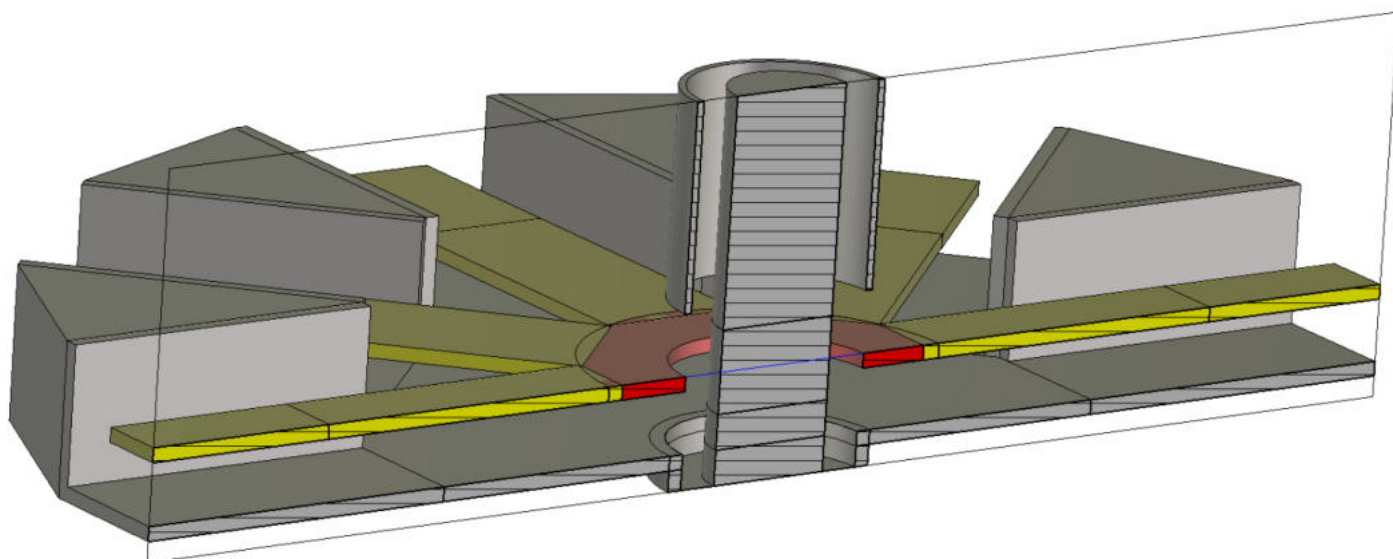


Figure 52 - Coaxial to stripline transition 8-way Power divider – Crossection

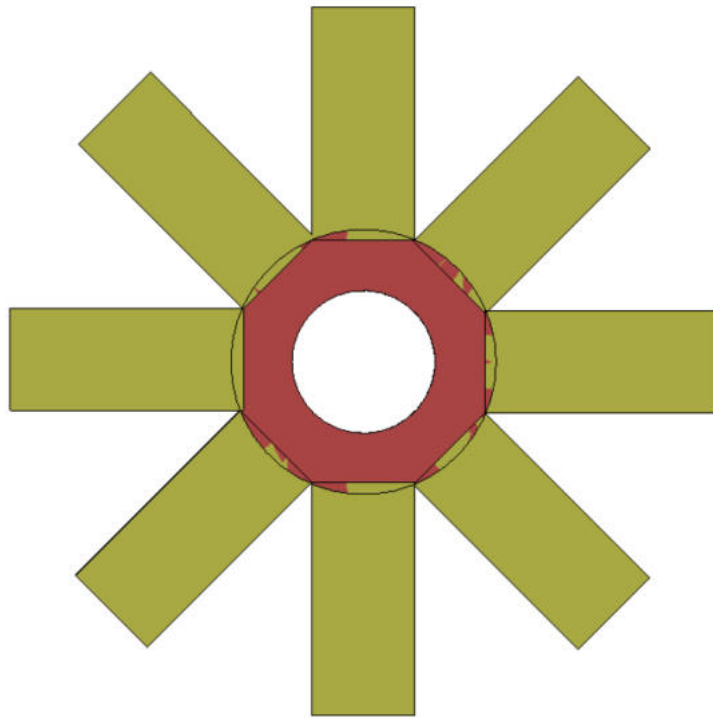


Figure 53 Metallic ring with 8 copper strips

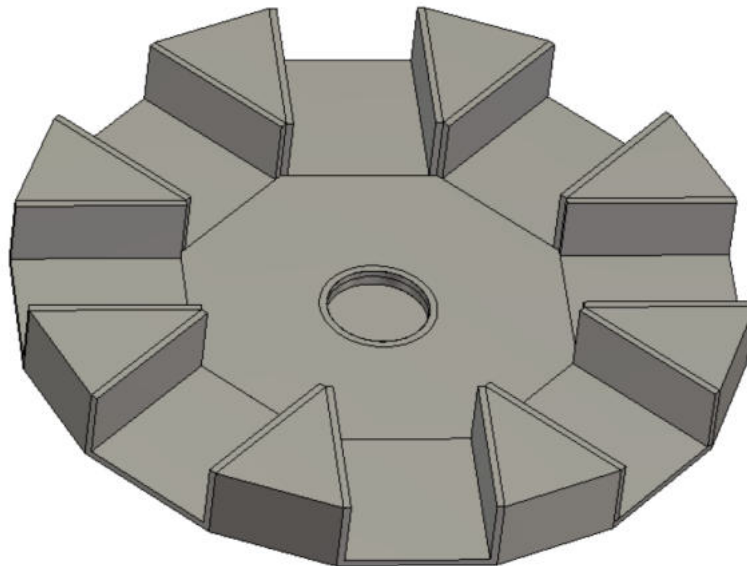


Figure 54 - Bottom side

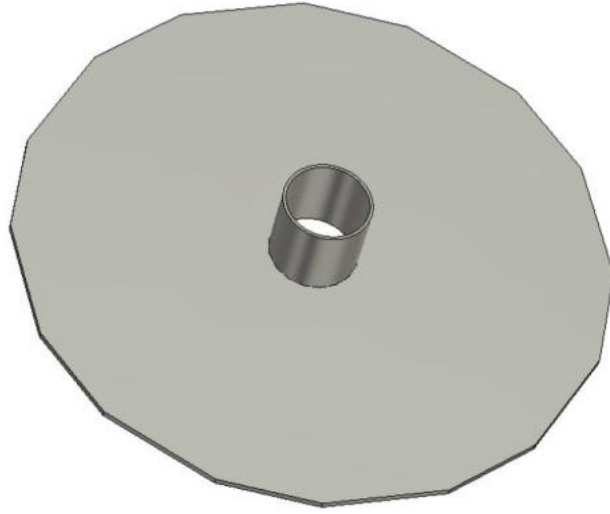


Figure 55 - Top side

The performance data are very interesting. Operating across a broadband 36.5 - 43.5 GHz. The combination loss is remarkably low because the signal division is so quite similar to the theoretical value, as following:

$$S_{n1} = 10 * \log\left(\frac{1}{8}\right) = -9 \text{ dB}$$

where n is the number of output ports.

Furthermore, the reflection coefficient at the output ports remains below -10 dB, a peak resonant value of -40 dB is achieved at 40.8 GHz. It means that the proposed division structure has a very good impedance matching over the bandwidth of interest.

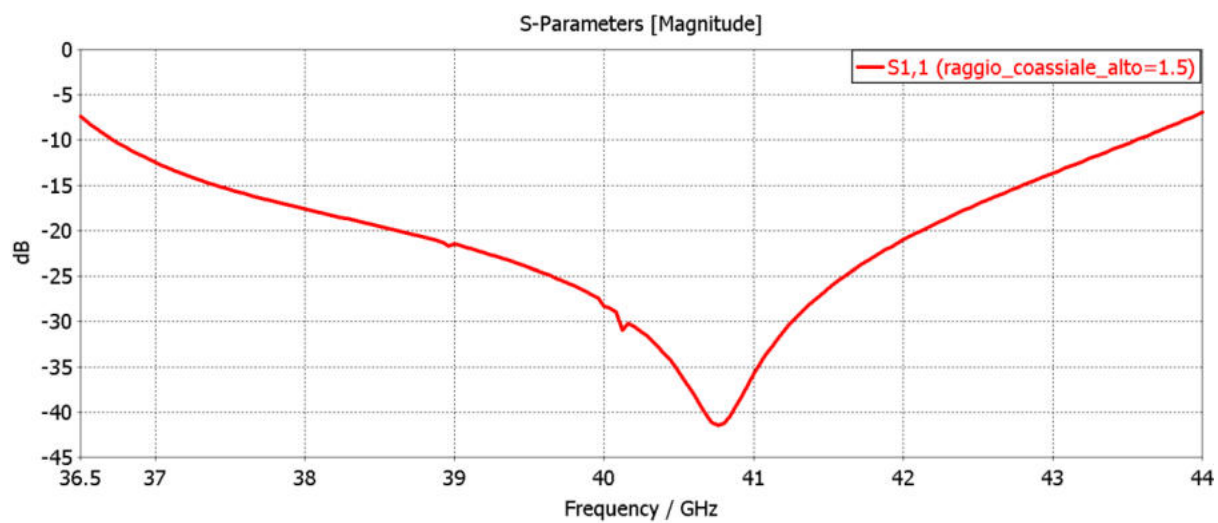


Figure 56 - Return loss

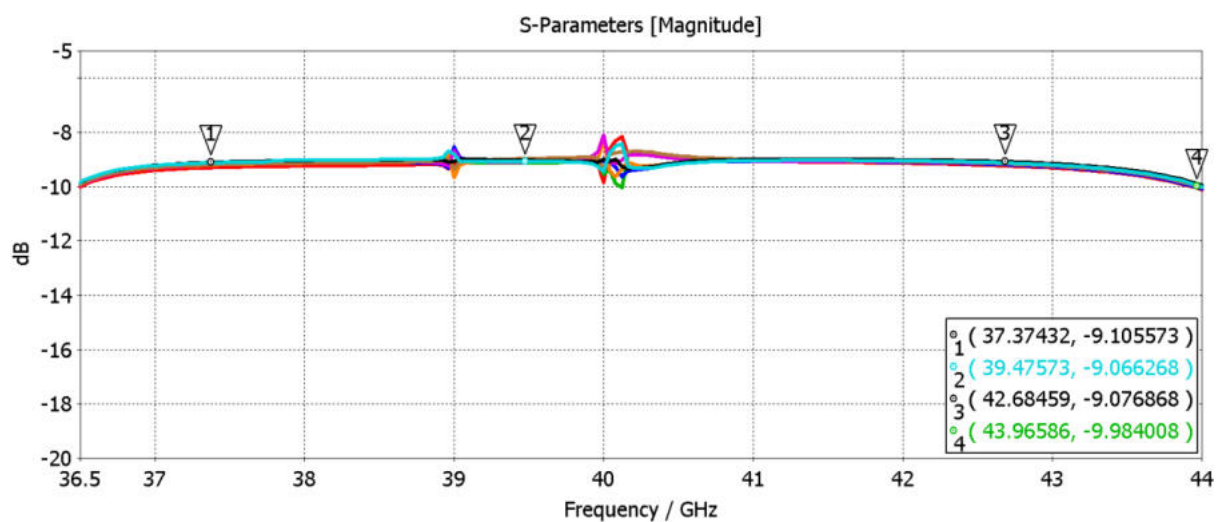


Figure 57 Trasmission signals division

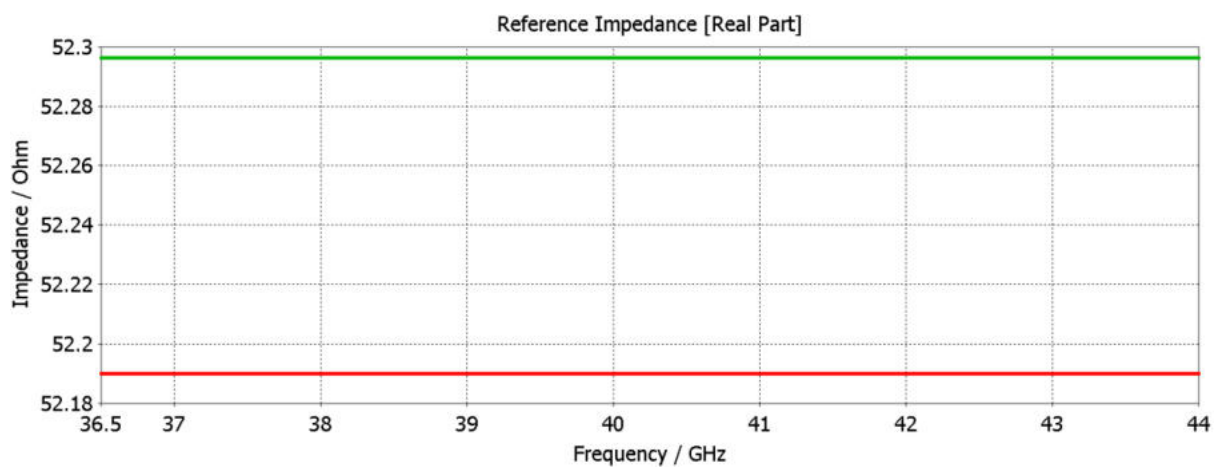


Figure 58 - Input and output impedance ports - real part

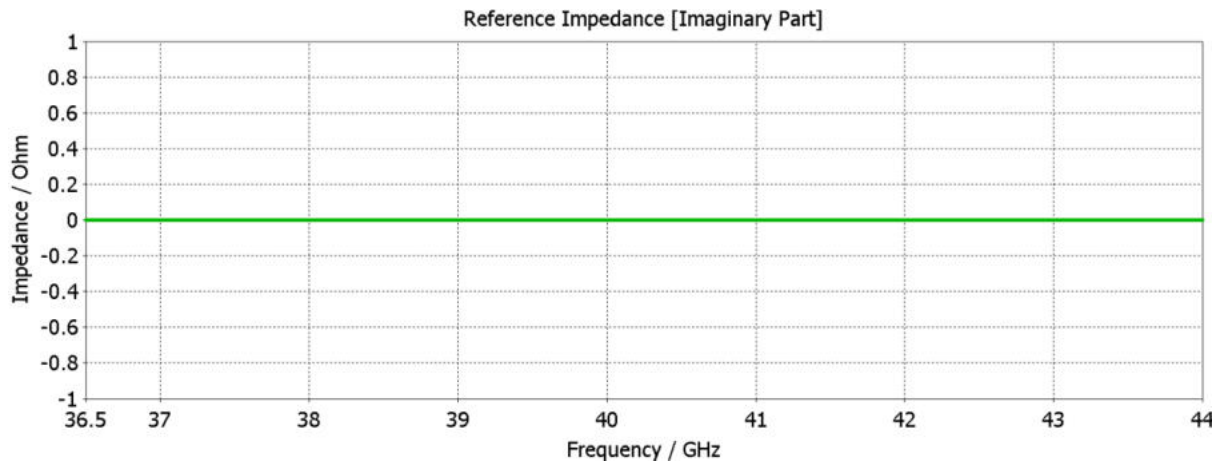


Figure 59 - Input and output impedance ports - imaginary part

Perhaps, the most practical advantage of this radial architecture is the layout itself. By arranging the output ports radially, the perfect symmetry is guaranteed, the EM field distribution gets splitted equally, as shown in Figure 60.

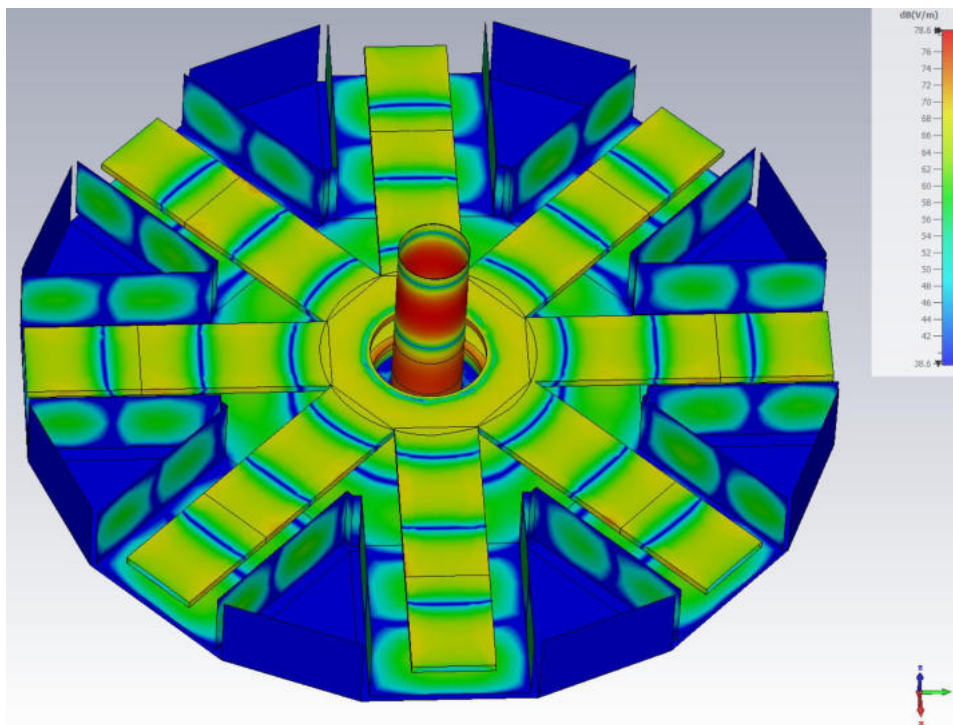


Figure 60 - Propagation of TEM mode in the eight-way stripline combiner

In addition, we can imagine placing mmic in correspondence at each output lines. This configuration proposes a possible solution for a thermal management system.

A significant drawback of the proposed architecture emerges when considering its use in a full power-combining chain — that is, when two divider structures must be integrated back-to-back to form a complete divider/combiner assembly. Because the output ports are arranged radially and oriented perpendicularly to the main propagation axis, recombination requires each output port of the dividing stage to be connected to the corresponding input port of the combining stage. This interconnection must be realized through individual coaxial-to-stripline / stripline-to-coaxial transitions, one per port, effectively introducing an additional transition interface at each branch of the network [100].

The consequences are many. First, each additional transition is a potential source of insertion loss, impedance mismatch, and phase imbalance, all of which degrade combining efficiency in the recombination stage. Second, perhaps more critically from a system integration point of view, the radial arrangement of the ports forces the overall combined structure to expand significantly in the transverse plane.

This limitation suggests that the proposed topology is better suited for use as a standalone power divider.

4 Planar Power Combiner

The proposed 4-way microstrip power combiner is a planar passive model designed to split an input signal into four equal-amplitude, equal-phase output signals over a wide frequency band. The divider is designed on a microstrip substrate — such as Rogers Duroid 4003C laminate, as reported in the Table 6.

Parameter	Value
Material	Rogers RO4003C (lossy)
Type	Normal
Epsilon (ϵ_r)	3.55
Electric $\tan\delta$	0.0027 (const. fit)
Mu (μ_r)	1
Thermal cond.	0.71 W/K·m

Table 6 - Electric parameters

At the input, the divider network presents a bifurcation into two symmetric branches. Each branch consists of two consecutive microstrip sections of different widths, carefully dimensioned to implement the impedance transformation required for wideband matching. This stepped-width approach ensures a smooth transition of the characteristic impedance along each branch, preserving the

integrity of the propagating TEM mode and avoiding spurious reflections at the discontinuities.

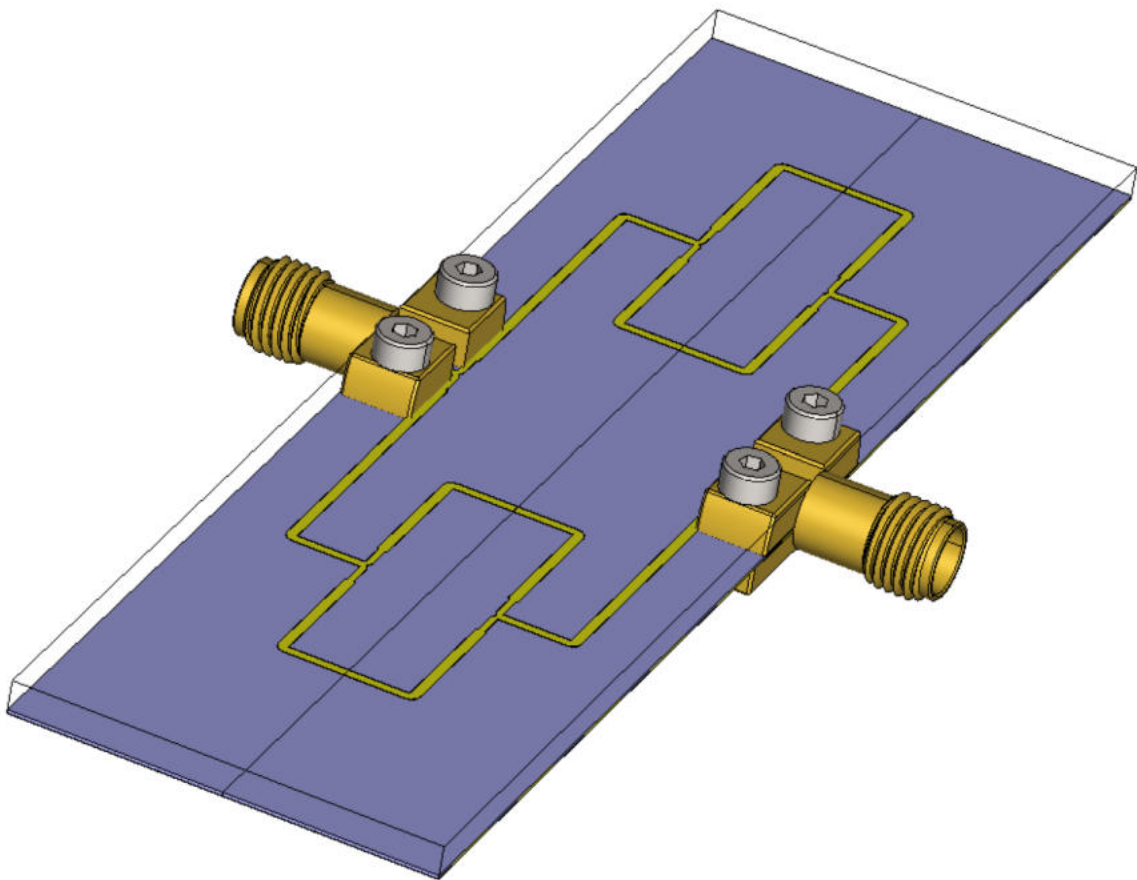


Figure 61 - 4-way power combiner in perspective

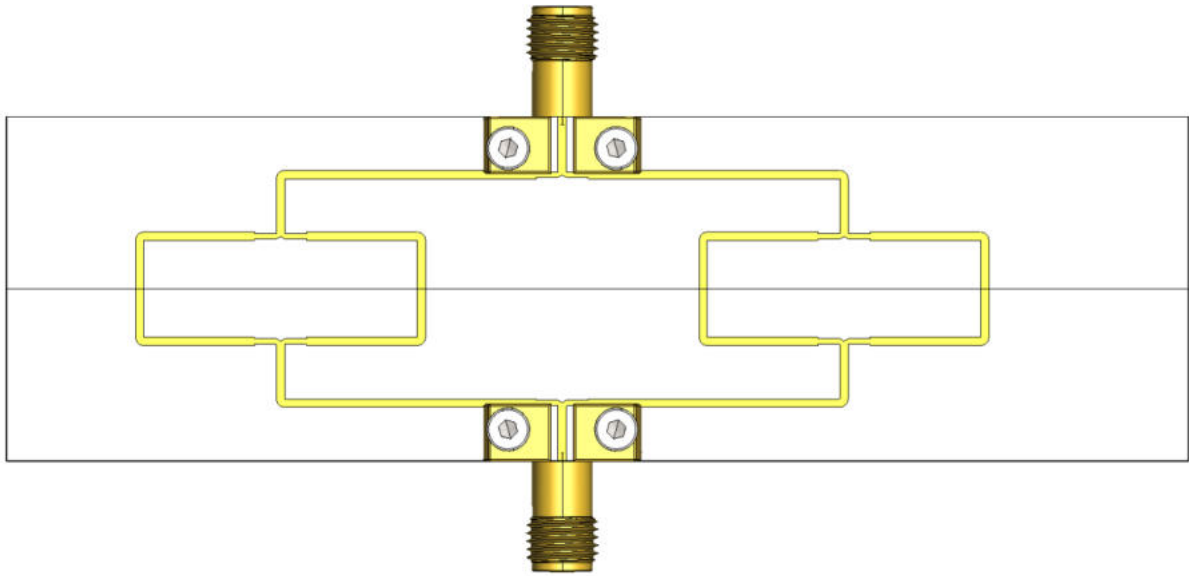


Figure 62 - 4-way power combiner - top view

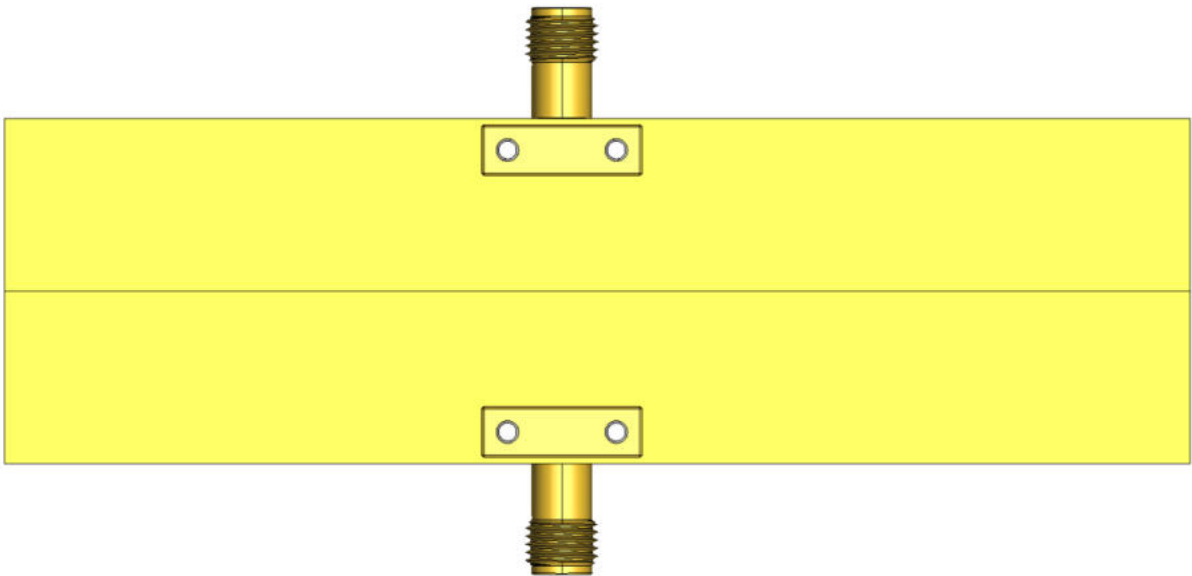


Figure 63 - 4-way power combiner - bottom view

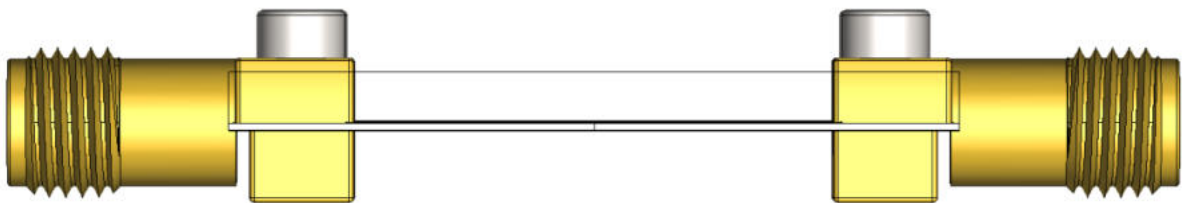


Figure 64 - 4-way power combiner - side view

For RF characterization and performance evaluation, SMA connectors are mounted at the input and output ports. The connectors provide the coaxial-to-microstrip interface necessary for network analyzer measurements.

In Table 7, variables' dimensions are listed:

Variable	Expression	Calculated Value (mm)	Description
r_conduttore_interno_1	1.3/2	0.65	Inner conductor radius 1
L_conduttore_interno_1	15	15	Inner conductor length 1
L_dielettrico_1	15	15	Dielectric length 1
r_dielettrico_1	3.9/2	1.95	Dielectric radius 1
r_gnd_1	5.5	5.5	Ground radius 1
L_gnd_1	4	4	Ground length 1
r_pin	0.15	0.15	Pin radius
H_block	15	15	Block height
L_pin	8	8	Pin length
epsilon	1	1	Relative permittivity (ϵ_r)

Table 7 - Variables of the Planar PC

Figure 65 shows the simulated input and output return loss of a planar microstrip power combiner operating in the frequency range from 17 to 21 GHz. The two

curves overlap almost perfectly across the entire band, confirming the structural symmetry of the combiner network between input and output ports.

The combiner exhibits a dual-resonance behavior. At 18.40 GHz, S11 reaches its deepest value of -33.66 dB, representing the point of optimal impedance matching. Near 19.22 GHz, a second minimum of approximately -30 dB is observed, extending the well-matched region.

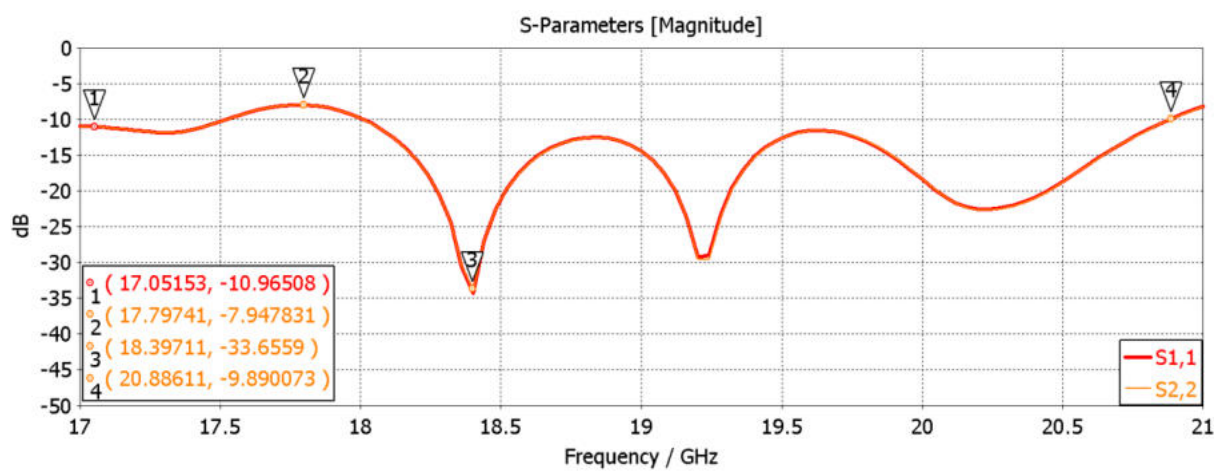


Figure 65 - Return loss

This dual-resonance profile is a direct consequence of the two-section stepped-width microstrip branches at the input bifurcation, combined with the septum inserted at the junction, which get better the return loss response around the design frequency.

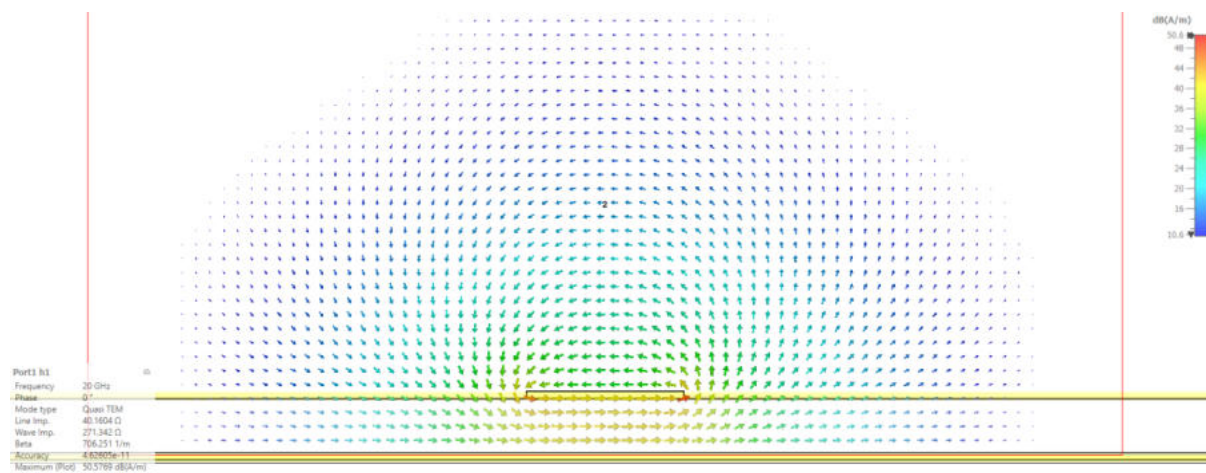


Figure 68 - H field distribution - vectorial representation

5 Radial Coaxial Power Divider

This section introduces an advanced 8-way Ka-band radial power combiner designed for getting low insertion loss and wideband performance at millimeter wave frequencies, addressing several advantages in high-power solid-state amplification systems [101]. The model is shown in Figures below (top view and perspective view).

Parameter	Value
Central pin diameter	1.27 mm
Outer pin diameter	2.92 mm
Total dimensions	22.80 x 22.90 mm

Table 8 - Assembly drawing size

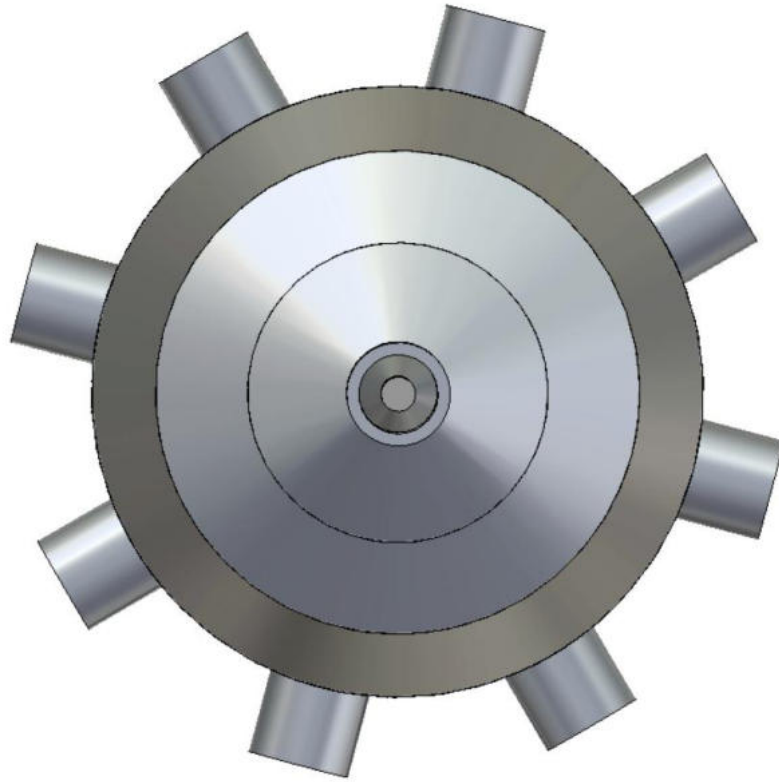


Figure 69 - Radial Coaxial Power Divider - top view

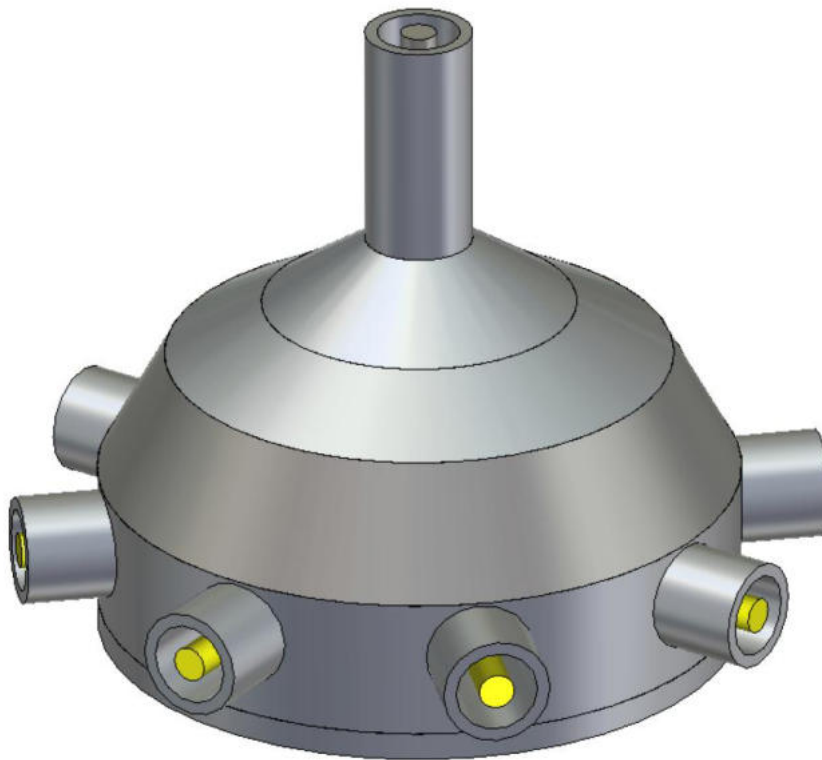


Figure 70 - Radial Coaxial Power Divider - Perspective view

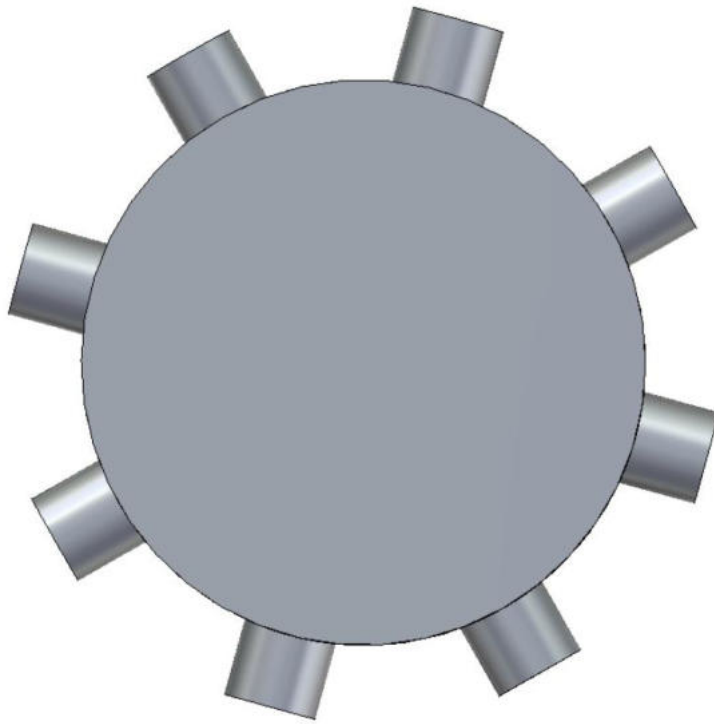


Figure 71 - Radial Coaxial Power Divider - bottom view

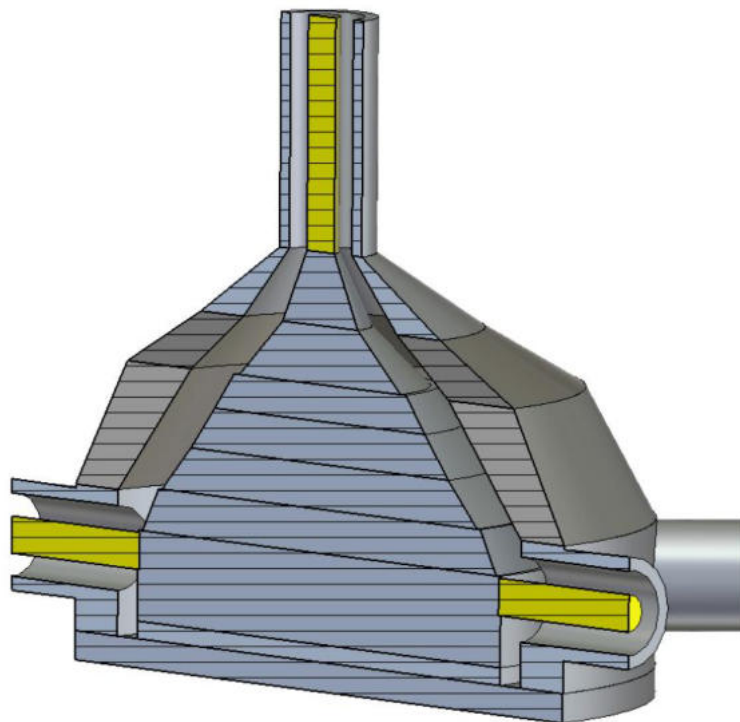


Figure 72 - Radial Coaxial Power Divider - cross section

5.1 Electrical Performance

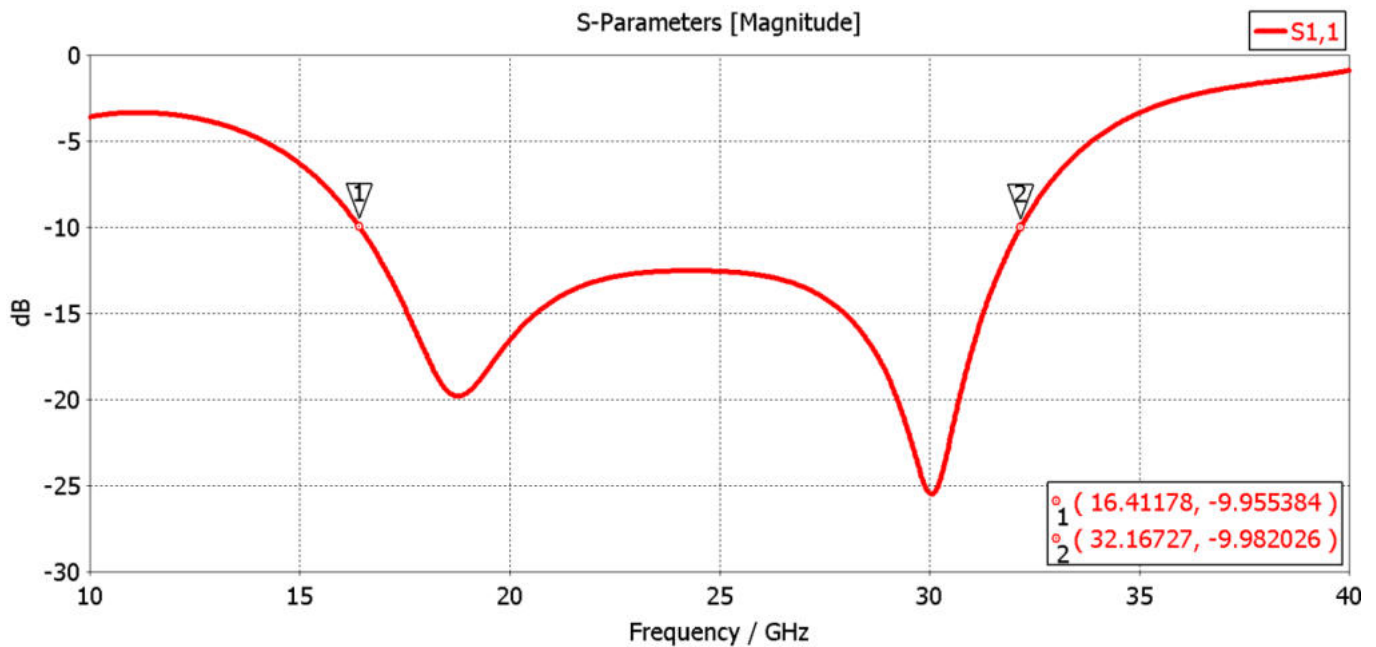


Figure 73 - Return loss

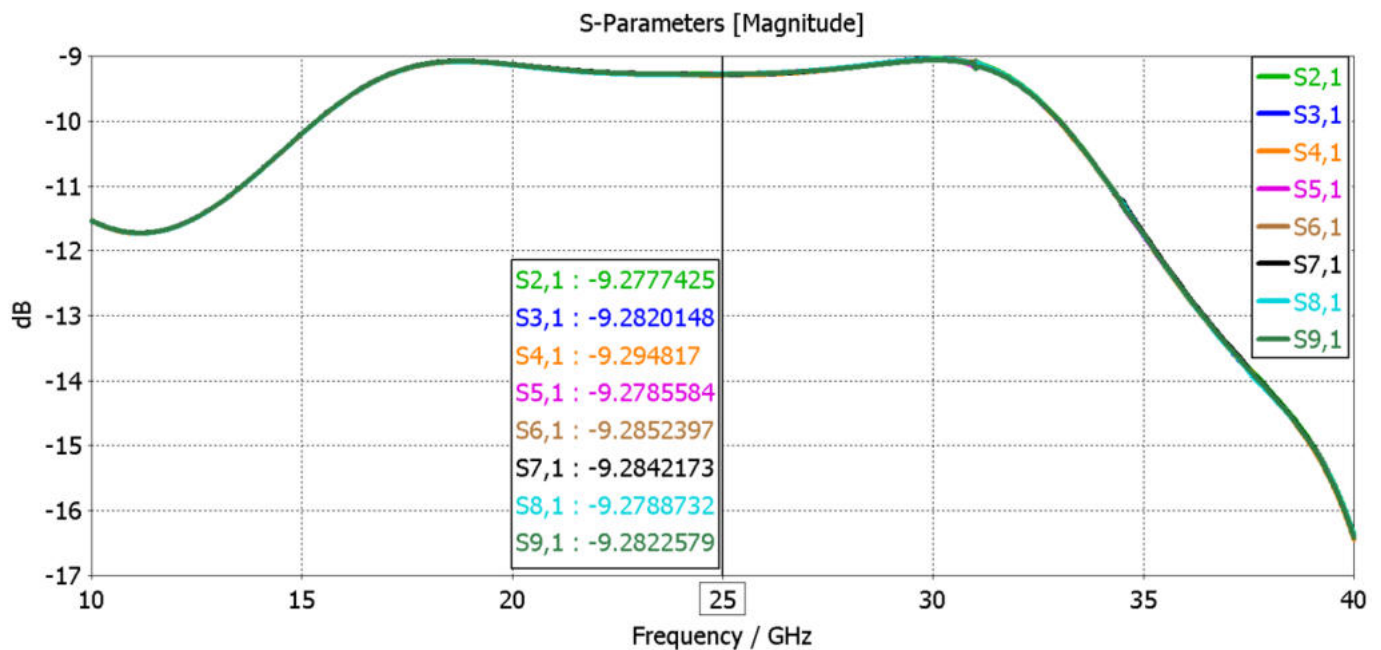


Figure 74 - Transmission output Power division

5.1.1 Zoomed plots over the bandwidth of interest

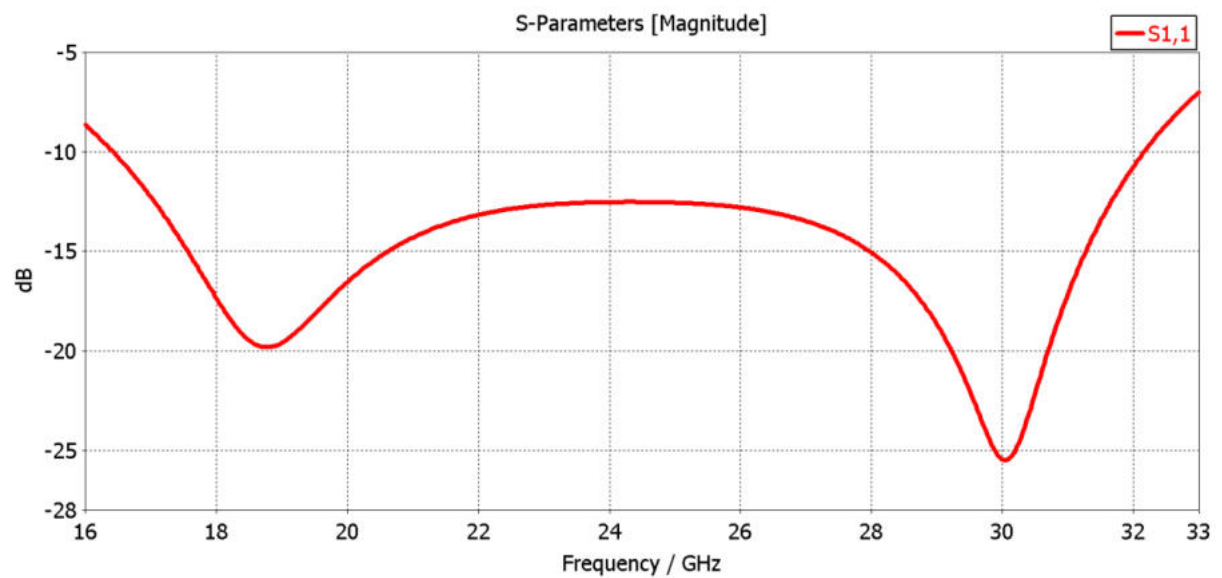


Figure 75 - Return loss – Narrower bandwidth representation

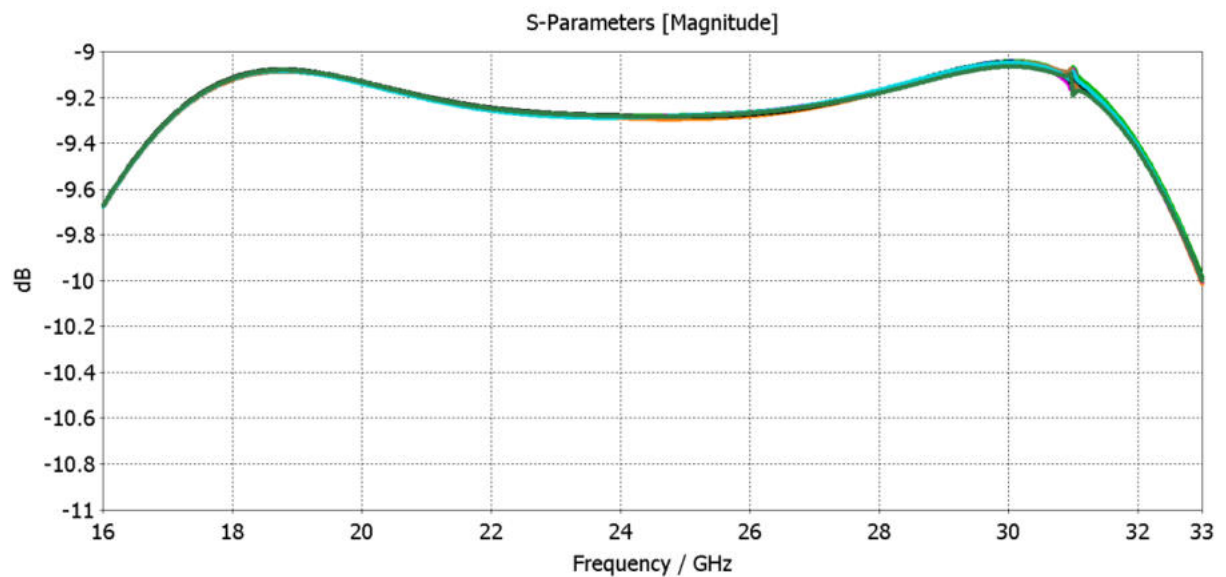


Figure 76 - Transmission output Amplitude Power division - Narrower bandwidth representation

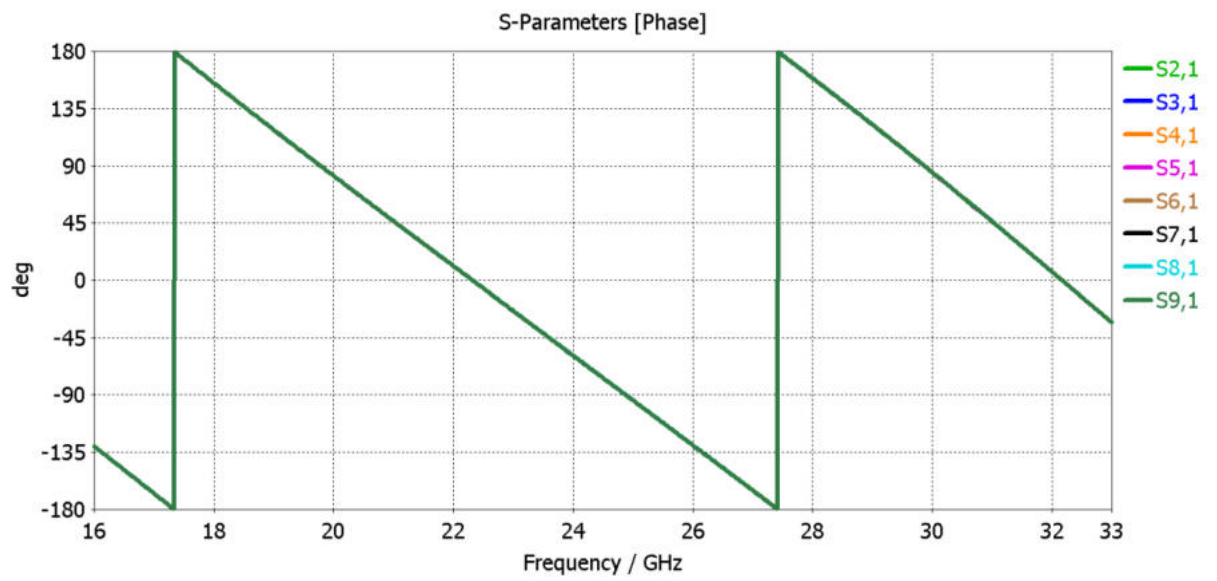


Figure 77 - Transmission output Phase Power division - Narrower bandwidth representation

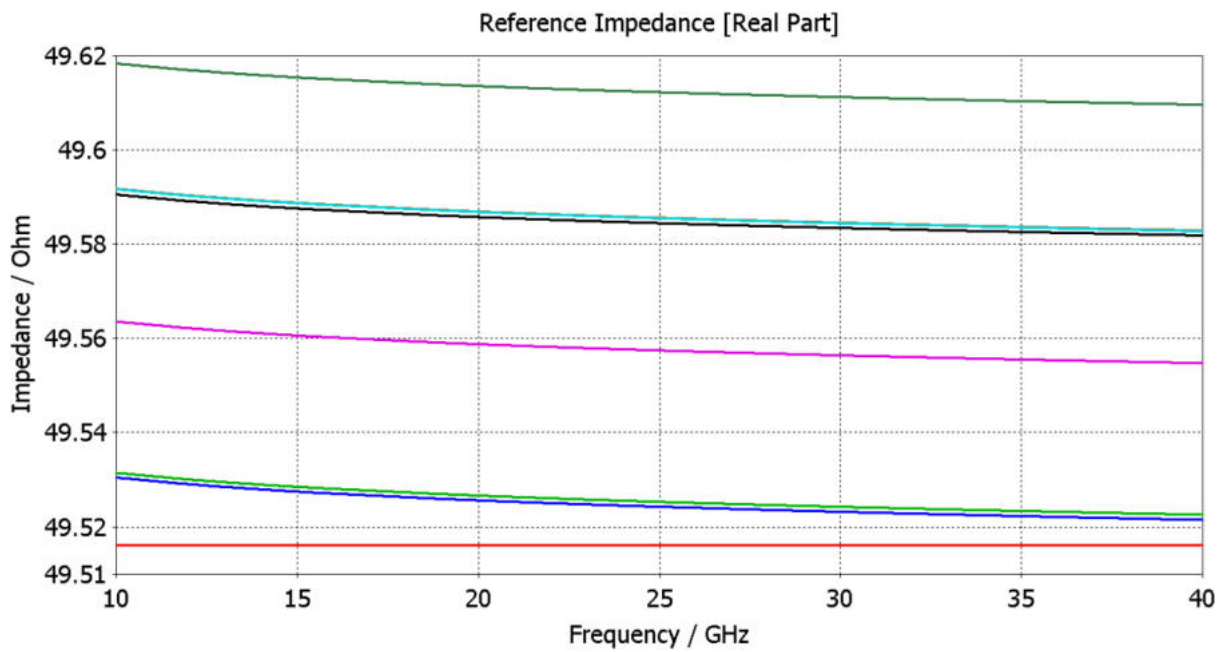


Figure 78 - Input and output impedance ports - real part

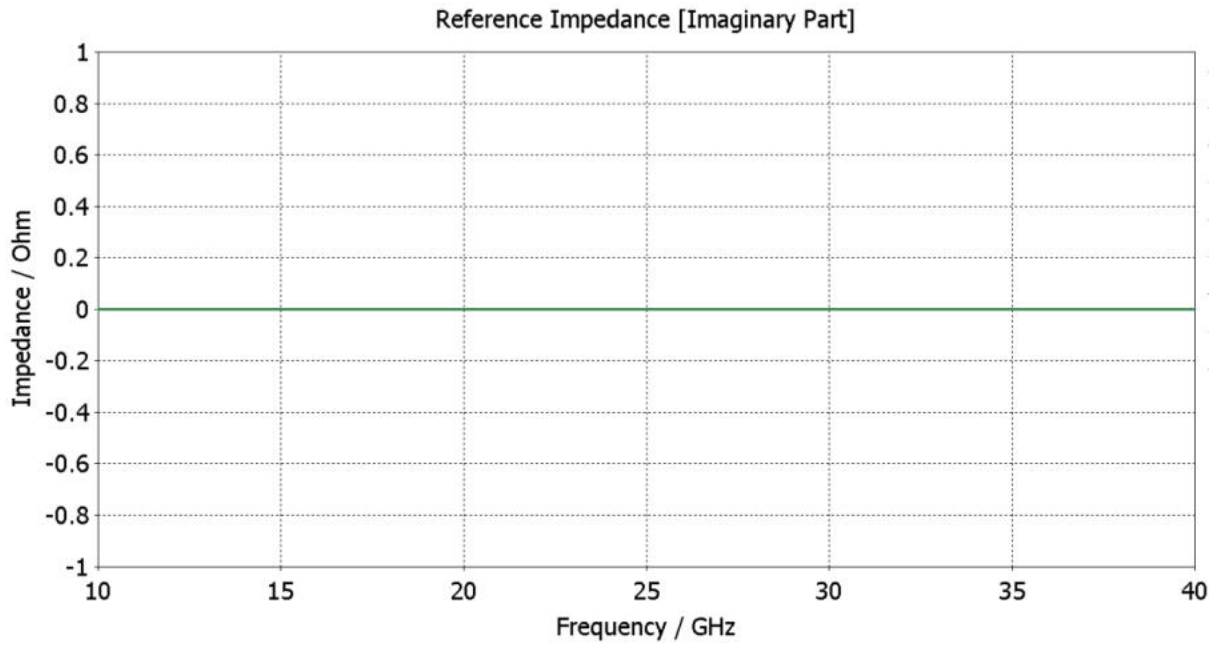


Figure 79 - Input and output impedance ports - imaginary part

5.2 Prototype Idea:

The prototype architecture was optimized for making fabrication feasibility and cost minimization as simply as possible, ensuring negligible perturbation to RF performance metrics such as insertion loss and return loss. The 3D solid model was decomposed into three stand-alone sub-blocks that are assembled like “legos” blocks. Renderings of the integrated mechanical configuration are provided in orthographic and isometric projections: plan view (top), perspective view, and port-side elevations (left/right), as shown in Figure 80 and Figure 81.

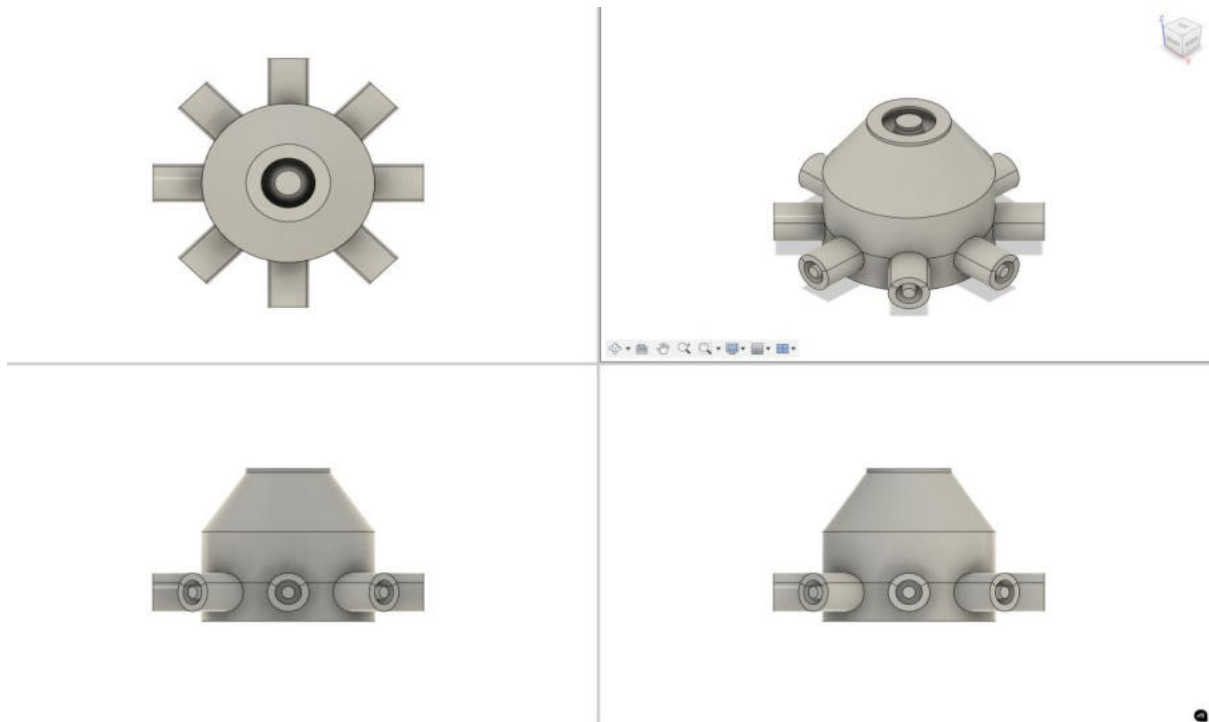


Figure 80 - Prototype Side views



Figure 81 - Prototype Side views in transparency

The CAD assembly model of the prototype, developed using Autodesk Fusion 360 mechanical design software, is presented in Figure 82.

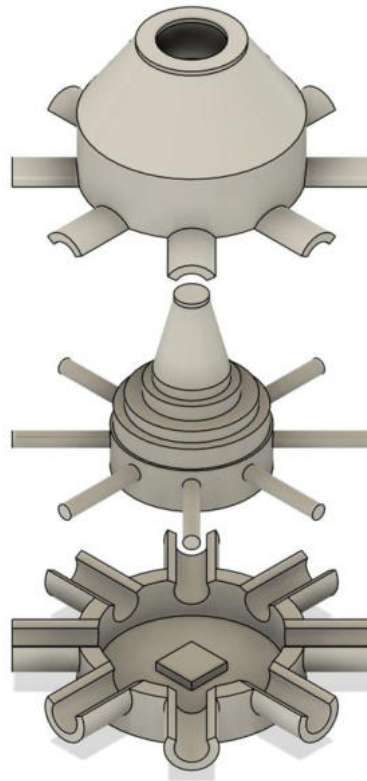


Figure 82 - Assembly CAD model

The central block of the model was dimensioned as follows (see Figure 83):

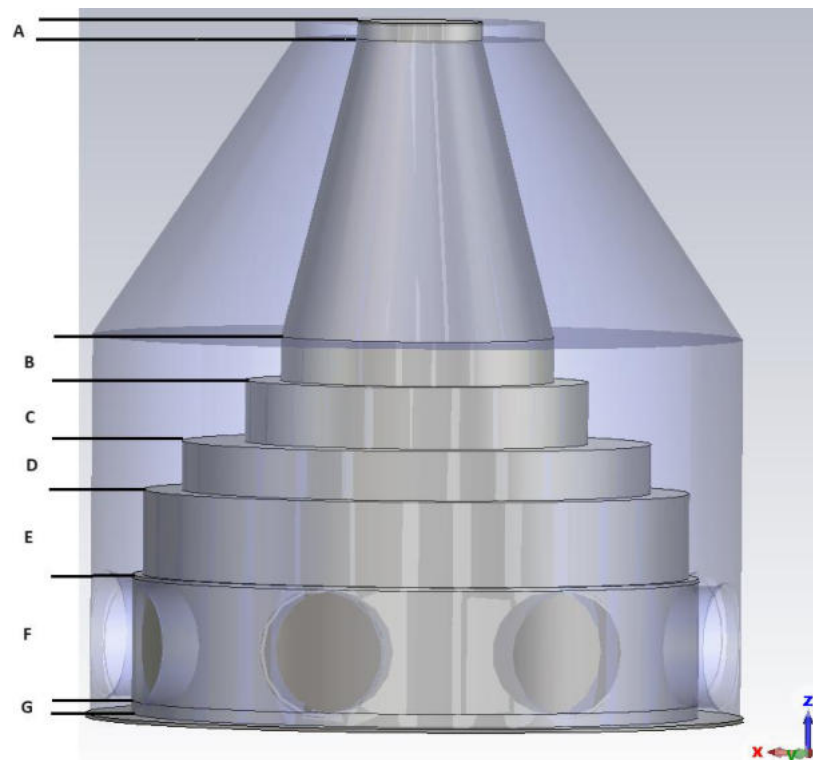


Figure 83 - Coaxial Spatial combiner with different section labelled

Section	Diameter [mm]	Height [mm]	Impedence [Ω]
A	Inner: 3.58; Outer: 7.8	9	50
B	7.8	1.26	50
C	9.8	1.75	36.45
D	13.4	1.4	17.69
E	15.6	2.42	8.58
F	16.2	3.5	6.31
G	16.2	0.37	6.31

Table 9 - Dimensions of the stepped inner coaxial structure

In order to feed the radial line, it's required a field distribution that ensures an axial symmetry, which is guaranteed by a coaxial TEM mode. Power is divided into eight ways using eight equi-spaced identical radial coaxial probes which are connected to the oversized central coaxial waveguide and are designed to have an impedance of 43Ω each. To gain insight into theory of the proposed combiner structure and its operation, We have presented an approximate equivalent circuit model for this structure, as shown in Figure 83.

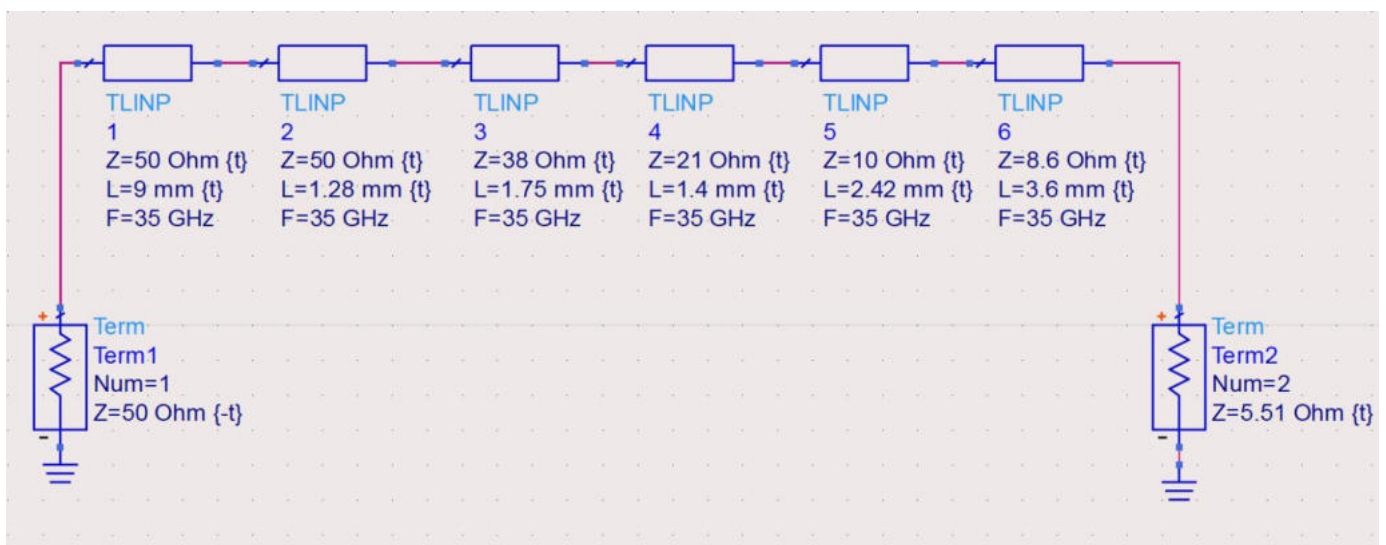


Figure 84 - Equivalent circuit model

As can be seen in the model above, the characteristic impedance Z_0 varies strongly as the function of the changing diameter value according to the Table 9. We propose a new tapered coaxial feeder design with the aim of making an improvement of the coaxial 's reflection coefficient starting from the N-section quarter-wave impedance transformer theory. At the first junction, the reflection coefficient is:

$$\Gamma_0 = \frac{Z_1 - Z_0}{Z_1 + Z_0}$$

In general, at the n-th junction, the reflection coefficient is:

$$\Gamma_n = \frac{Z_{n+1} - Z_n}{Z_{n+1} + Z_n}$$

The last reflection coefficient is

$$\Gamma_N = \frac{Z_L - Z_N}{Z_L + Z_N}$$

In order to achieve the best reflection coefficient with multi section impedance matching, some optimization was performed with ADS software based on the initial values of impedance lines and height of sections.

The prototype exhibits exceptional compactness and was fabricated using pink-colored ABS (acrylonitrile-butadiene-styrene) thermoplastic filament via additive manufacturing on a 3D printer.

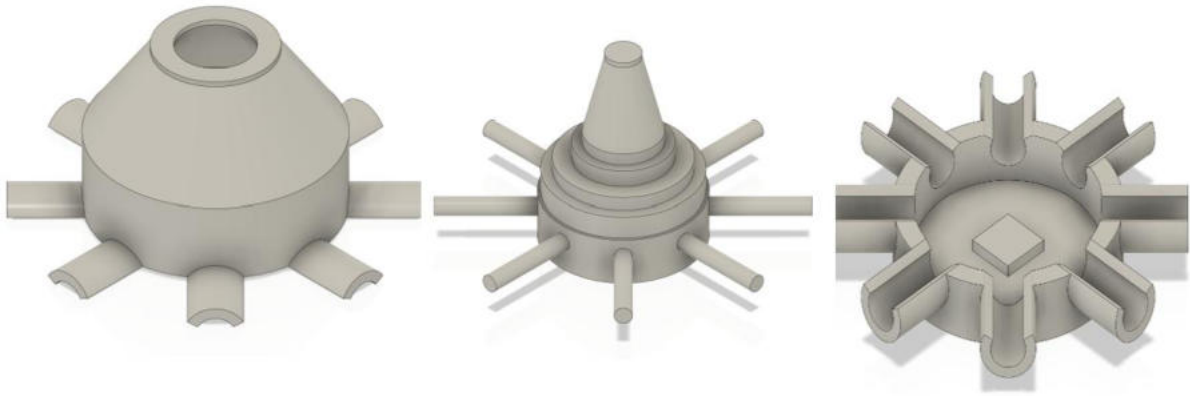


Figure 85 - Prototype sub-blocks (1)



Figure 86 - Prototype sub-blocks (2)

As shown in the image above, the bottom subassembly incorporates a square-section alignment septum that gets inserted into the precision pocket machined at the bottom side in Part 2, ensuring perfect concentric alignment of the eight central rods representing the output coaxial inner conductors.

5.2.1 Fabricated Radial Power Combiner

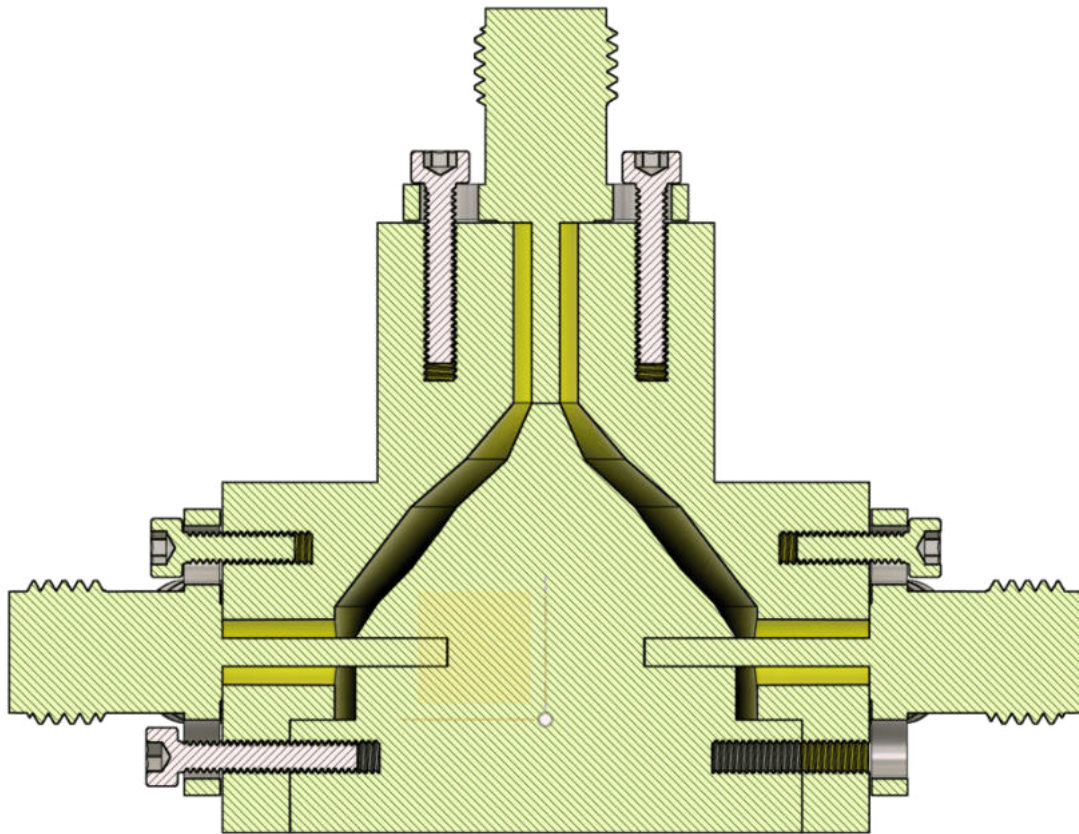
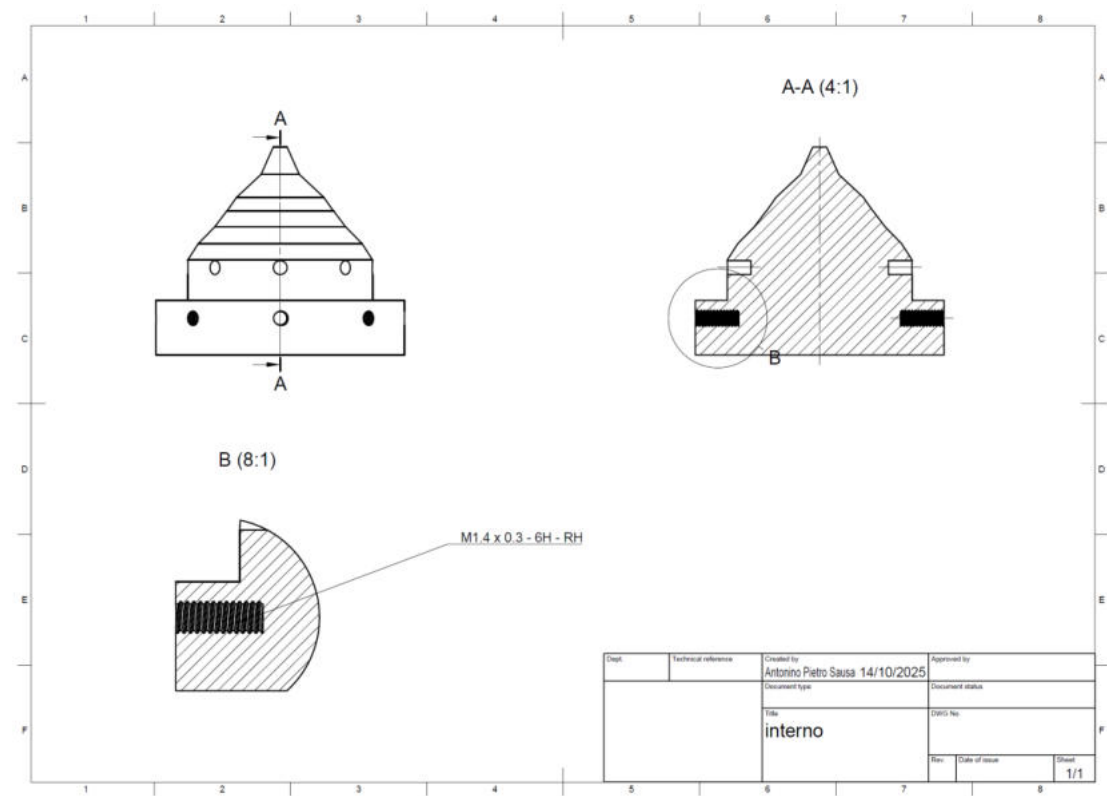
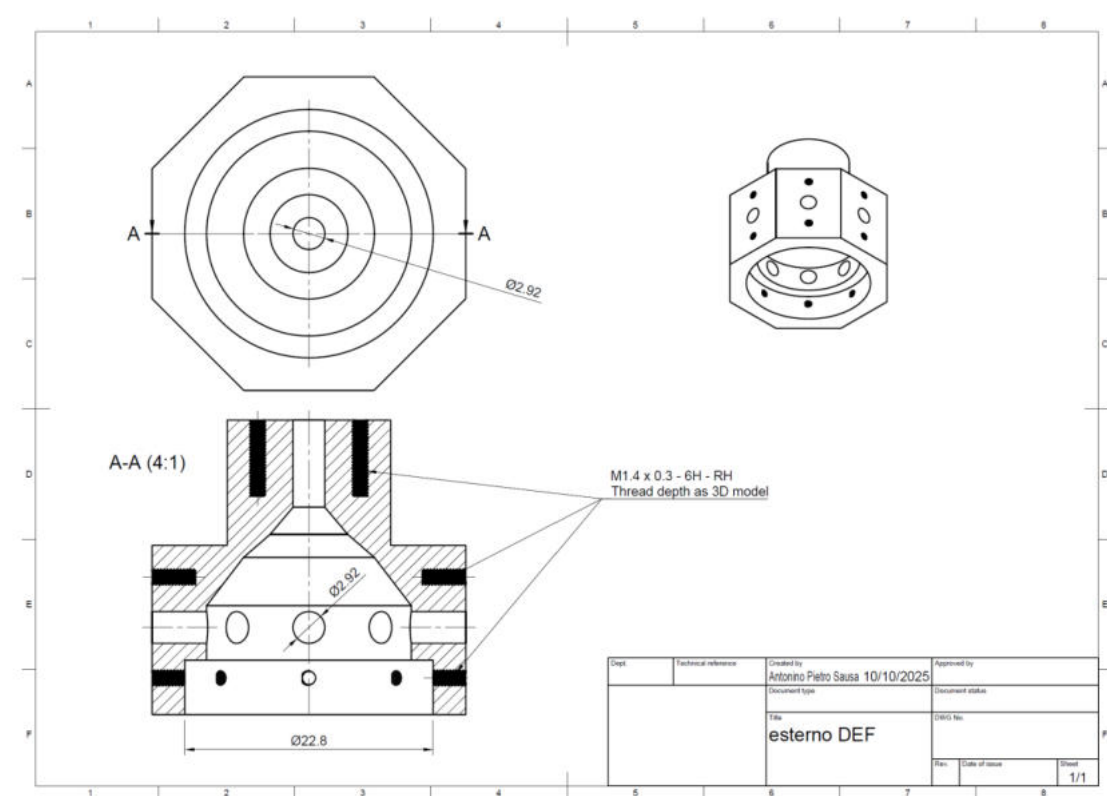


Figure 87 - Power divider cross-section

● Inner part



● Outer part



Connector: MicroRFConnector:

2.92mm Female Airline Bulkhead Connector 12.7 x 4.8mm / 0.500 x 0.189inch

Flange Jack with $\Phi 1.27\text{mm}$ / .050" Pin

DATA SHEET

2.92mm Female Airline Bulkhead Connector

12.7 x 4.8mm / 0.500 x 0.189inch Flange Jack

with $\Phi 1.27\text{mm}$ / .050" Pin

3. KEY FEATURES

Item	Description
Polarity	Standard
Interface	2.92mm
Connector Type	Jack
Connector Gender	Female
Impedance	50 Ω
Body Shape	Straight
Body Mounting	2 Hole Flange
Attachment Method	Solder

4. PERFORMANCE FEATURES

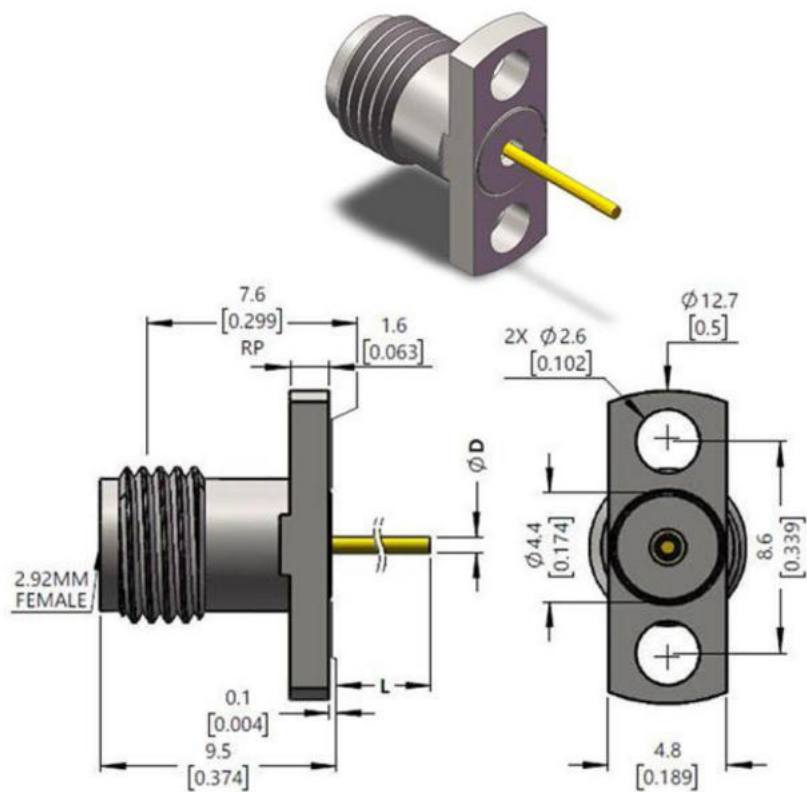
Item	Description
Frequency Range	DC~40GHz
VSWR	DC~40GHz: 1.15:1 (Max)
Contact Resistance	Center Conductor $\leq 3.0\text{m}\Omega$, Outer Conductor $\leq 2.0\text{m}\Omega$
Insulation Resistance	$\geq 5000\text{m}\Omega$
Withstanding Voltage	750V
Durability	500 Cycles
Operating Temperature	-55°C ~ +165°C

6. ADDITIONAL SIZE INFORMATION

Item	Description
Diameter of Pin (refer to the below drawing)	ΦD : 1.27mm / 0.050"
Length of Pin (refer to the below drawing)	L: 8mm / 0.314"

7. DRAWING

Drawing (unit: mm)



5.2.2 Experimental setup and measured results

In this section the fabricated model and the experimental setup are shown.



Figure 88 - Fabricated Radial PC (1)



Figure 89 - Fabricated Radial PC (2)

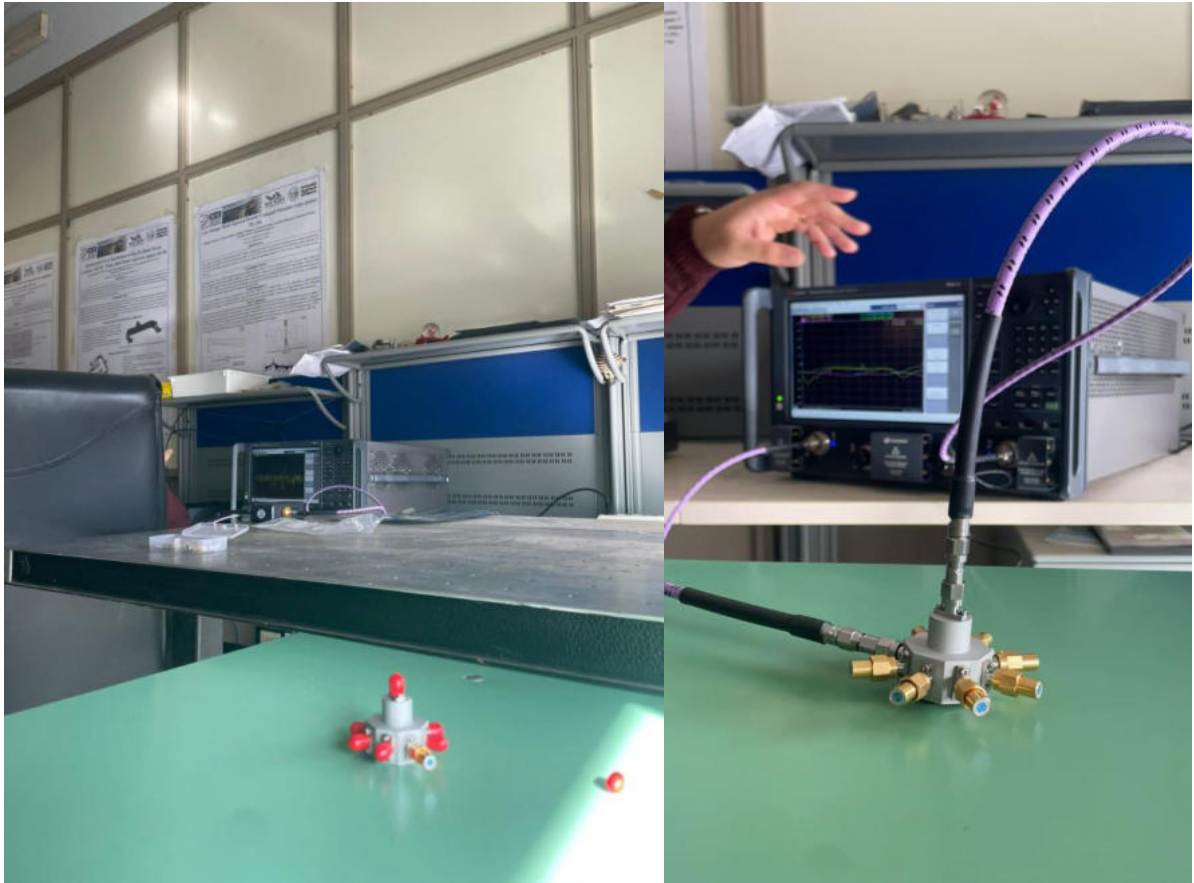


Figure 90 - Figure 89 - Experimental setup (VNA - Constant phase cables - DUT)

Measurements were carried out by using the Keysight PNA-X N5247B vector network analyzer (10 MHz–67 GHz bandwidth). The device under test (DUT) was connected to the analyzer via phase-stable coaxial cables. A full two-port calibration was first performed using a calibration kit (Cal Kit), ensuring the reference plane coincides with the terminal parts of the phase stable coaxial cables. Because of the DUT is a 9-port architecture (1 input port + 8 output ports), measurements proceeded step by step. Port 1 of the analyzer connected to the DUT input, while Port 2 connected to one output port at a time. All remaining output ports were terminated with matched 50 Ω loads to maintain proper even-

mode excitation and prevent odd-mode reflections. The measurement results obtained are presented in Figure 91.

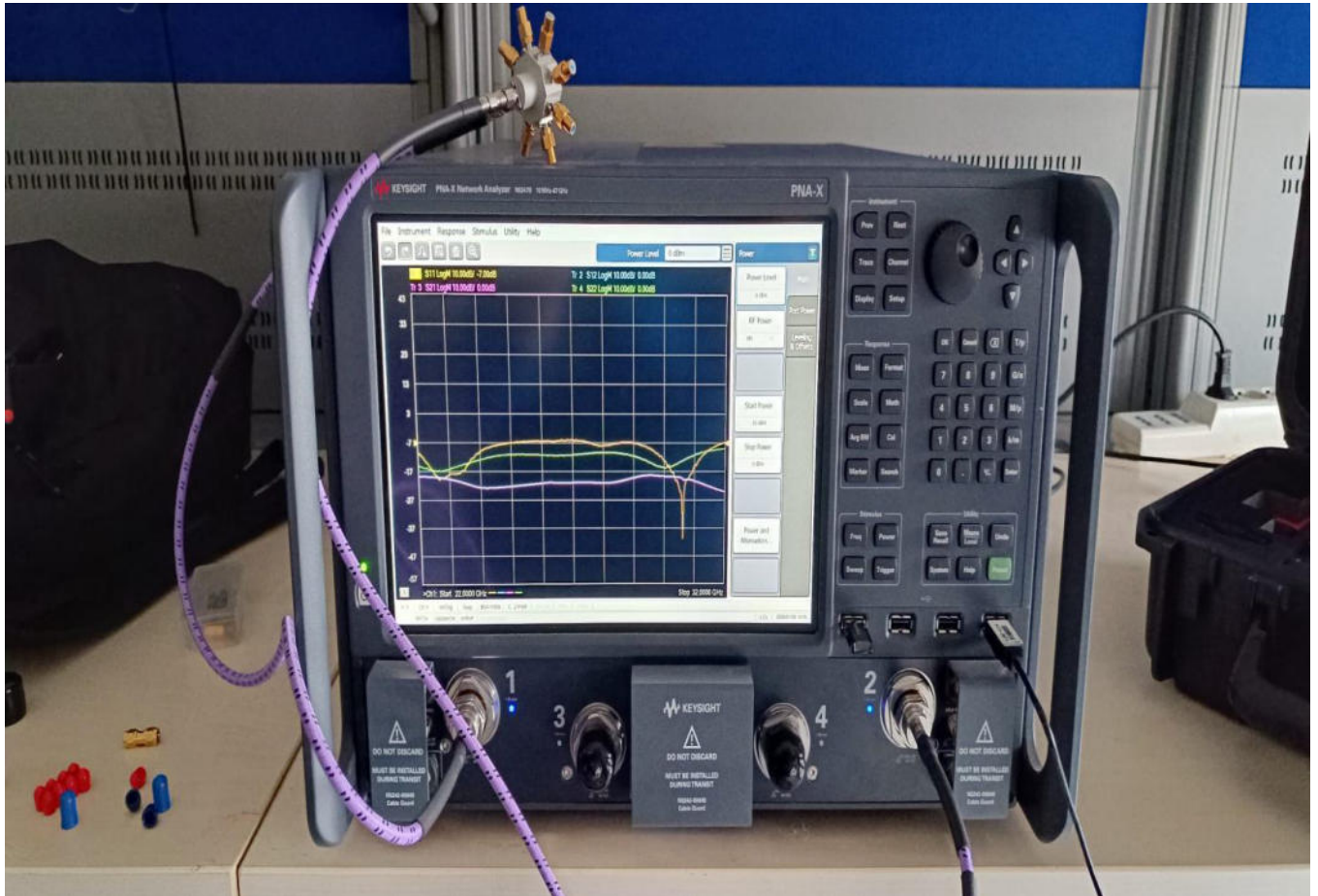


Figure 91 - Keysight PNA-X N5247B vector network analyzer

The measured return loss curves are presented. They show a very similar behavior across the band of interest, confirming the expected symmetry of the dividing structure [102]. In particular, excellent impedance matching is observed at the upper band frequency, around 30–31 GHz, where the return loss reaches approximately 25 dB. No further resonance peak appears over the frequency band in which the return loss achieves about 16 dB.

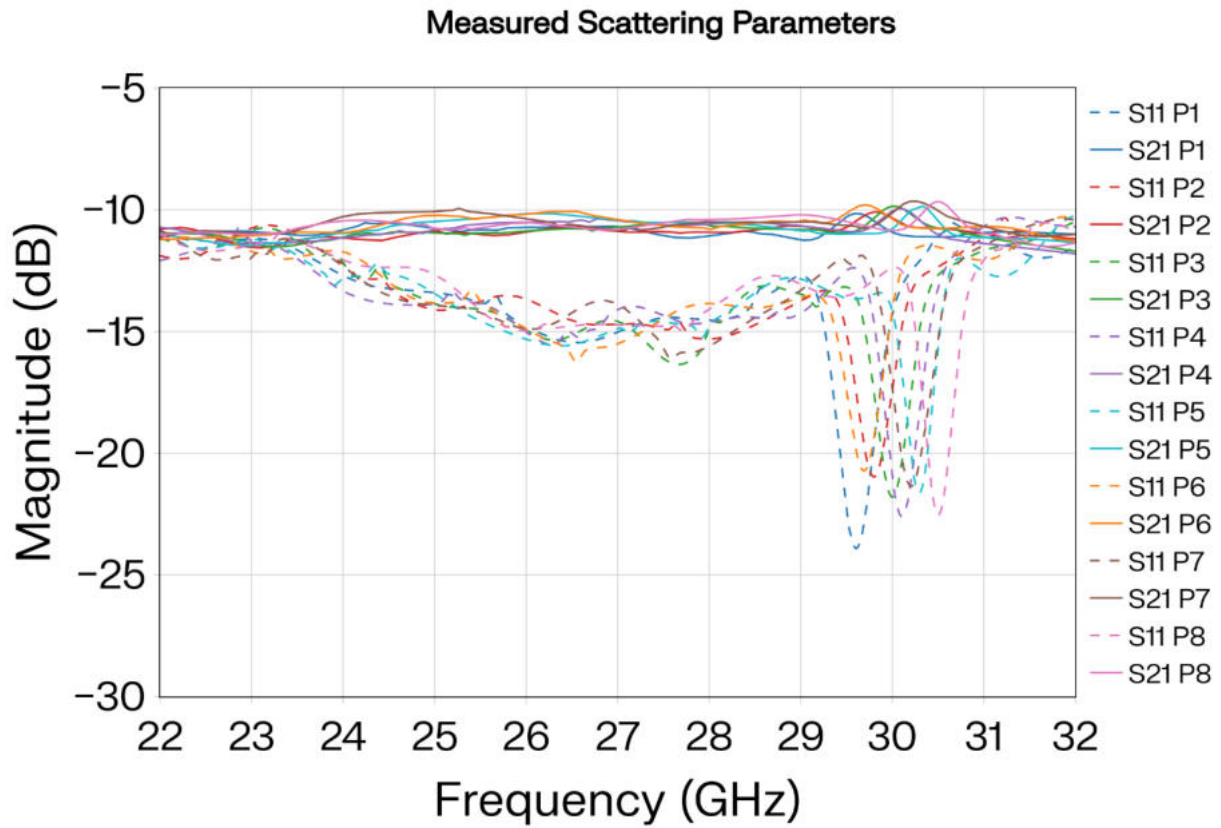


Figure 92 - S11 Magnitude curves

Meanwhile, the corner region is the most critical and least matched, with a return loss of roughly 11 dB. Several mechanisms can contribute to this degradation. First of all, the similarity of all measured curves supports the effectiveness of the symmetric divider topology. The mechanical parts are manufactured in aluminum: the drilling process may introduce surface roughness and geometric deviations that distort the integrity of the fundamental propagating mode, such as the intrinsic tolerances of the machining process.

Another possible source of performance degradation is the set of $50\ \Omega$ matched loads connected to the unused ports, which may introduce a small residual mismatch between ports and thus impact the overall return loss.

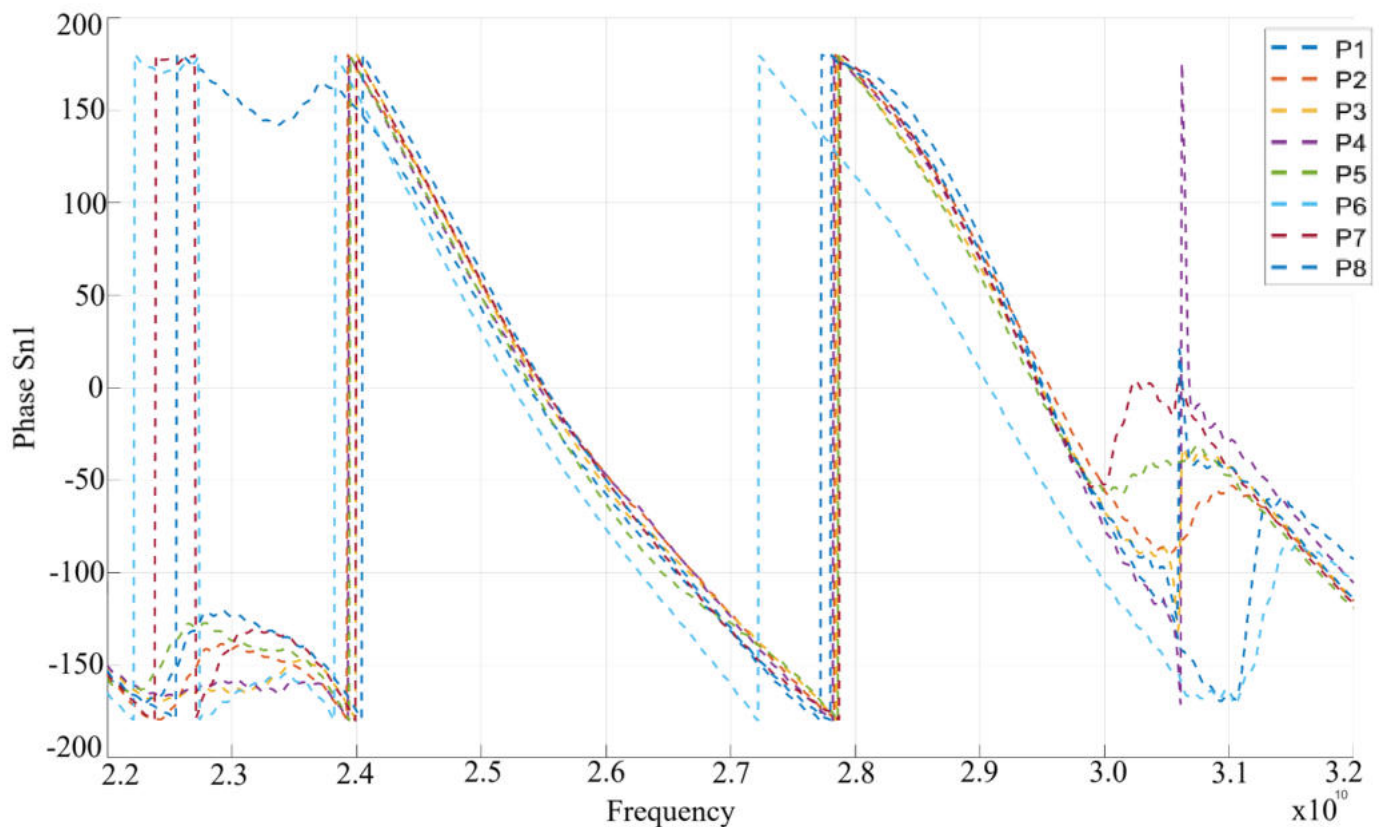


Figure 93 – S₁₁ Phase curves

Finally, the measured S_{11} phase responses are consistent with the simulated trends, further validating the accuracy of the electromagnetic model in predicting the combiner behavior. Indeed, the insertion loss curves exhibit correspondingly lower values at frequencies where impedance matching is optimal. Compared to the simulated -9 dB ideal target, measured values reach -10/-11 dB over the band.

The corner-band region remains the most critical, in accordance with the return loss behaviour observed before.

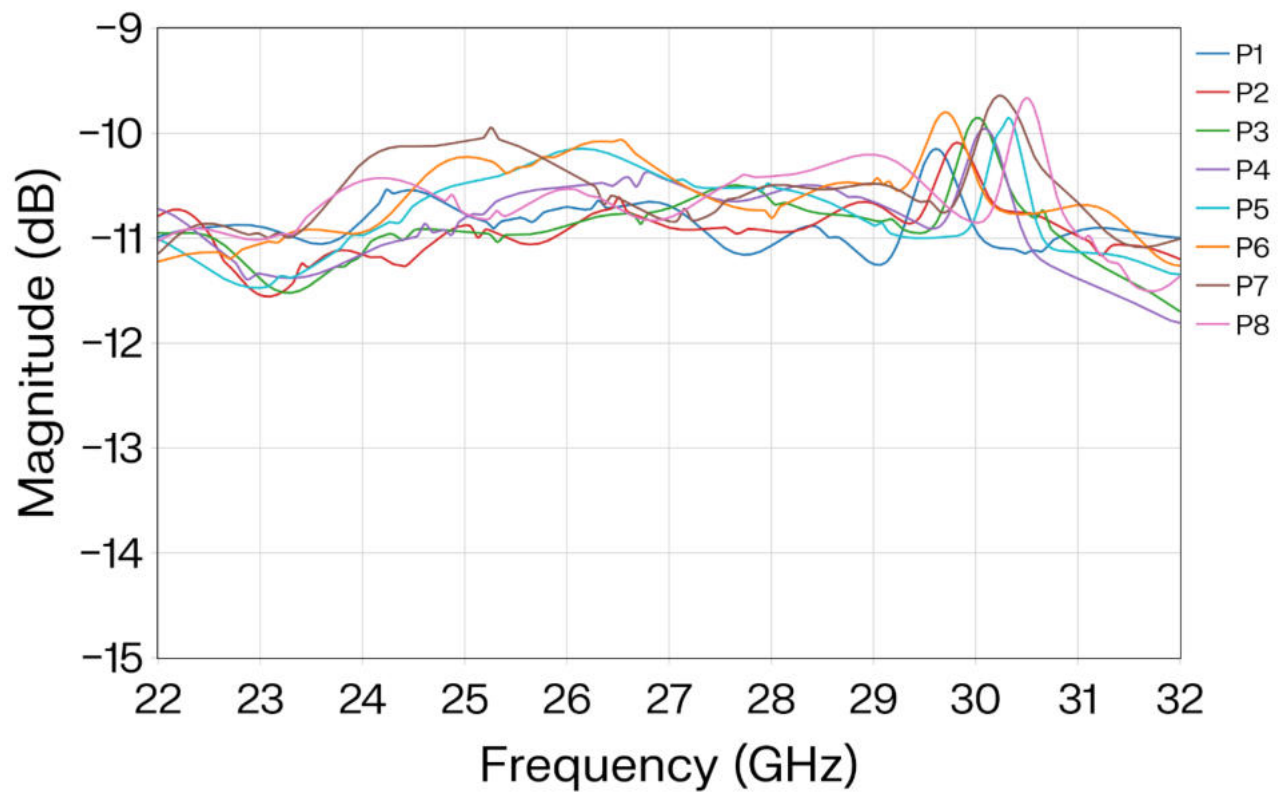


Figure 94 - Sn1 Magnitude parameters

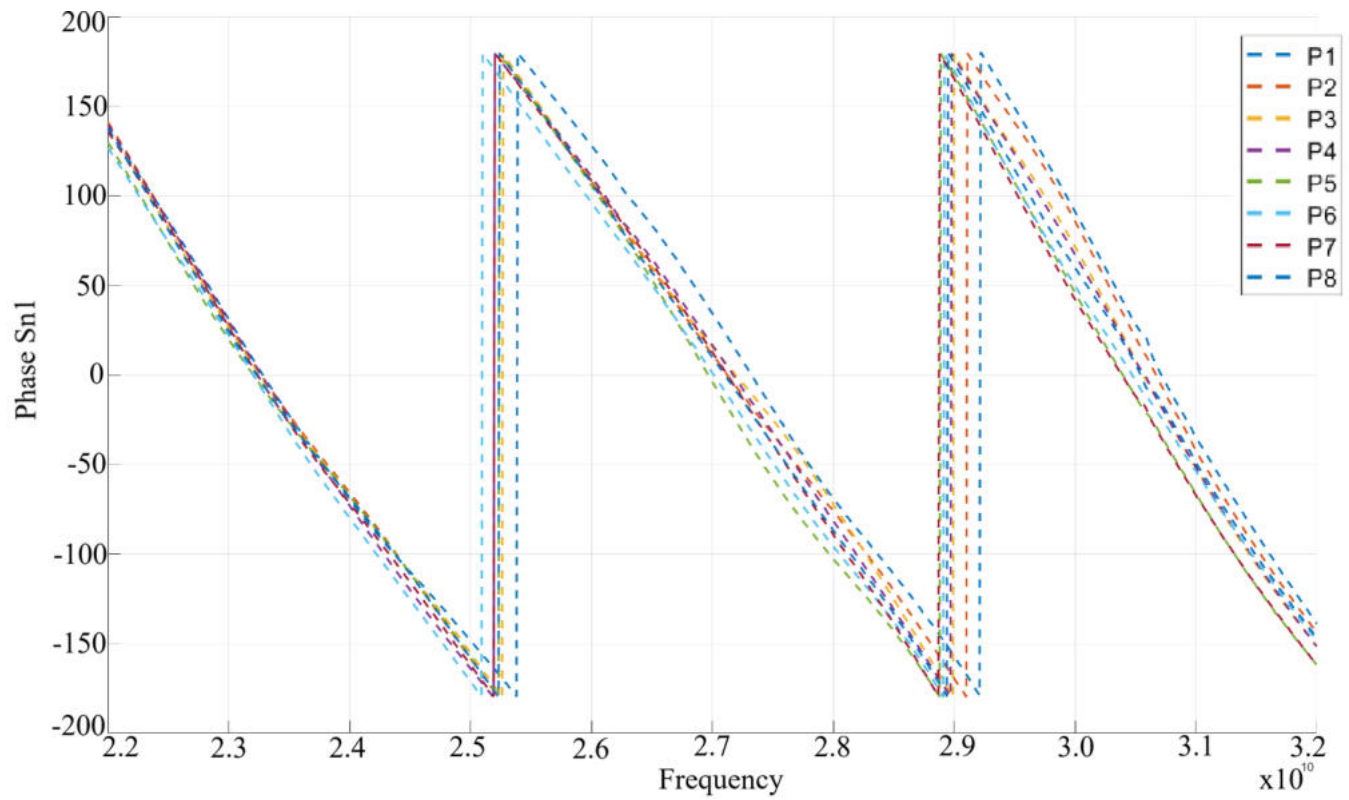


Figure 95 - Sn1 Phase parameters

S-parameter data for individual channels were extracted from measurements in .s2p format. The complete S-parameter matrix for each frequency point is listed below.

Frequency (Hz)	S11 Mag (dB)	S11 Phase (°)	S21 Mag (dB)	S21 Phase (°)	S12 Mag (dB)	S12 Phase (°)	S22 Mag (dB)	S22 Phase (°)
22000000000	-13,906692	-157,54276	-9,5483513	129,36539	-9,5198679	129,43065	-9,5166531	-40,318825
22100000000	-15,795349	-167,20479	-9,5504122	118,44614	-9,5181723	118,55487	-9,9464178	-45,839417
22200000000	-17,734171	-173,69649	-9,5592537	107,64218	-9,5278988	107,72329	-10,393826	-50,949329
22300000000	-20,308243	-175,49802	-9,5681677	96,890396	-9,5412169	96,998718	-10,693215	-55,483131
22400000000	-23,632872	-167,70071	-9,5742855	86,129417	-9,5459023	86,272705	-10,862626	-58,727573
22500000000	-25,136047	-146,98779	-9,5756102	75,277565	-9,5568457	75,426453	-10,953167	-62,016232
22600000000	-23,077816	-129,98032	-9,5887156	64,361908	-9,5626125	64,477119	-10,834215	-64,926773
22700000000	-20,604618	-125,38316	-9,6204071	53,297089	-9,5907402	53,451591	-10,448664	-68,566048
22800000000	-18,807343	-131,45654	-9,6804256	42,091778	-9,6489153	42,303619	-10,055399	-72,245941
22900000000	-17,338161	-136,73283	-9,7854338	31,020138	-9,7548103	31,121494	-9,4567509	-76,967255
23000000000	-16,477585	-144,74608	-9,9219799	19,996817	-9,8970613	20,139612	-8,9070377	-82,454544
23100000000	-16,148968	-152,56876	-10,096926	9,1698523	-10,063476	9,3297291	-8,3023825	-88,949944
23200000000	-15,796678	-157,41251	-10,286361	-1,3573864	-10,250845	-1,2667861	-7,6653228	-96,294067
23300000000	-15,553098	-161,25958	-10,488229	-11,729761	-10,457413	-11,574265	-7,1296153	-104,40455
23400000000	-14,853048	-163,90335	-10,670909	-21,740644	-10,650972	-21,66515	-6,6570702	-112,21873
23500000000	-13,720008	-165,95688	-10,853534	-31,576477	-10,820571	-31,542791	-6,0919757	-121,46941
23600000000	-12,448855	-171,49297	-10,979033	-41,304127	-10,951526	-41,183388	-5,782022	-130,2648
23700000000	-11,131994	-178,63779	-11,112269	-51,038582	-11,079699	-50,953827	-5,375875	-139,31442
23800000000	-9,7319365	172,62103	-11,226453	-60,803963	-11,203017	-60,735565	-5,0617027	-148,89629
23900000000	-8,741044	161,78648	-11,33215	-70,58654	-11,309187	-70,508385	-4,8315101	-158,05621
24000000000	-7,835937	150,53526	-11,454814	-80,447311	-11,418735	-80,296402	-4,6028342	-167,25778
24100000000	-7,1530461	138,65886	-11,562777	-90,273003	-11,531035	-90,143295	-4,434761	-176,71609
24200000000	-6,5542669	127,28085	-11,684823	-99,867561	-11,6558	-99,724319	-4,2853975	173,8732
24300000000	-6,1516328	115,92649	-11,795286	-109,4505	-11,765857	-109,29012	-4,1524467	164,59514
24400000000	-5,828958	104,91473	-11,865407	-118,91387	-11,840203	-118,88526	-4,0599847	155,60713
24500000000	-5,5139651	94,617218	-11,944146	-128,55406	-11,915553	-128,49048	-3,9507911	146,29503
24600000000	-5,3520293	83,798332	-12,004802	-138,32893	-11,969017	-138,24289	-3,9040253	137,14934
24700000000	-5,2034407	74,035561	-12,072655	-148,271	-12,034123	-148,17636	-3,8348992	128,04045
24800000000	-5,0255547	63,332382	-12,176873	-158,29568	-12,153909	-158,22717	-3,7823896	119,00568
24900000000	-4,9251356	53,317799	-12,312229	-168,11484	-12,292364	-168,07974	-3,7292442	109,5711
25000000000	-4,8735061	43,57103	-12,475858	-177,85768	-12,445527	-177,72084	-3,7331672	100,39031
25100000000	-4,8764586	33,62075	-12,649803	-172,60187	-12,614261	-172,698	-3,6809402	91,183601
25200000000	-4,9396081	23,939392	-12,808381	163,29007	-12,77646	163,41579	-3,7279794	81,681442
25300000000	-4,9802632	14,7979	-12,992907	154,41754	-12,958975	154,37157	-3,7594755	72,390869
25400000000	-5,0629768	5,9900856	-13,133694	145,52269	-13,103751	145,6543	-3,8200424	62,990143
25500000000	-5,2117982	-2,9709103	-13,260952	136,92218	-13,238513	137,02162	-3,8981729	53,291695
25600000000	-5,3484125	-11,108016	-13,356745	128,46101	-13,317559	128,59901	-4,029356	44,226788
25700000000	-5,4231038	-18,780794	-13,383847	120,064	-13,366412	120,13259	-4,1436625	34,80891
25800000000	-5,4942889	-26,988691	-13,394542	111,631	-13,363081	111,73676	-4,3311424	25,580902
25900000000	-5,5736675	-34,70332	-13,366632	103,19536	-13,335857	103,30619	-4,5006475	16,545427
26000000000	-5,6082664	-42,320927	-13,310947	94,552322	-13,275303	94,592934	-4,6750493	7,5183215
26100000000	-5,5611744	-50,462807	-13,207631	85,605255	-13,181158	85,748863	-4,8911071	-1,3950273
26200000000	-5,6119142	-58,401615	-13,162372	76,038933	-13,126522	76,255875	-5,1310105	-9,9566393
26300000000	-5,4538937	-66,028732	-13,319554	67,897713	-13,271725	68,100143	-5,3885136	-17,907253
26400000000	-5,3171058	-75,126106	-13,093811	60,062992	-13,064489	60,190414	-5,529912	-26,16844
26500000000	-5,3286309	-84,021118	-12,855742	51,150669	-12,817673	51,27705	-5,7071357	-33,932114
26600000000	-5,3255978	-93,600876	-12,634808	41,839458	-12,619162	41,95919	-5,8024735	-41,491501
26700000000	-5,3672991	-102,79253	-12,437683	32,234314	-12,40288	32,405109	-5,9357314	-49,403324
26800000000	-5,38062	-112,58139	-12,255312	22,512983	-12,235072	22,534948	-5,9988575	-57,033676
26900000000	-5,5089102	-122,4443	-12,065997	12,485587	-12,043605	12,577772	-6,0280852	-64,330528
27000000000	-5,5860324	-131,41544	-11,90994	2,5358517	-11,885203	2,5020773	-6,0277944	-72,432655
27100000000	-5,7106271	-141,39345	-11,75353	-7,8796101	-11,714938	-7,9114194	-5,9451661	-79,761749
27200000000	-5,8899102	-150,52252	-11,59885	-18,417404	-11,580224	-18,363203	-5,8212762	-87,938255
27300000000	-6,0855722	-160,32877	-11,508865	-29,082113	-11,496679	-29,033918	-5,7355037	-95,709061
27400000000	-6,3963771	-169,18405	-11,415999	-39,753173	-11,376915	-39,782055	-5,6211233	-104,01936
27500000000	-6,6666861	-178,13501	-11,359772	-50,825218	-11,343001	-50,68404	-5,491209	-112,67502
27600000000	-6,9912357	173,11783	-11,367476	-61,478069	-11,355493	-61,67852	-5,3687763	-121,20351
27700000000	-7,2738051	165,3728	-11,430775	-72,416702	-11,404928	-72,432877	-5,272841	-130,07069
27800000000	-7,6425939	158,08311	-11,490368	-83,19429	-11,461959	-83,023643	-5,1791868	-139,00075
27900000000	-7,7278204	151,70067	-11,603036	-93,747131	-11,566978	-93,702637	-5,0150008	-148,02238
28000000000	-7,8399792	145,75862	-11,724208	-104,07241	-11,684464	-103,97594	-4,961432	-156,933
28100000000	-7,7361269	139,33931	-11,891535	-114,15137	-11,847594	-113,9471	-4,9331679	-166,10548
28200000000	-7,4689903	133,37865	-12,018953	-123,90956	-12,005738	-123,77919	-4,8656673	-175,53999
28300000000	-7,188365	126,17599	-12,156068	-133,20709	-12,107989	-132,98999	-4,865746	175,47794
28400000000	-6,7365923	118,36543	-12,229607	-142,60526	-12,235043	-142,53043	-4,8782306	165,91373
28500000000	-6,3298082	109,63816	-12,31863	-151,54942	-12,268023	-151,53836	-4,9094152	156,60768
28600000000	-5,9518032	100,53851	-12,374786	-160,64137	-12,319264	-160,49202	-4,9802861	146,92111
28700000000	-5,5870018	90,51207	-12,357433	-169,73363	-12,325445	-169,49304	-5,1412458	137,62375
28800000000	-5,3376913	80,975449	-12,336154	-178,68147	-12,298995	-178,46159	-5,235702	128,38126
28900000000	-5,1394539	70,216606	-12,280564	172,26933	-12,276111	172,53725	-5,4922104	118,84118
29000000000	-5,0294275	59,550541	-12,225988	163,16003	-12,20787	163,47144	-5,7580247	109,64019
29100000000	-4,9494362	49,036495	-12,117563	154,1261	-12,057064	154,32951	-6,0638938	100,92947
29200000000	-5,00421	37,509384	-11,997537	144,79565	-11,977296	144,98697	-6,435575	91,57119
29300000000	-5,2225142	26,579268	-11,877172	135,61105	-11,84726	135,73024	-6,8779979	83,353737
29400000000	-5,4236951	15,723498	-11,695129	126,32761	-11,658225	126,38313	-7,4149184	75,690147
29500000000	-5,8228307	4,3048782	-11,494336	116,70057	-11,476028	116,91895	-7,9062624	68,233406
29600000000	-6,3995295	-6,2131462	-11,218803	107,09551	-11,20387	107,34247	-8,4938726	61,533039
29700000000	-7,172163	-16,944384	-10,961895	96,989899	-10,939323	97,353546	-9,1121731	55,354939
29800000000	-8,081233	-26,687168	-10,650681	86,521713	-10,637489	86,926765	-9,6259212	50,607262
29900000000	-9,092947	-34,71088	-10,357552	75,725967	-10,358233	75,827194	-10,188594	47,064411
30000000000	-10,541929	-42,052193	-10,074226	64,242004	-10,055782	64,543159	-10,561872	44,869633
30100000000	-11,756906	-45,647945	-9,8401356	52,230034	-9,8114319	52,440392	-10,718641	43,058857
30200000000	-13,148131	-49,279598	-9,6842155	39,880974	-9,6560078	40,114502	-10,542119	42,065201
30300000000	-14,221875	-48,384045	-9,6102257	27,17609	-9,5863047	27,531309	-9,9930811	40,059799
30400000000	-15,117842	-46,632259	-9,5941124	14,687183	-9,5808411	14,791776	-9,2216415	36,307339
30500000000	-15,85946	-41,733204	-9,7037811	2,0192423	-9,6555853	2,198828	-8,4790268	30,702642
30600000000	-15,658315	-36,475071	-9,9331598	-10,392238	-9,8647184	-10,124532	-7,588438	24,170038
30700000000	-14,385692	-35,047504	-10,046346	-21,867626	-10,01859	-21,776531	-6,8457503	15,676226
30800000000	-14,015933	-34,772583	-10,198044	-33,73967	-10,135511	-33,500046	-6,1620145	6,3546977
30900000000	-13,244514	-36,067039	-10,383323	-45,578114	-10,382154	-45,44389	-5,5967236	-3,3308868
31000000000	-12,247388	-39,260345	-10,720457	-56,890667	-10,654738	-56,715534	-5,0576425	-13,317352
31100000000	-11,364135	-42,444099	-10,983947	-68,058357	-10,948765	-67,954643	-4,6082859	-24,00872
31200000000	-10,376287	-49,945992	-11,240093	-78,822426	-11,192313	-78,806725	-	

6 WR-28 to microstrip Two way Power Combiner

The design process of this power combiner was developed in several stages. The first step consisted of a detailed analysis of the input power splitting between the two branches, with the aim of ensuring a balanced and controlled distribution of power. The geometry of the two transmission arms was investigated in order to optimize their electrical behaviour and maintain the desired performance over the operating bandwidth. A stepped configuration, as shown in Figure 96, made of six sections was adopted. This solution was selected because it allows a stepped impedance transition and offers greater flexibility in tailoring the power division profile. The resulting power distribution is shown in Figures below in terms of S-parameters.

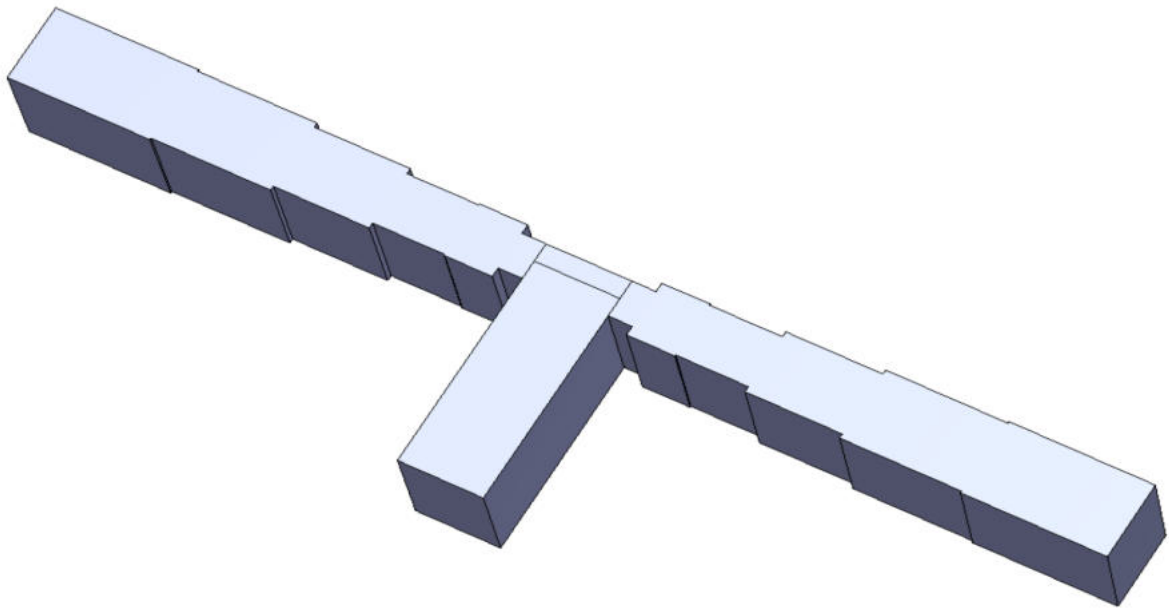


Figure 96 - Stepped t junction

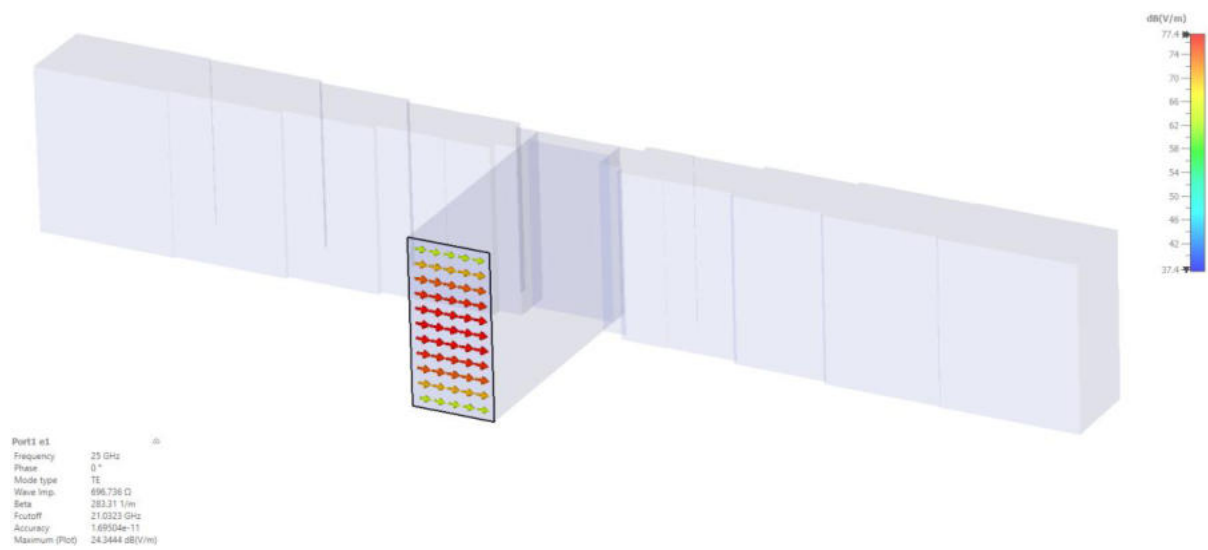


Figure 97 - E field

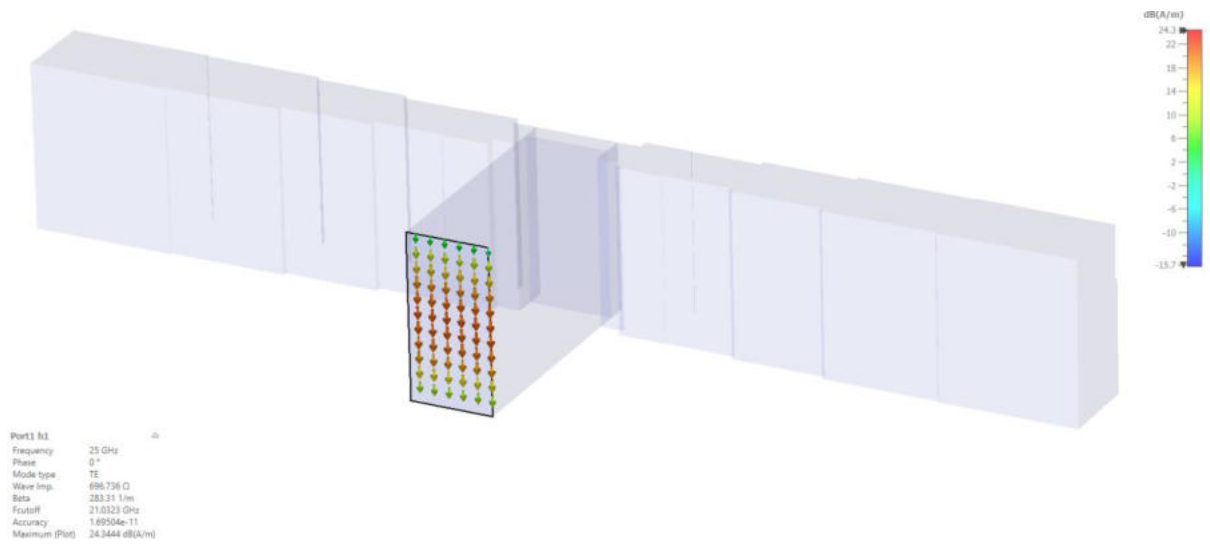


Figure 98 - H field

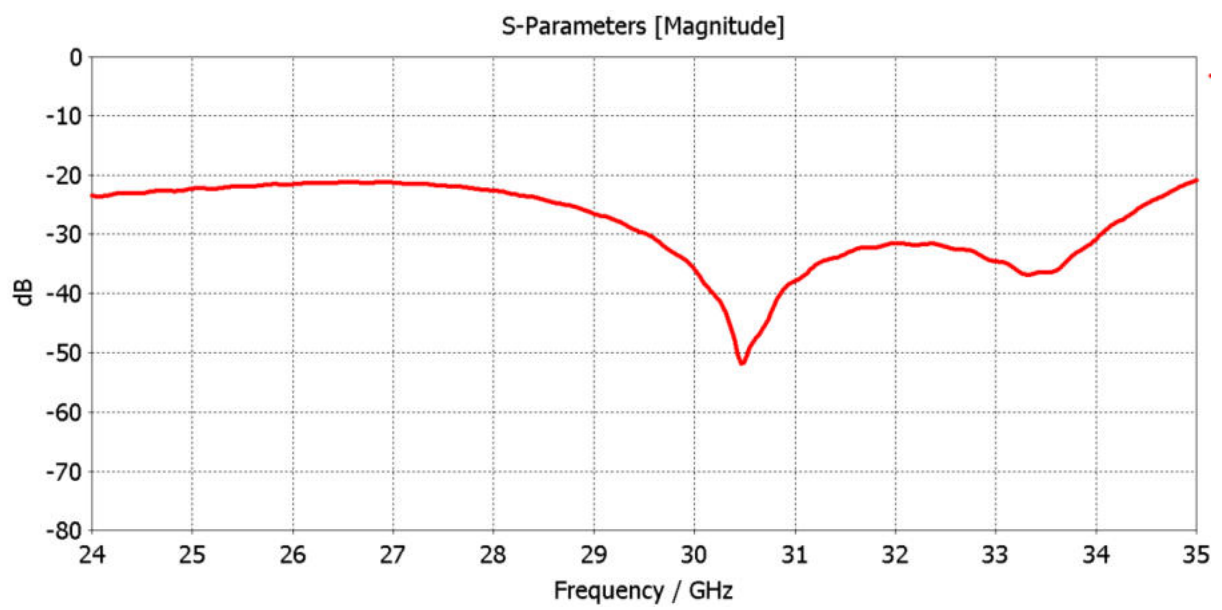


Figure 99 - Return loss

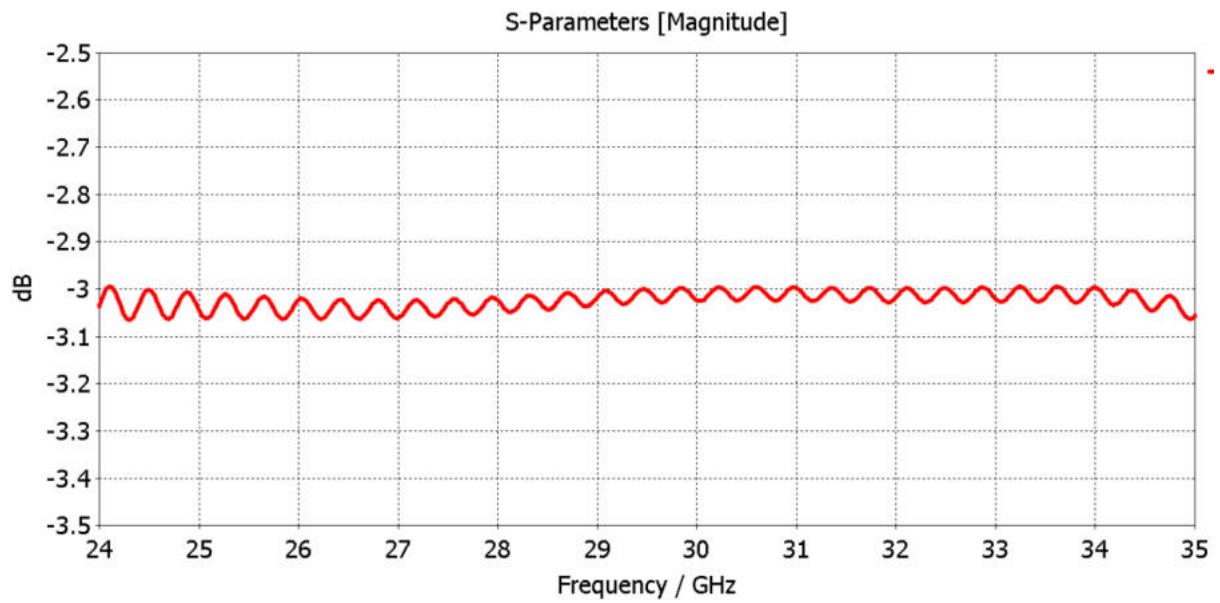


Figure 100 - S21, S31 Amplitude

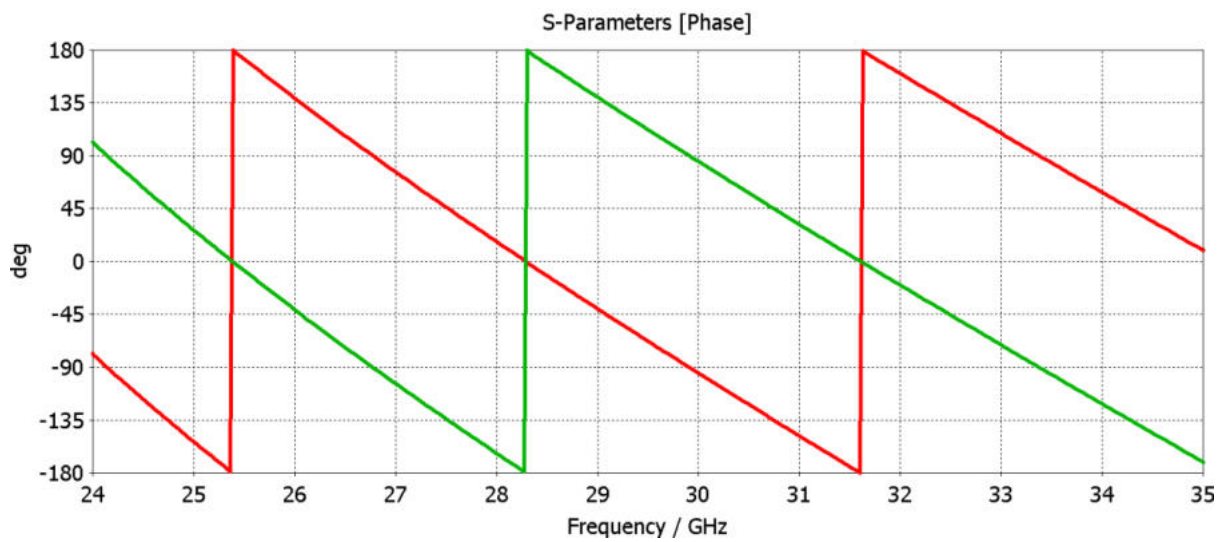


Figure 101 - S21, S31 phase shift

Subsequently, an alternative two-way input splitting configuration was adopted. The structure shown in Figure 102 was selected because it provides improved impedance matching with respect to the stepped solution previously considered.

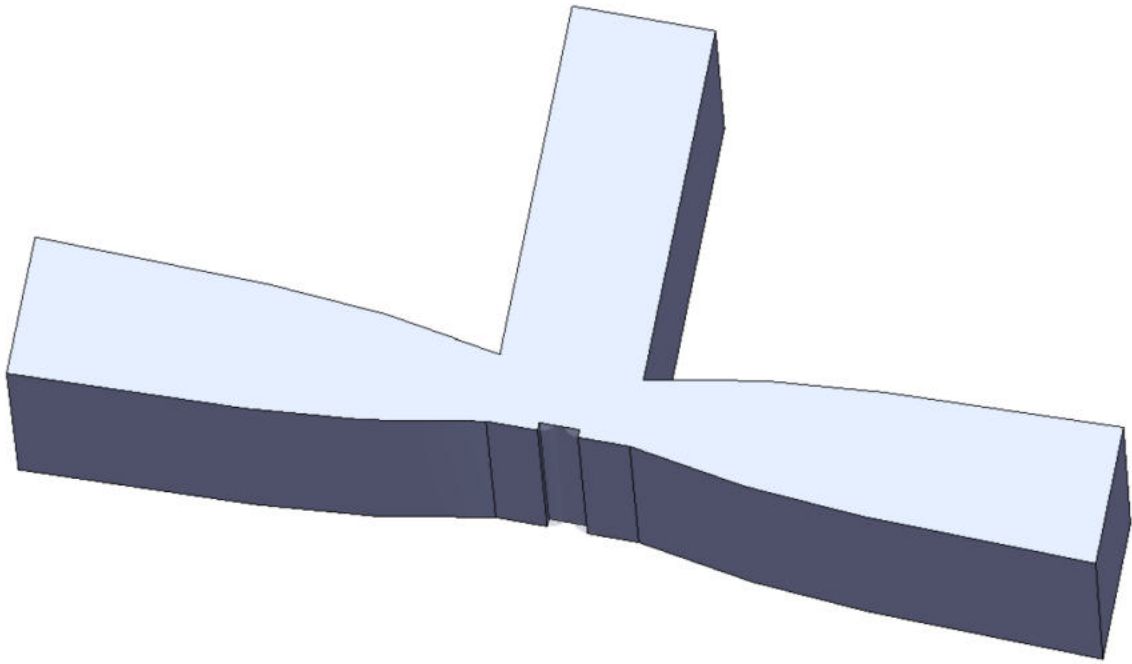


Figure 102 - T junction

This better matching condition helps reduce the input reflection coefficient and therefore contributes to a more efficient power transfer toward the two output branches.

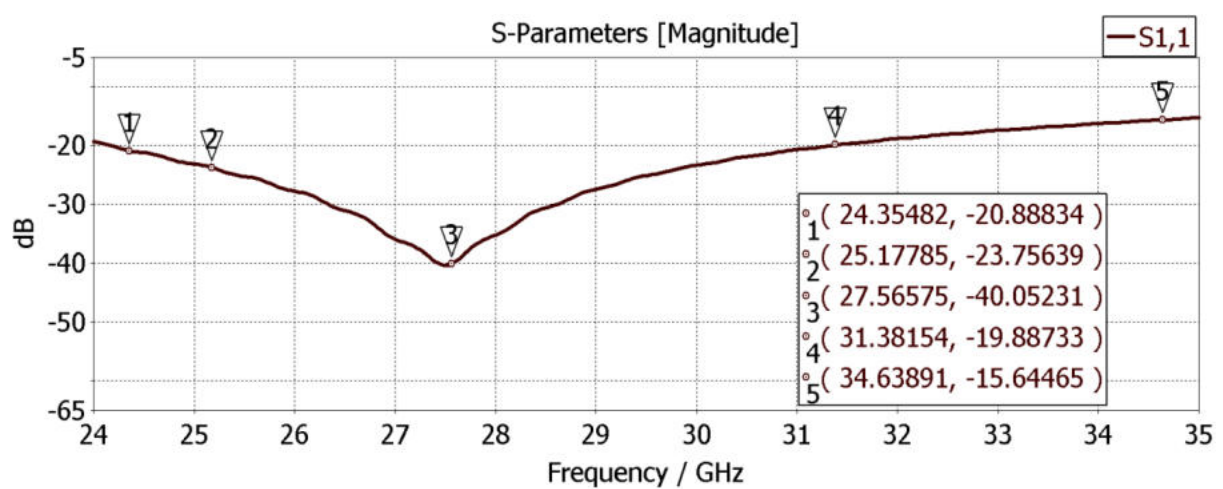


Figure 103 - Return loss

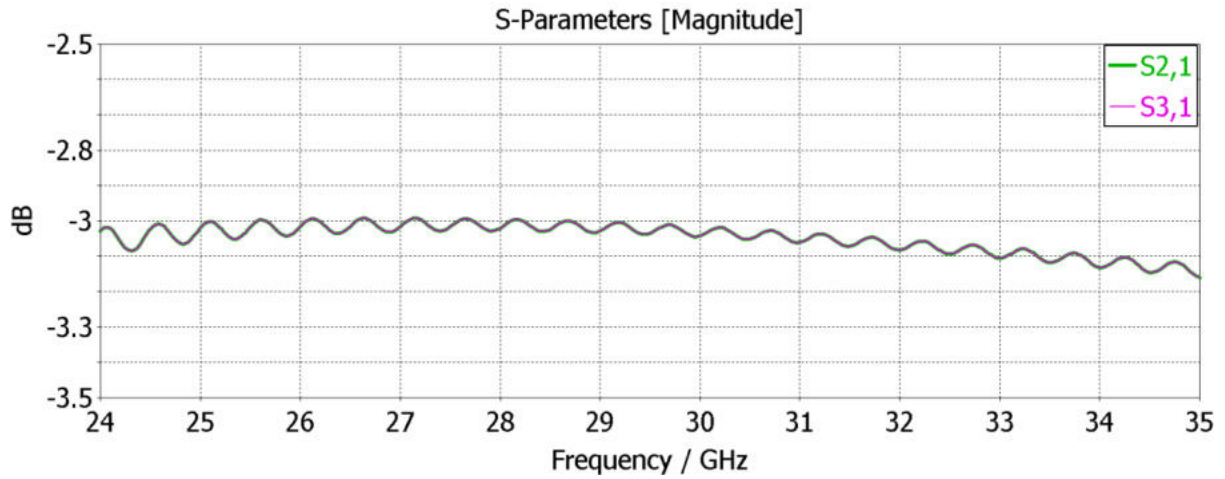


Figure 104 - Trasmission Power division

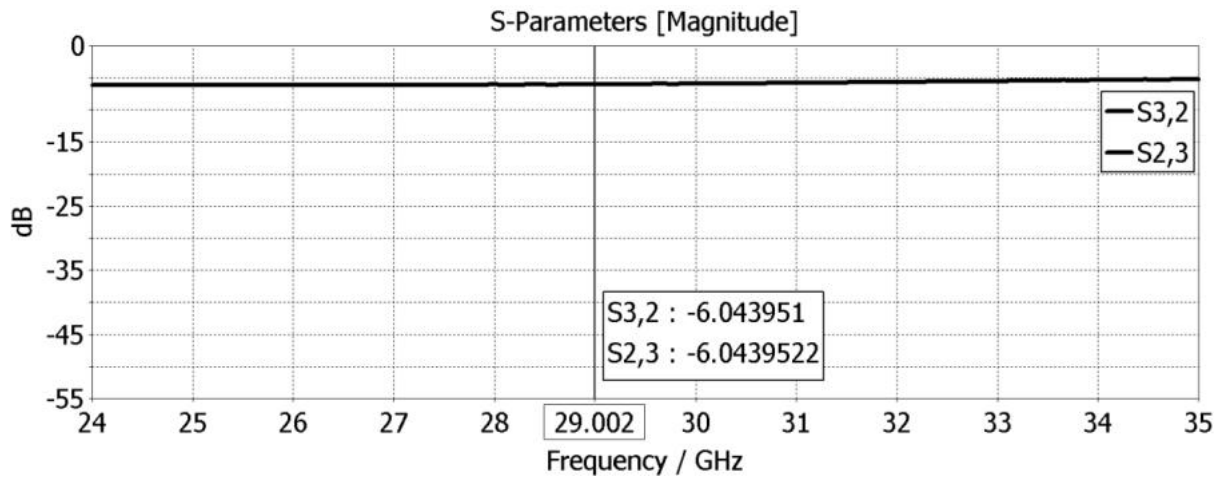


Figure 105 - Isolation between ports

In addition, the new configuration offers a good transition between the input section and the subsequent network. For this reason, this solution was preferred for the final implementation.

The microstrip transmission line was designed with the specific goal of achieving broadband impedance matching across the operating frequency range. To this end, a stepped impedance microstrip geometry was implemented, consisting of 7

distinct sections characterized by progressively varying widths and lengths. This multi-step configuration enables a gradual impedance taper, effectively minimizing reflections and optimizing power transfer over a wide bandwidth, as illustrated in Figure 106.

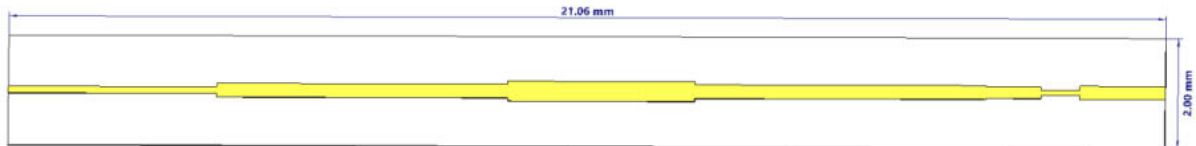


Figure 106 - stepped microstrip

The wider-width line section is dimensioned at $50\ \Omega$. In this section, it is envisaged to position an MMIC HPA to amplify the signal.

The dimensions of the stepped microstrip are shown in Table 10.

Parameters	Values
H_substrato	0.127
L_micro1	2.2
H_micro1	0.035
w_micro1	0.19061973590129
L_micro2	1.5
H_micro2	0.035
w_micro2	0.12981956583934
L_micro3	16.740881346965
H_micro3	0.035
w_micro3	0.28109835567144
L_tot_microstriscia	L_micro1+L_micro2+L_micro3+L_micro4+L_micro5+L_micro6
L_micro4	1
w_micro4	0.27568694407986
L_micro5	0.7
w_micro5	0.097988777806472
L_micro6	1.6
w_micro6	0.16797431012117
L_micro_1_2	1.6+0.178
L_micro_2_2	0.7
L_micro_3_2	1
L_micro_4_2	7
L_micro_5_2	L_micro_4_2
H_substrato_2	0.127
w_substrato_2	2
w_micro_1_1	0.21
w_micro_2_2	0.09
w_micro_3_2	0.21
w_micro_4_2	0.25
L_micro_6_2	1.5
L_micro_7_2	2.3
w_micro_5_2	w_micro_4_2
w_micro_6_2	0.12
w_micro_7_2	0.12001

Table 10 - Overall size of the microstrip [mm]

Finally, the complete model was developed, as shown in Figure 107.

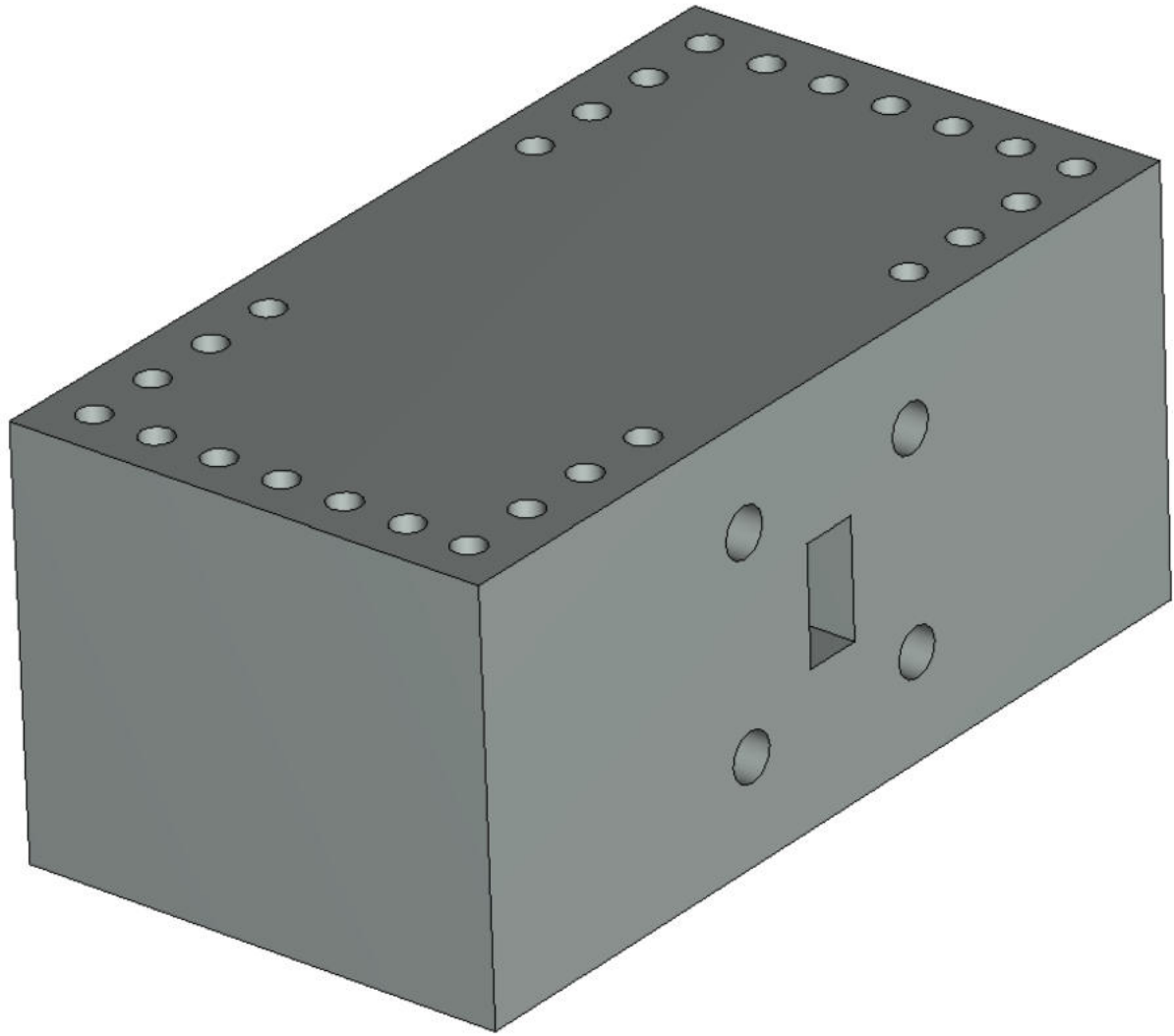


Figure 107 - Power combiner - Perspective view

The two-way combiner has an overall size of $55.068 \times 29.878 \times 25.112$ mm. The model is made by aluminium. In the structure, holes have been created to screw in the connector for the WR-28 rectangular waveguide-to-coaxial transition, enabling interfacing with the measurement instrumentation.

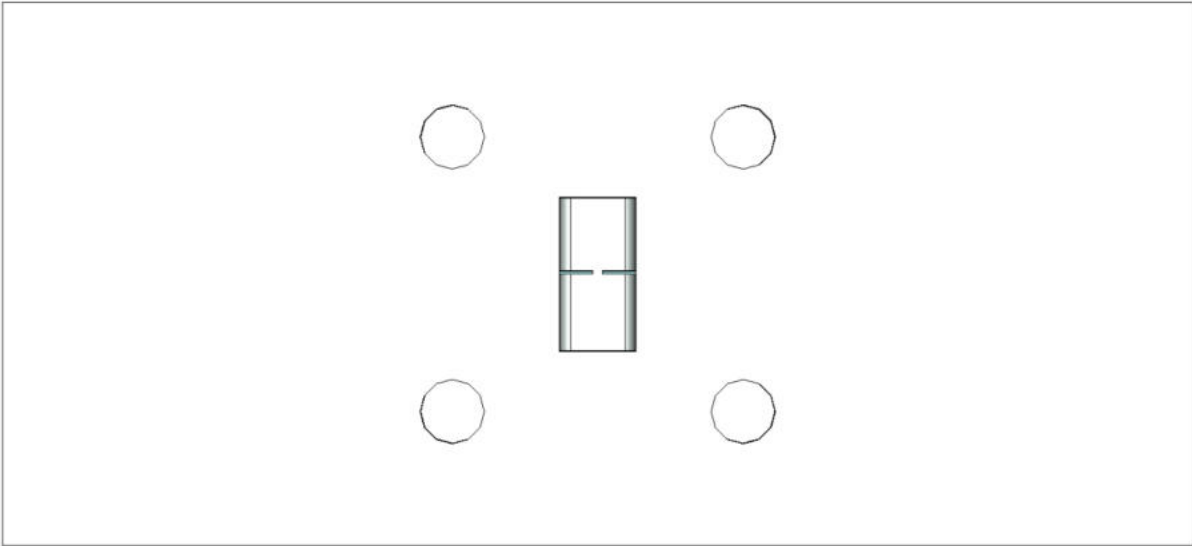


Figure 108 - output Front view

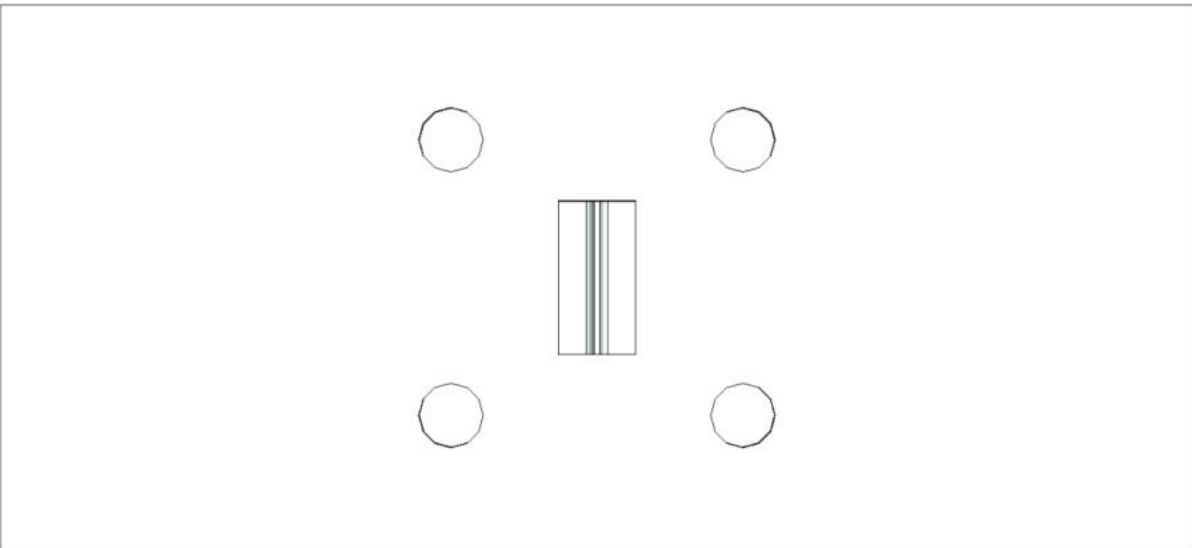


Figure 109 - Input front view

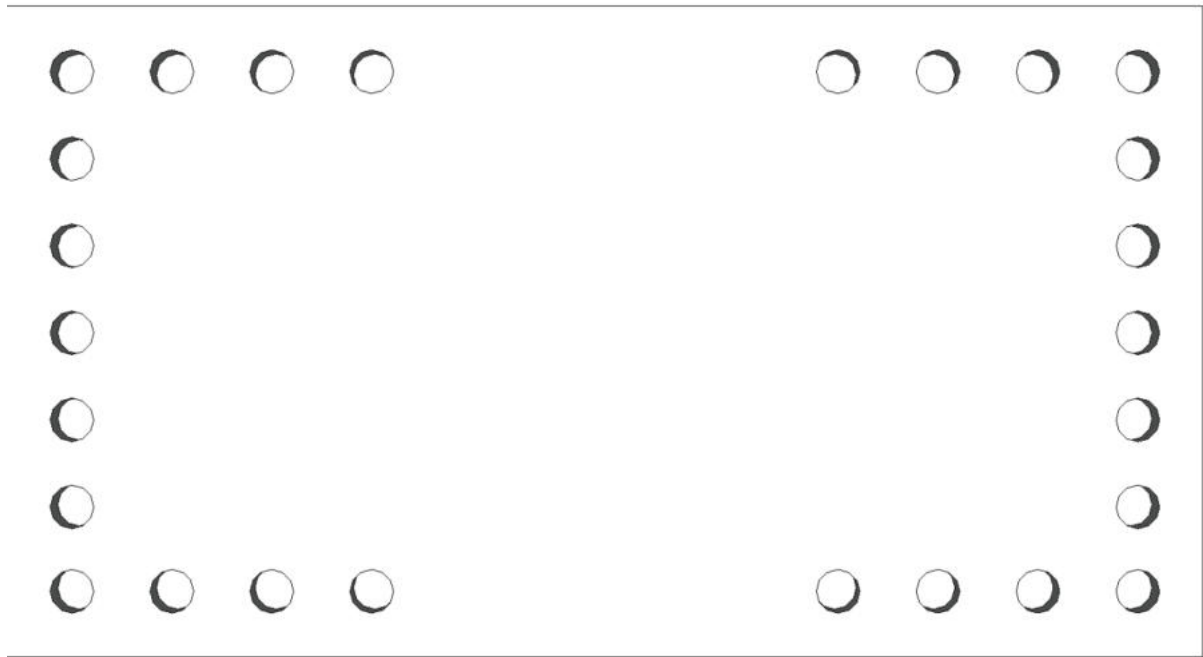


Figure 110 – Top/Bottom view

Once the T-junction was properly designed, the process moved to the next step. This consisted of optimally positioning the recombination microstrips to ensure excellent coupling during the rectangular waveguide-to-microstrip transition. In Figure 11, the overall model cross-section is illustrated.

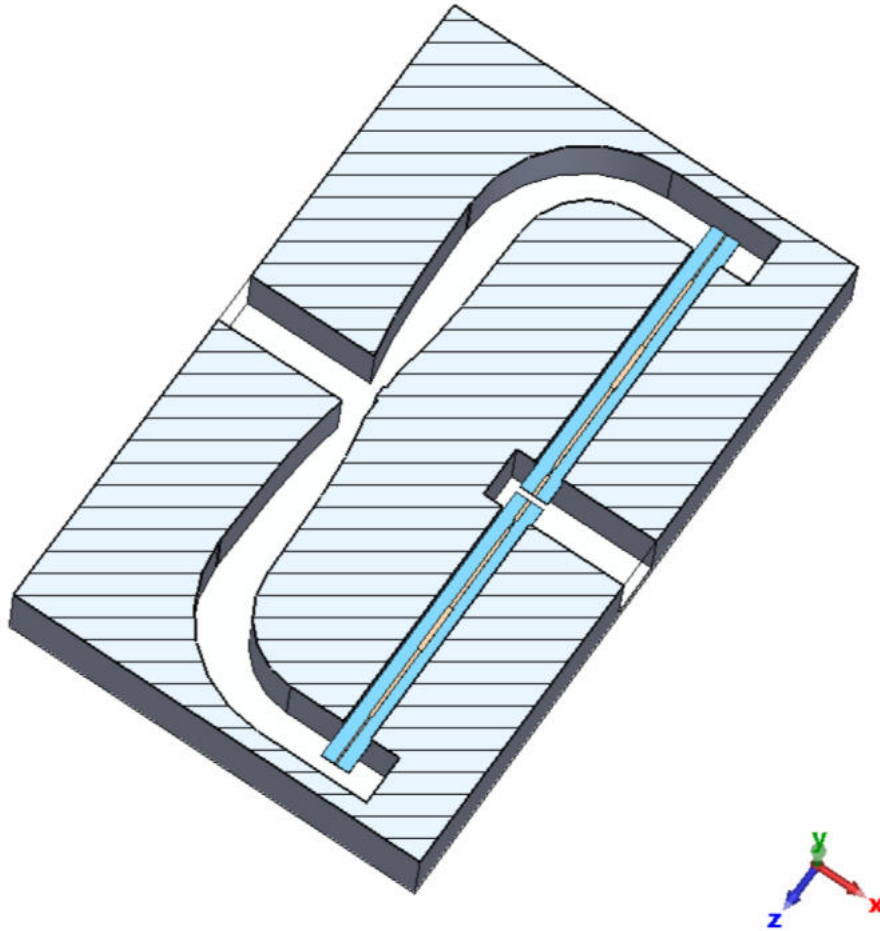


Figure 111 - 2-way power combiner cross-section

Optimal coupling is achieved by positioning the microstrip at a distance of 3.1 mm (corresponding to $0.22 \lambda_g$) from the short-circuit (along the x-axis) and 1.256 mm from the waveguide wall (along the z-axis). In formula we get: the free-space wavelength (λ_0) at 30 GHz is 10.000 mm. For the WR-28 rectangular waveguide ($a=7.112$ mm, $b=3.556$ mm):

- Cut-off frequency $f_c = 21.091$ GHz
- Cutoff wavelength $\lambda_c = 14.224$ mm

- Guide wavelength $\lambda_g = 14.062$ mm

In addition, the microstrip section within the waveguide behaves as a radiating element that captures the TE_{10} mode at the region of maximum field intensity and converts it to TEM mode during propagation (in the recombination phase). The microstrip penetration is 2.3 mm (along the z-axis). Figure 112 shows the drawing of the combining structure from different perspectives.

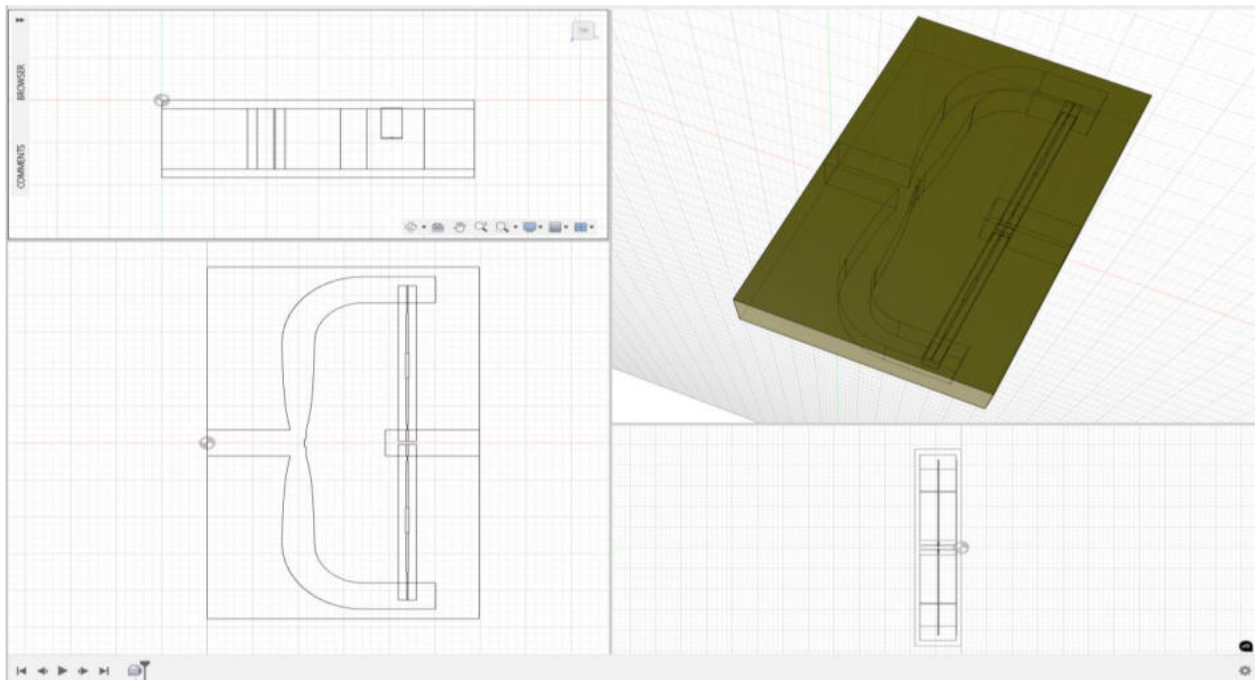


Figure 112 - Power combiner - sketch

6.1 Simulation results of the Power Combiner

Figure 113 and Figure 114 present the simulated performance of the two-way power combiner.

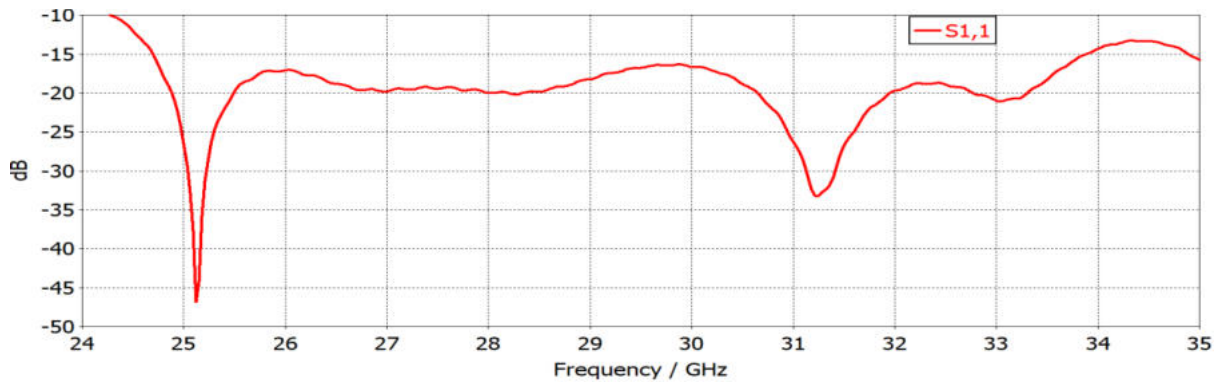


Figure 113 - Return loss

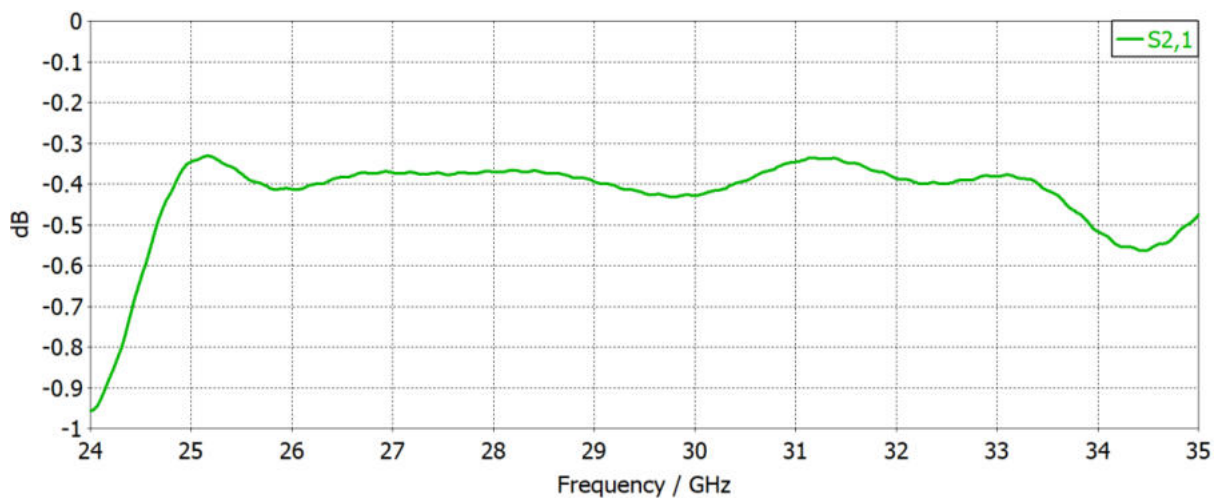


Figure 114 - Insertion loss

The frequency response curves reveal a good broadband impedance matching across the operational bandwidth. Two different resonant peaks are observed at the resonant frequencies of 25 GHz and 31 GHz, where insertion losses reach a minimum of ~ 0.3 dB, an exceptionally low value that validates the

goodness of the combining structure. The average insertion loss throughout the 24 - 35 GHz operating bandwidth measures approximately 0.35 dB, with characteristic degradation at the edges where impedance matching progressively gets worse.

6.2 Step before the PC fabrication process

Considering the outstanding simulated performance achieved, the 3D model was subsequently modified to accurately represent the manufacturing process. The aim was to ensure that the simulated S-parameters closely matched the measured performance obtained with vector network analyzer instrumentation. Finally, Gerber files were generated for fabrication.

It was assumed to mill an aluminium block using a 2 mm diameter titanium milling tool, an extremely durable material. Accordingly, all 90° sharp corners in the original design were replaced with 2 mm radius circular fillets to faithfully represent the tooling constraints. This manufacturing-aware 3D model refinement achieved the high-fidelity detail level illustrated in Figures 115 and 116 (before and after).

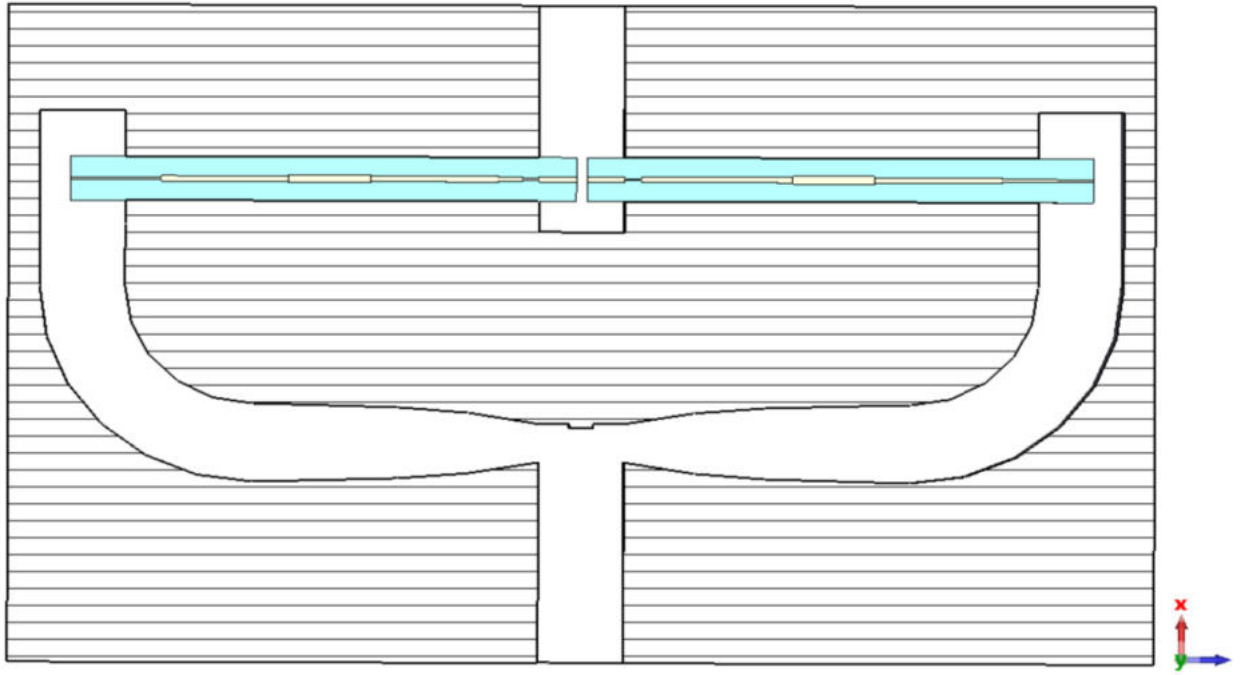


Figure 115 - Before

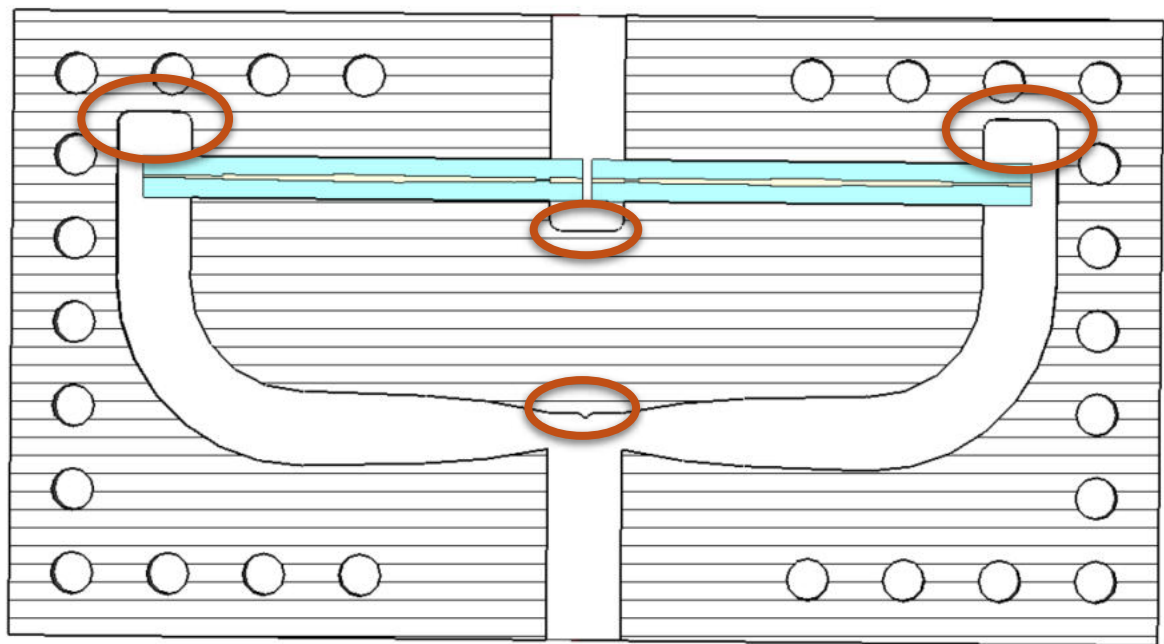


Figure 116 - After

6.2.1 Simulation results

Since the structure was adapted to the manufacturing process employing a 2 mm titanium end mill, the simulations were repeated. Figure 117 demonstrates that by modelling the corners with the 2 mm tool results in minimal performance changes.

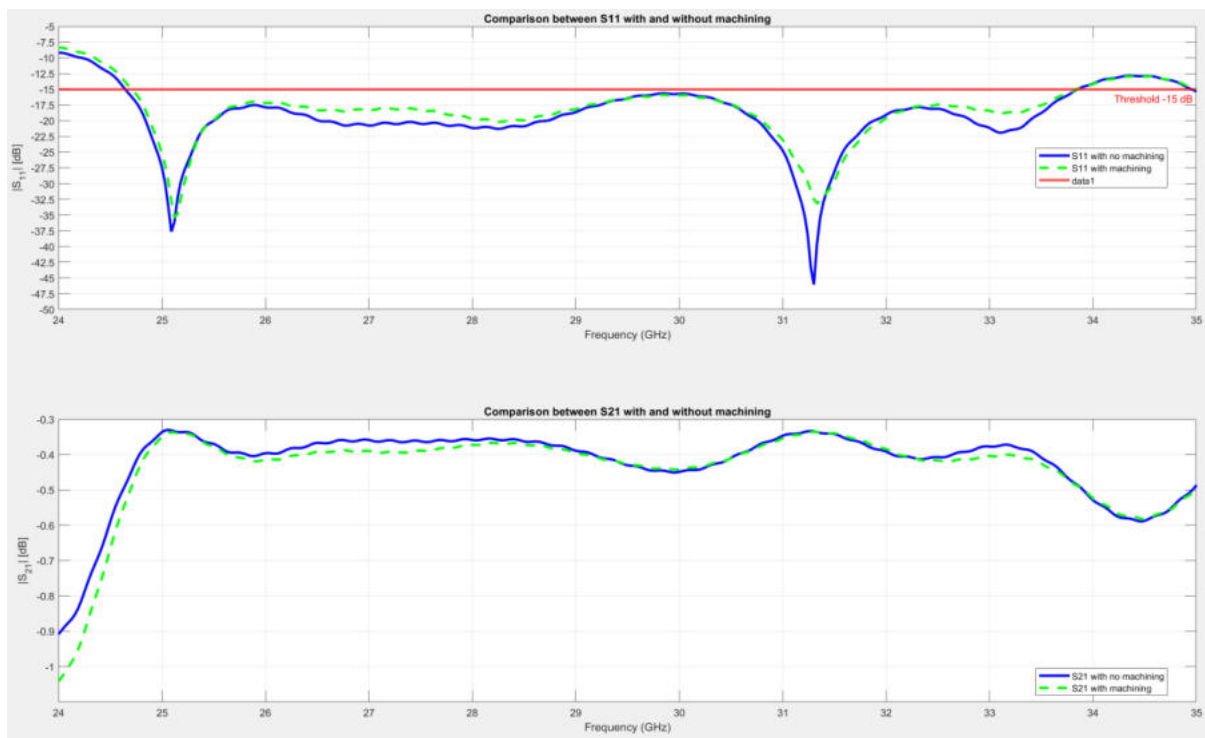


Figure 117 - Performance comparison

The S_{11} parameter exhibits a slight degradation between 24-28 GHz after machining, while insertion losses increase by less than 0.5 dB. The structure maintains excellent broadband matching across the operational bandwidth.

The S-parameters of the structure are separately re-proposed in Figures below.

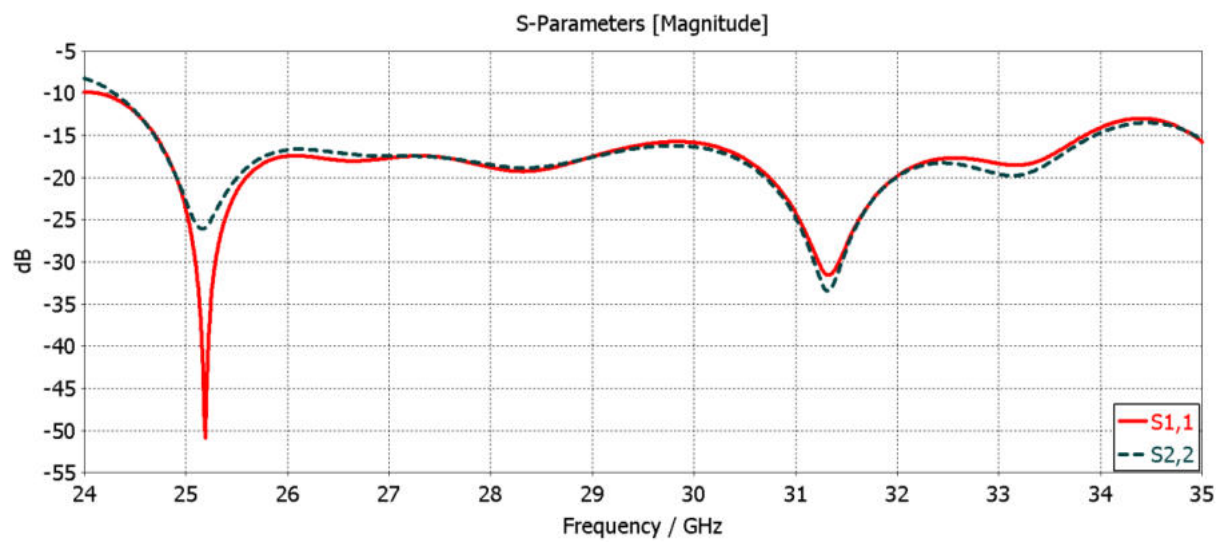


Figure 118 - Return loss

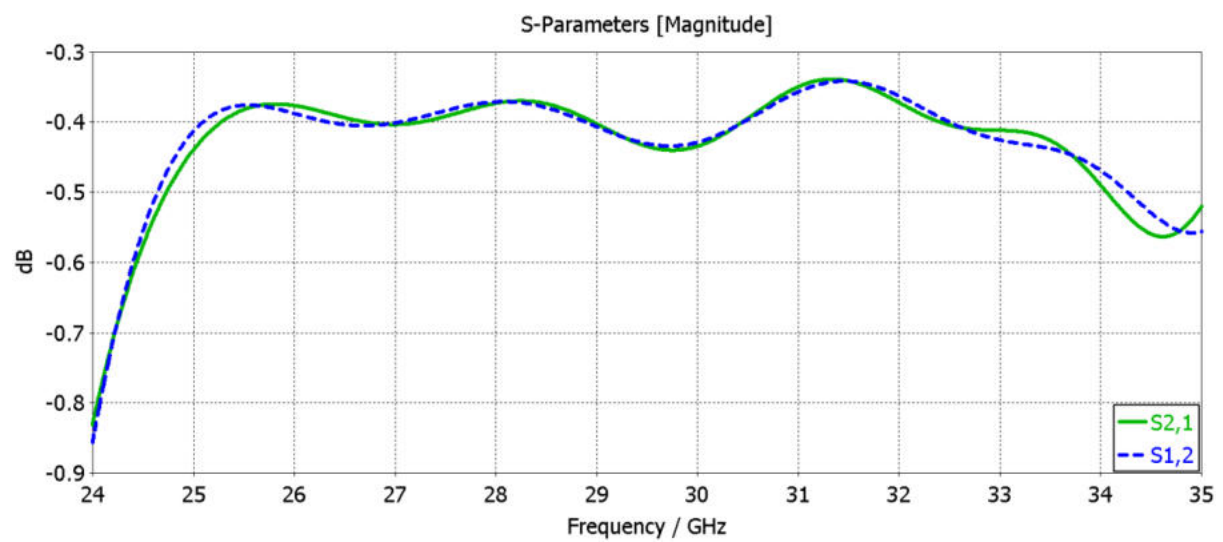


Figure 119 - Insertion loss

7 Scalability of the N-Way Power Divider Model

In the previous sections, it was explained how the 2-way combining model based on the waveguide-to-microstrip transition was extremely effective, particularly for broadband applications. However, you might think that a two-way combining structure offers limited utility since it would deliver low power levels at the output port. Actually, the presented combining structure has a key property that makes it extremely valuable for applications requiring high combined power at the output port. It is, in fact, highly scalable. This means that starting from the 2-way structure, it is possible to create an N-way division and combining network while maintaining low losses.

The approach adopted below uses a modular block-based methodology. This approach enables complete characterization of the structure using stand-alone characterized blocks that are simply interconnected via nets. The major advantage is a deeply reduction in computational power and simulation time (from several hours to just 1 minute or less).

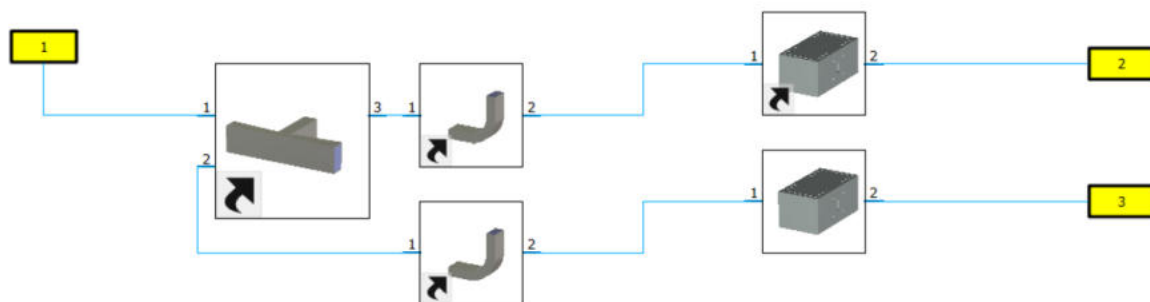


Figure 120 - 2 way (4 microstrips) - Power Divider

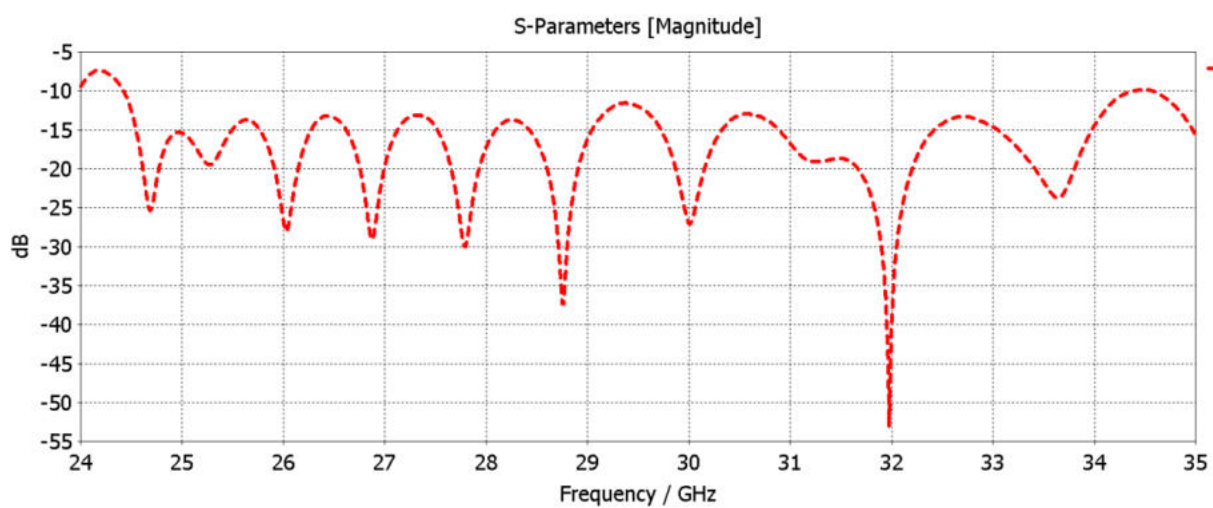


Figure 121 - Return loss

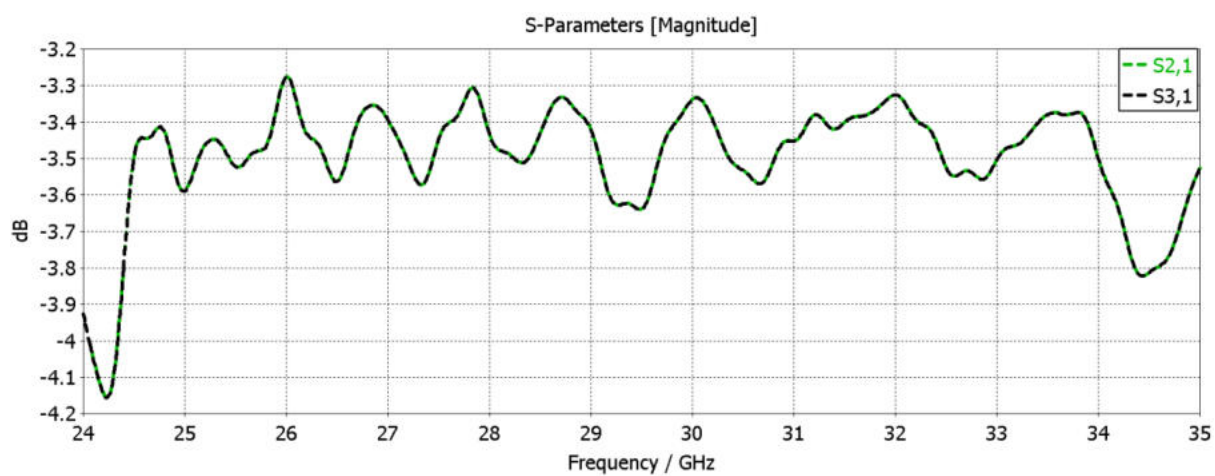


Figure 122 - Trasmission Power division

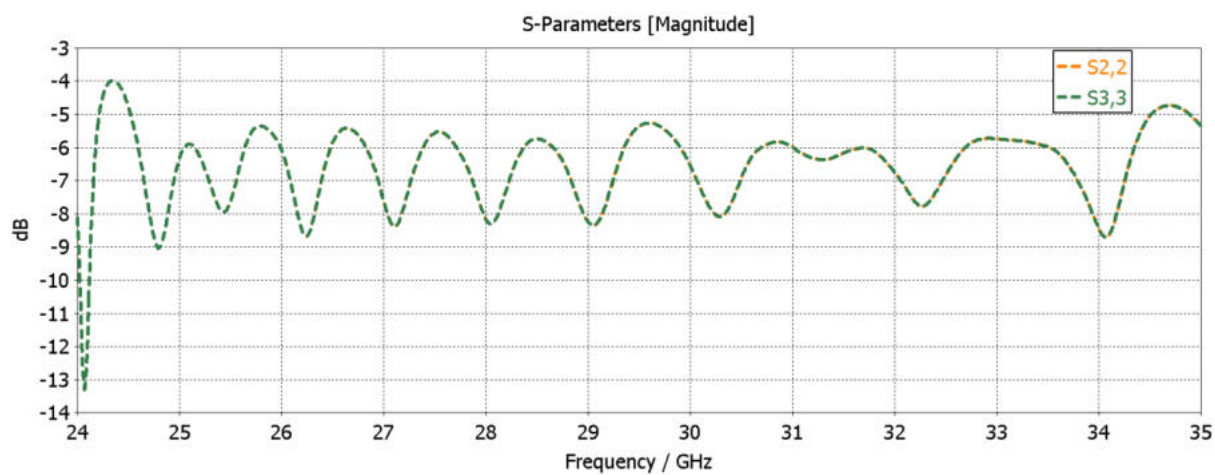


Figure 123 - Isolation between ports

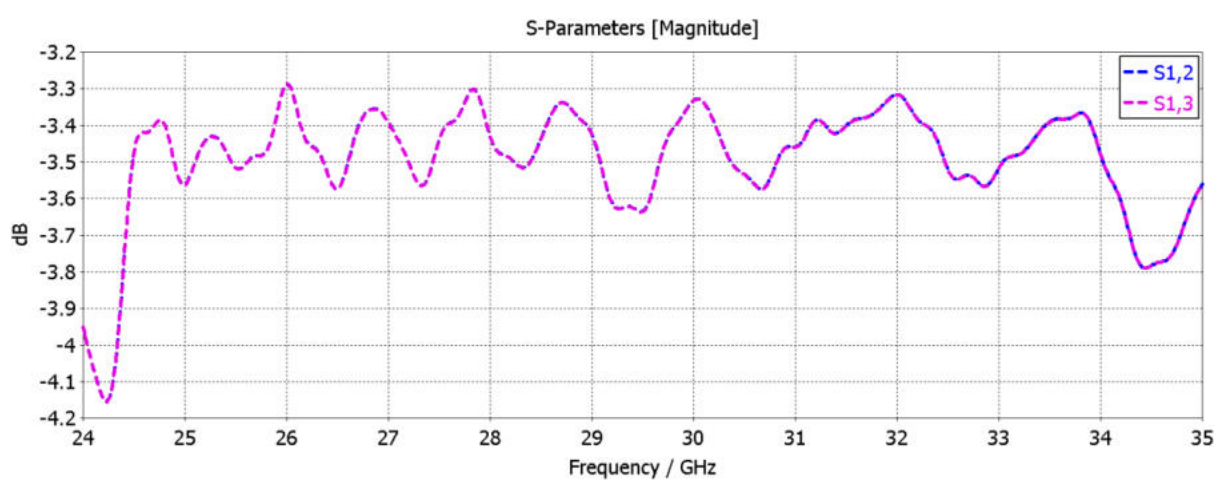


Figure 124 - Trasmission Power division

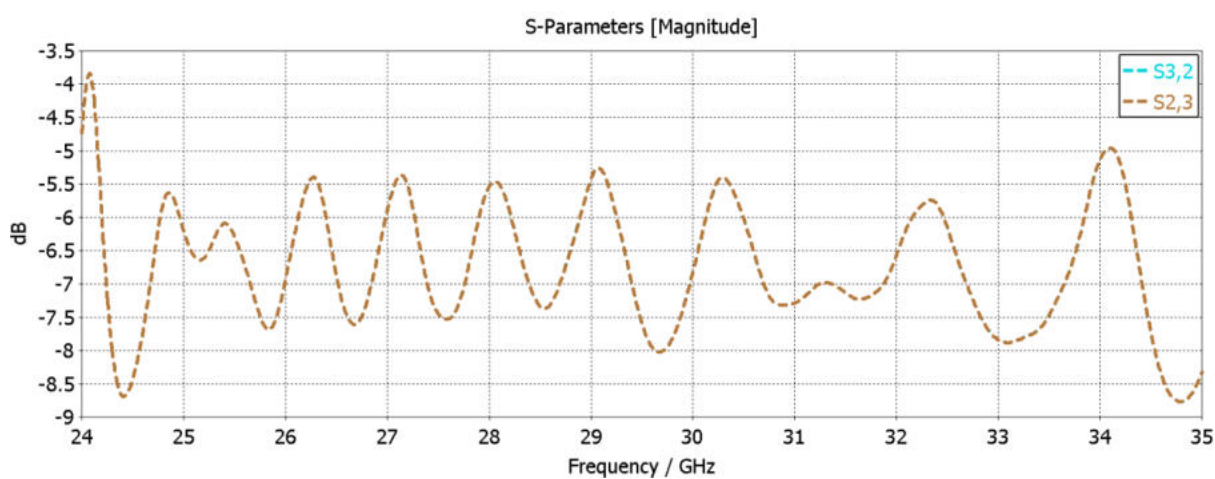


Figure 125 - Isolation between ports

7.1 4-Way Power Combiner Model

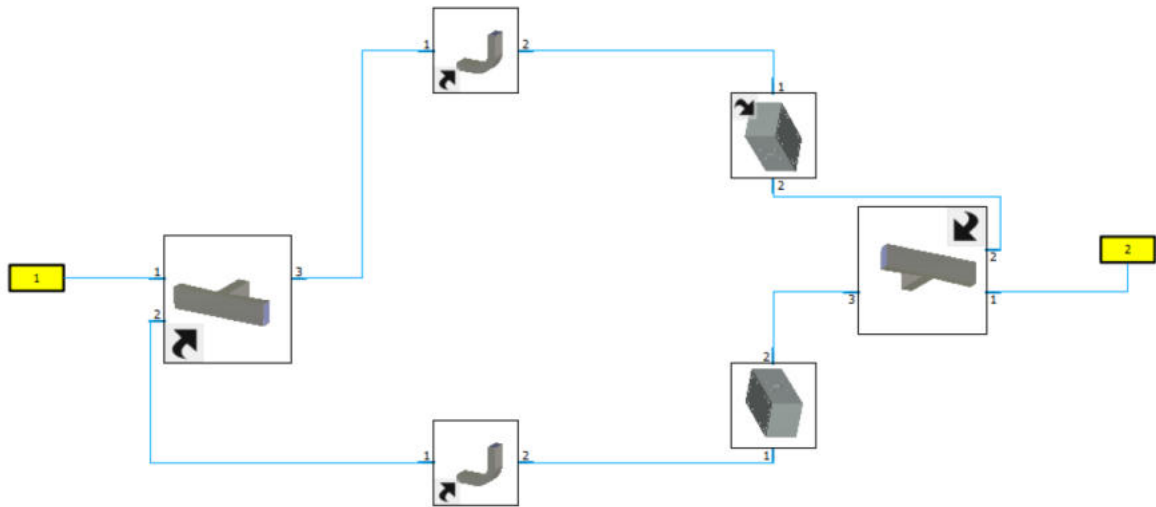


Figure 126 - 4-way power combiner block-based approach

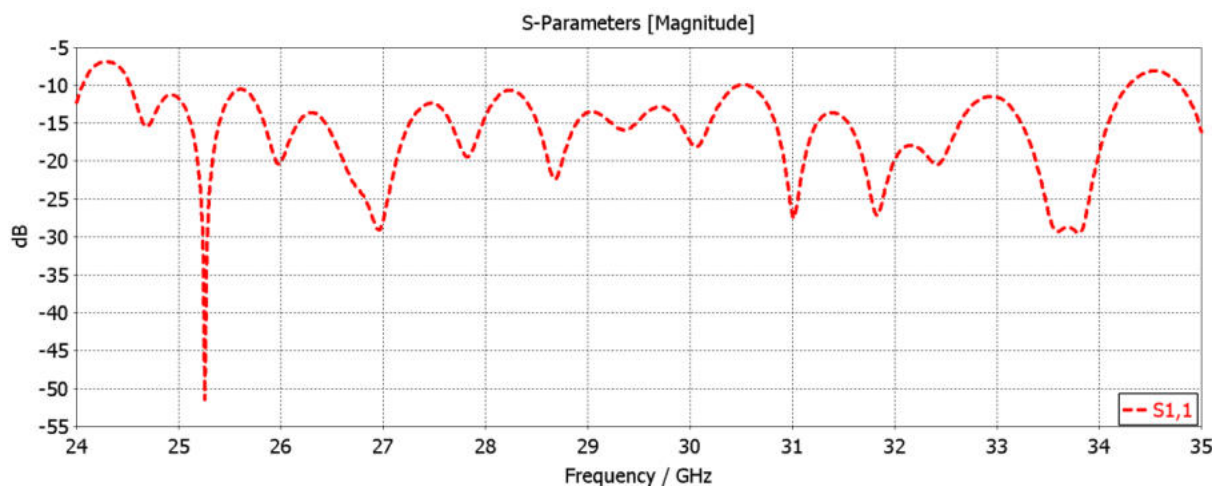


Figure 127 - Return loss

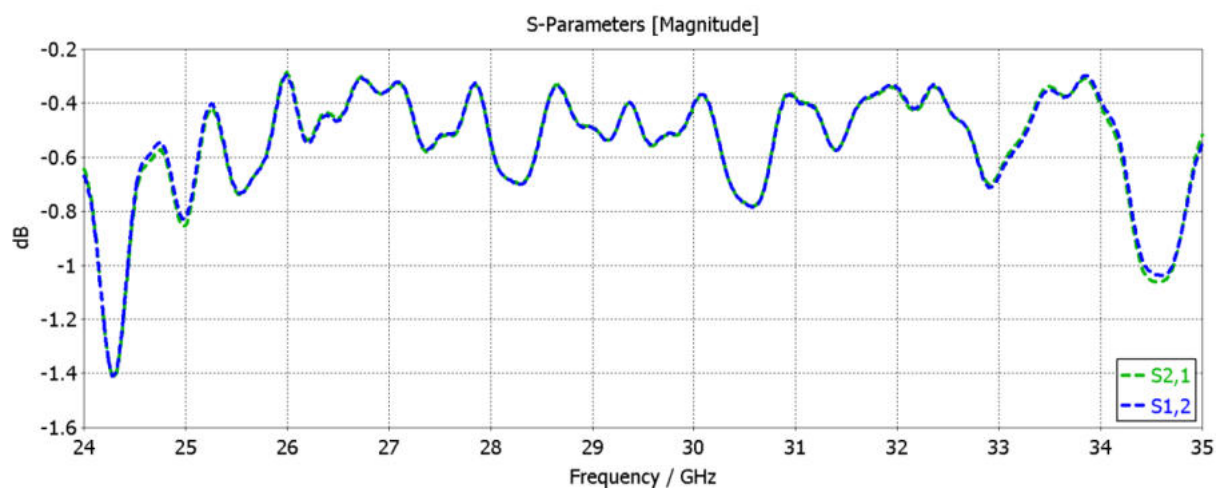


Figure 128 - Insertion loss

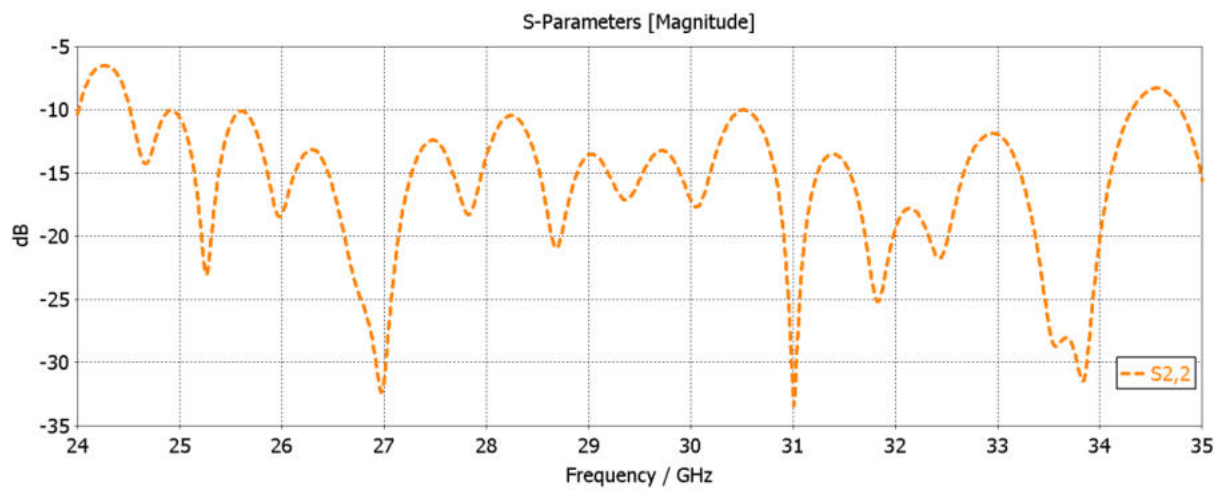


Figure 129 - Return loss

7.2 2-Way Power Divider Model with 8 microstrips

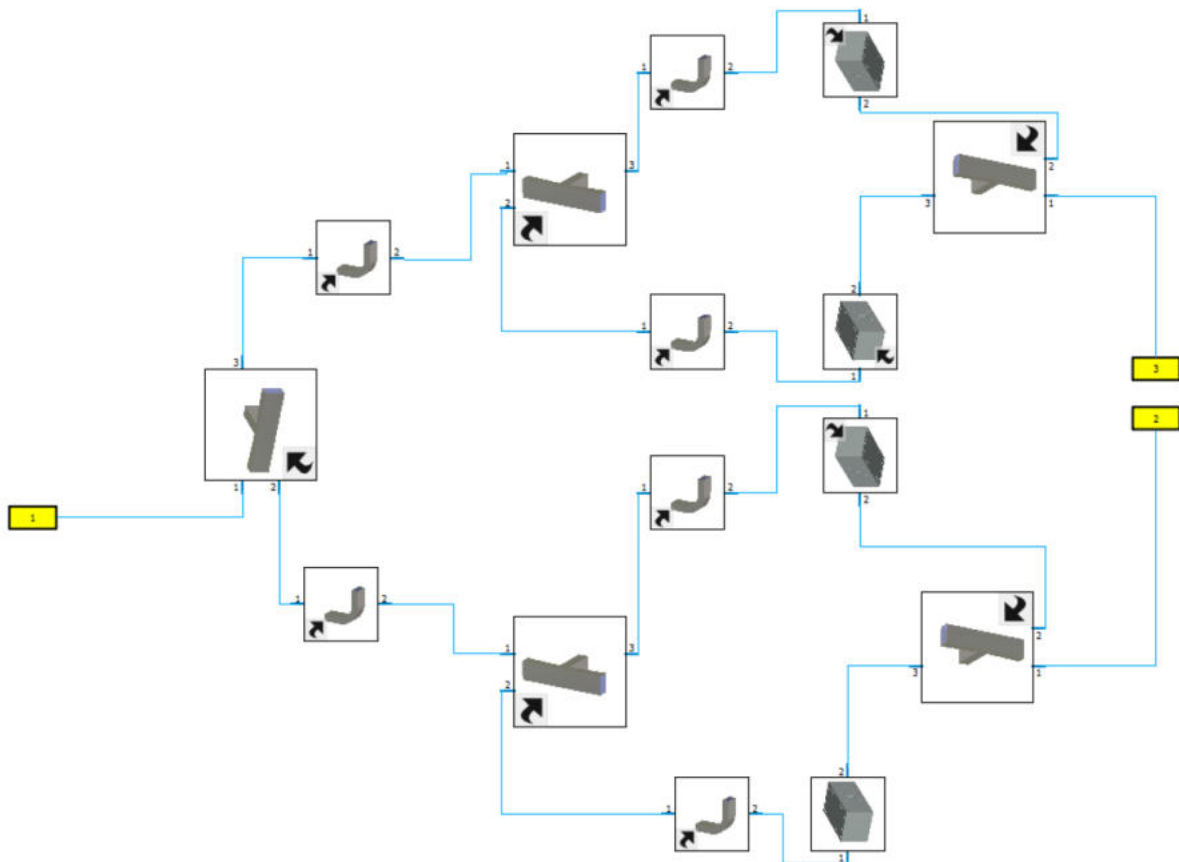


Figure 130 - 8 way power divider block-based approach

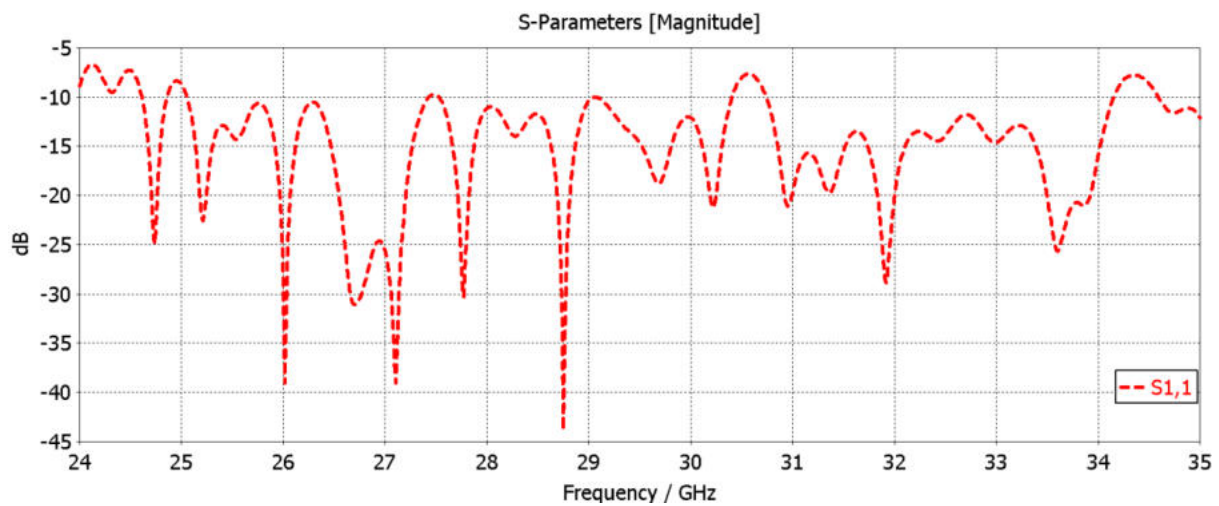


Figure 131 - Return loss

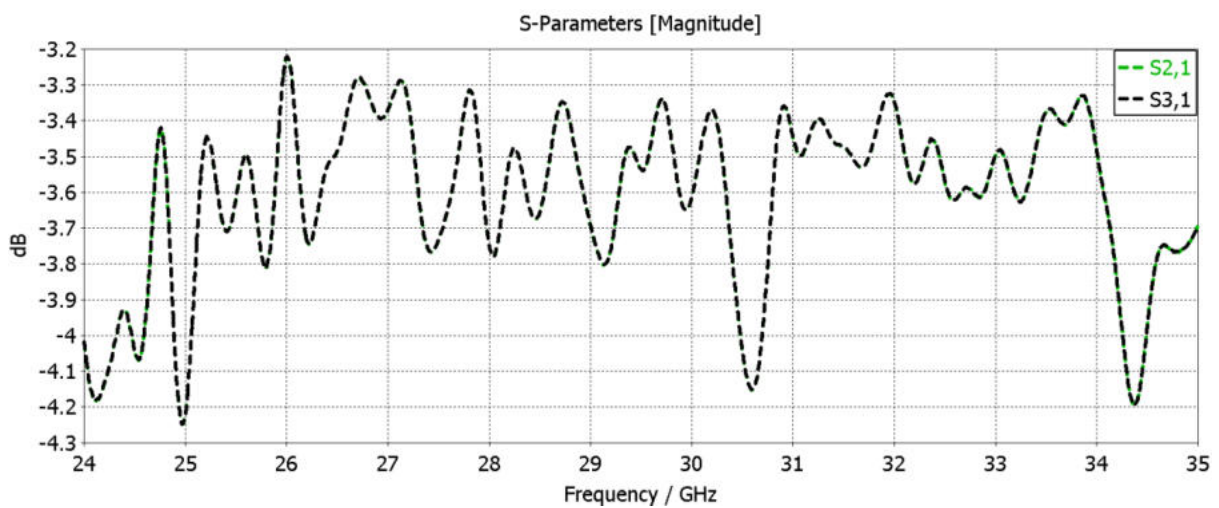


Figure 132 - Trasmission Power division

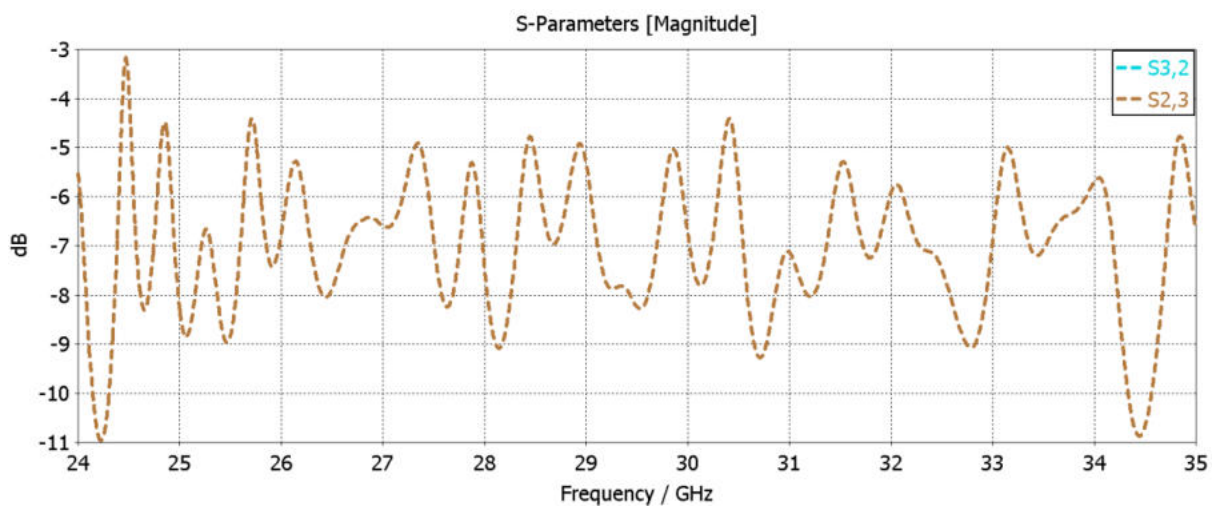


Figure 133 - Isolation between ports

7.3 8-Way Power Combiner Model

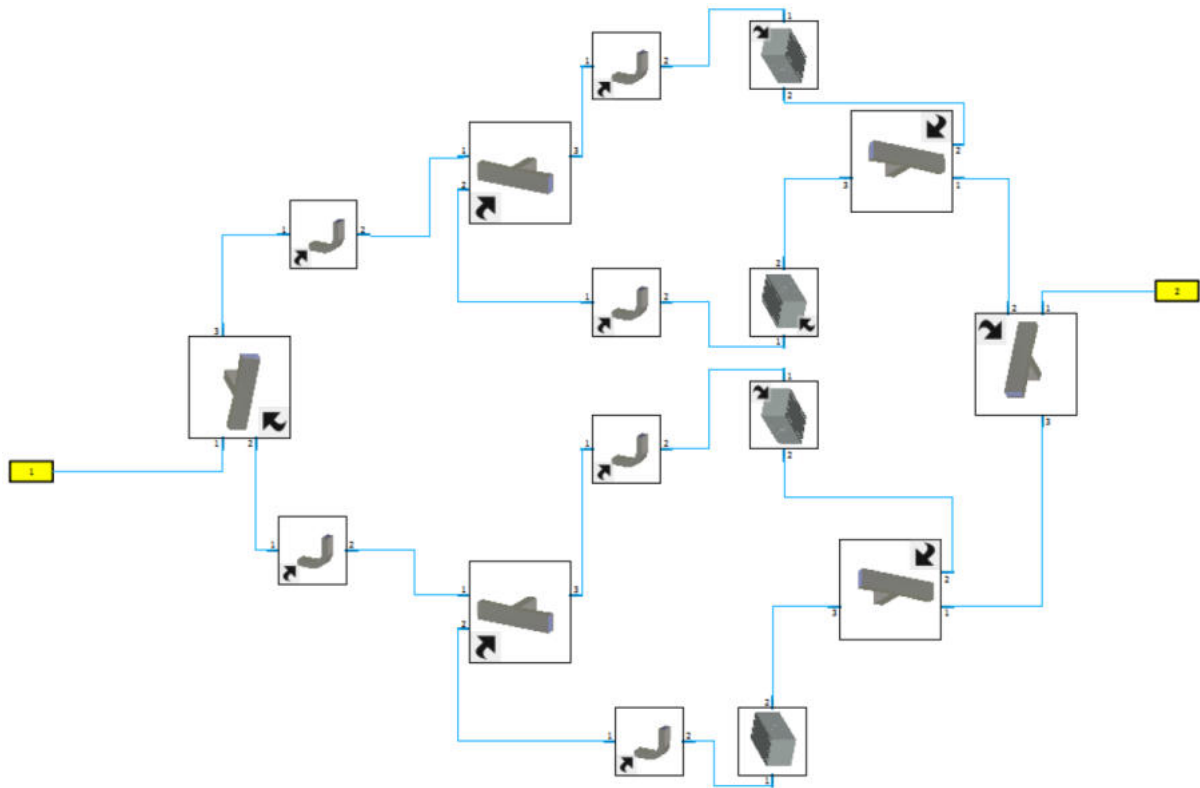


Figure 134 - 8 way power combiner block-based approach

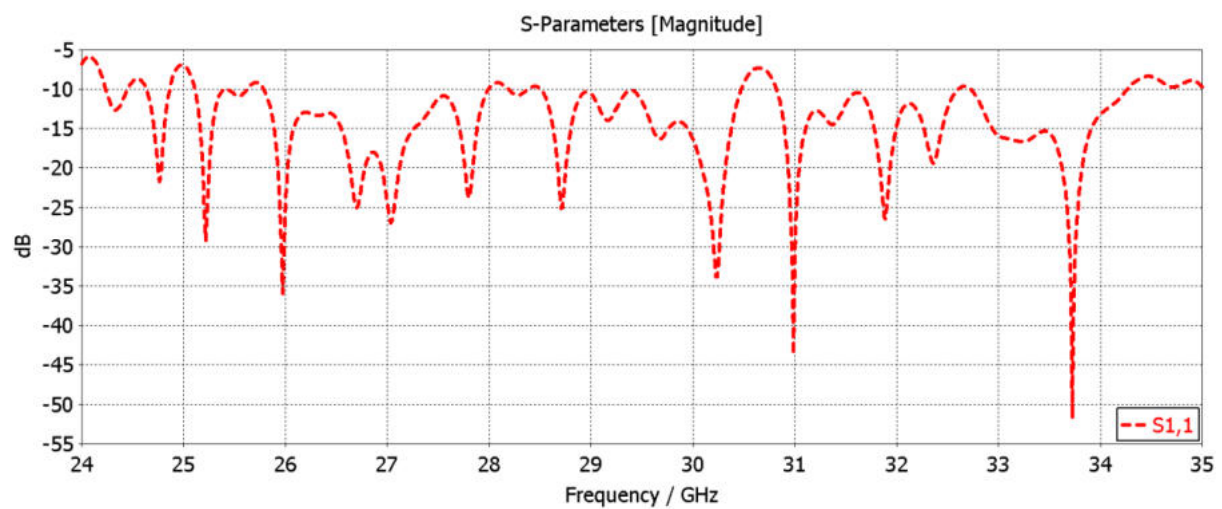


Figure 135 - Return loss

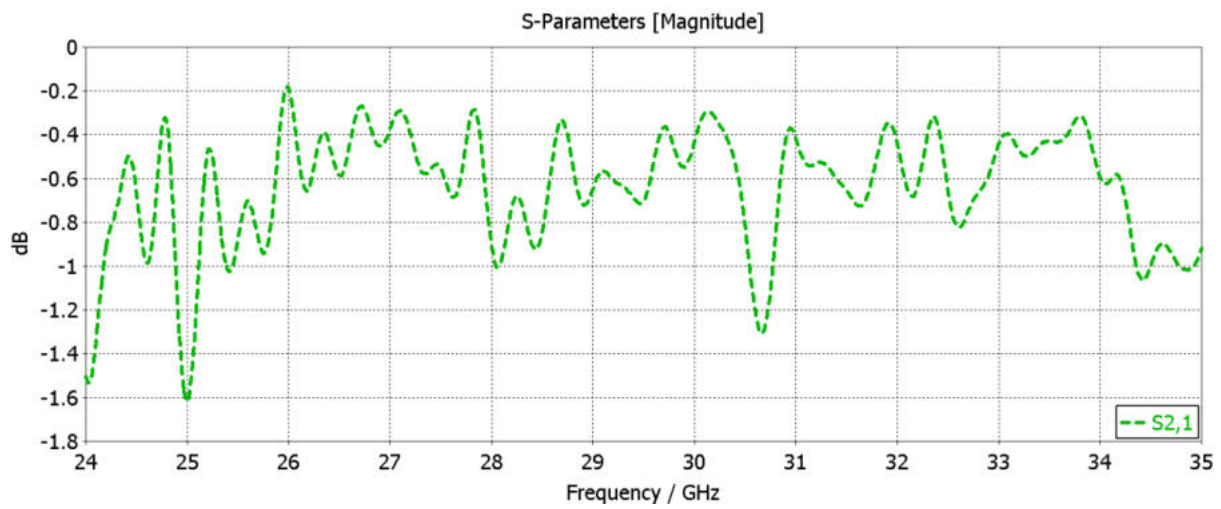


Figure 136 - Insertion loss from input port

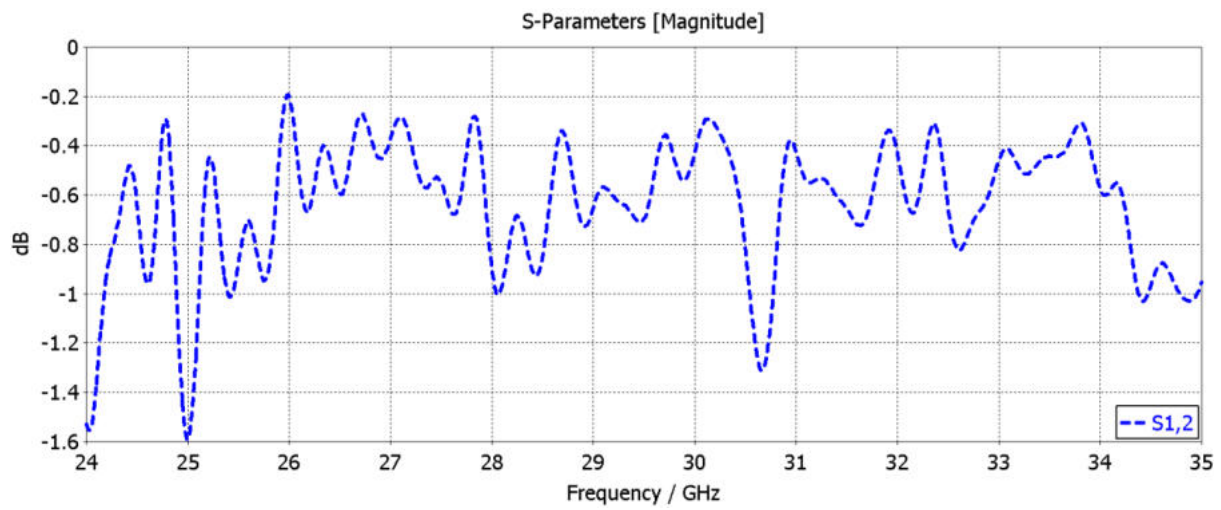


Figure 137 - Insertion loss from output port

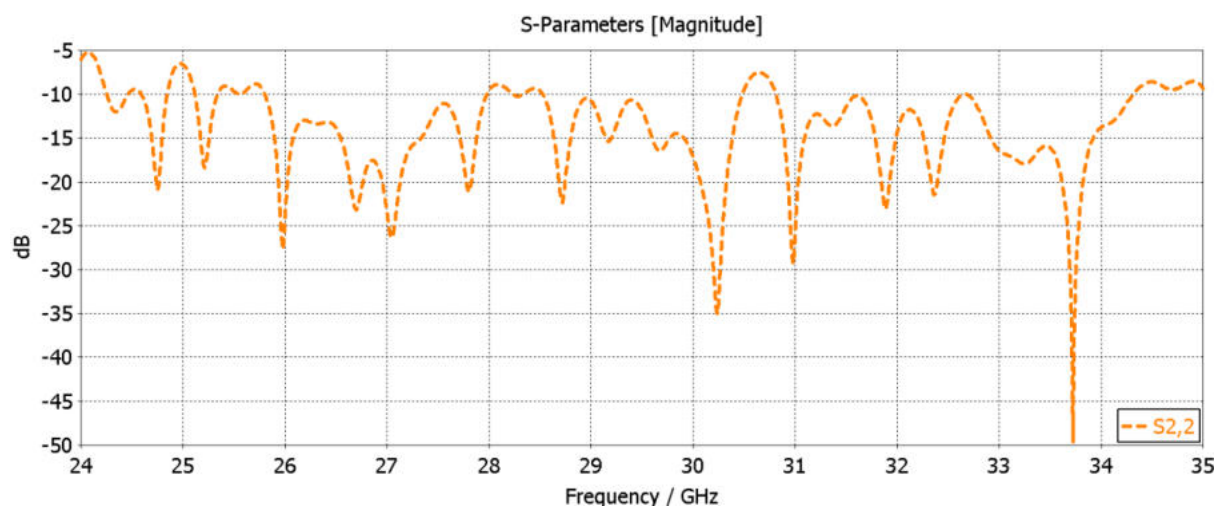


Figure 138 - Output Return loss

It should be noted that the block-based simulation methodology produces results that are almost equal to those obtained from full 3D single-model simulations. Each block is fully characterized by its scattering matrix (S-parameters), reducing block interconnection to a purely mathematical matrix multiplication operation. Individual blocks are linked to their respective low-mesh-count 3D sub models, ensuring computational efficiency.

This approach conclusively demonstrates the scalability of the waveguide-to-microstrip division/combination structure, which offers significant advantages in terms of physical footprint and power handling capability. Waveguide structures exhibit exceptional power handling—on the order of 100 kW under appropriate conditions including internal dielectric properties, operating frequency, and guide dimensions (a and b parameters). In Table 11 the comparison between traditional approach vs block-based simulation is presented.

xx	Block Simulation	Full 3D Simulation
Accuracy	Identical S-parameters	Reference (baseline)
Mesh cells/block	Low (<<1M)	Very high (>10M for N>4)
Simulation time	Minutes	Days
Power handling	100 kW potential	Same (physics-limited)

Table 11 - Comparison between traditional approach vs block-based simulation

The number of division and combination paths can be very high since the waveguide can withstand substantial power levels at the output port.

7.4 Physical Parameters and Constants

You may start with the standard physical dimensions of the WR-28 waveguide and the dielectric constants (air).

- Waveguide Dimensions (a, b):
 - $a = 7.112 \text{ mm} = 0.007112 \text{ m}$
 - $b = 3.556 \text{ mm} = 0.003556 \text{ m}$
- Breakdown Field (E_{br}): $3 \times 10^6 \text{ V/m}$ (30 kV/cm, typical value for dry air).
- Free Space Impedance (Z_0): $\approx 377\Omega$.

2. Cutoff Frequency Calculation (f_c)

The frequency below which the fundamental TE_{10} mode does not propagate.

$$f_c = \frac{c}{2a} = \frac{3 \times 10^8}{2 \times 0.007112} \approx 21.08 \text{ GHz}$$

3. Wave Impedance Factor Calculation

At 30 GHz, the wave does not travel as it would in free space but "bounces" between the walls. This effect is described by the term $\sqrt{1 - (f_c/f)^2}$.

$$\text{Factor} = \sqrt{1 - \left(\frac{21.08}{30}\right)^2} = \sqrt{1 - 0.493} = \sqrt{0.507} \approx 0.712$$

This number (0.712) indicates that the wave impedance Z_{TE} is higher than Z_0 ($Z_{TE} = Z_0/0.712 \approx 529\Omega$), which reduces the transportable power for the same maximum electric field.

4. Theoretical Maximum Power Calculation (P_{max})

The formula used for peak power before arcing (breakdown) in the TE_{10} mode is:

$$P_{max} = \frac{a \cdot b \cdot E_{br}^2}{4 \cdot Z_0} \times \sqrt{1 - \left(\frac{f_c}{f}\right)^2}$$

Substituting the values:

$$1. \text{ Geometric Term: } \frac{0.007112 \cdot 0.003556 \cdot (3 \cdot 10^6)^2}{4 \cdot 377} \approx 151,000 \text{ Watts}$$

(This would be the power if $f = \infty$)

$$2. \text{ Frequency Correction: } 151,000 \times 0.712 \approx 107,500 \text{ Watts}$$

Final Theoretical Result: 107.5 kW

The calculation above assumes ideal conditions like

- Perfect geometry (no surface roughness).
- Dry and clean air (no humidity or dust).
- VSWR = 1.0 (no reflections).

In the reality, assuming:

- $VSWR = 1.5$
- Safety: A factor of $1/4$ or $1/10$ of the theoretical value

You may get: $107.5 \text{ kW} \times 0.1 \approx 10 \text{ kW}$.

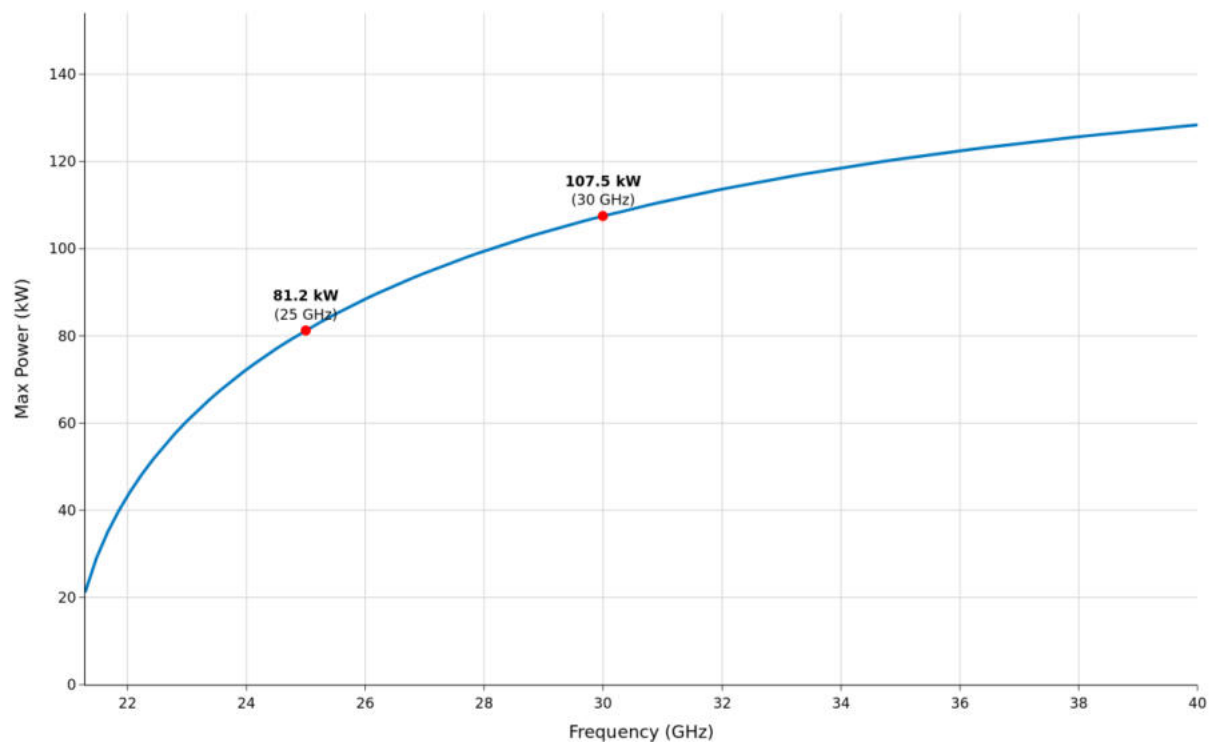


Figure 139 - WR 28 Power Handling

7.5 Fabricated Power Combiner

After demonstrating the scalability of the rectangular waveguide-to-microstrip power combiner, we focused on the fabrication of the two-way structure.

The fabrication of the prototype was carried out at Lancaster University, UK, where the structure was manufactured using precision machining techniques suitable for rectangular waveguide and microstrip such as the performance testing phase, where the fabricated combiner was characterized in terms of scattering parameters. The following images show the different stages of the manufacturing process. In Figures below, is shown the input and output ports of the assembly model.

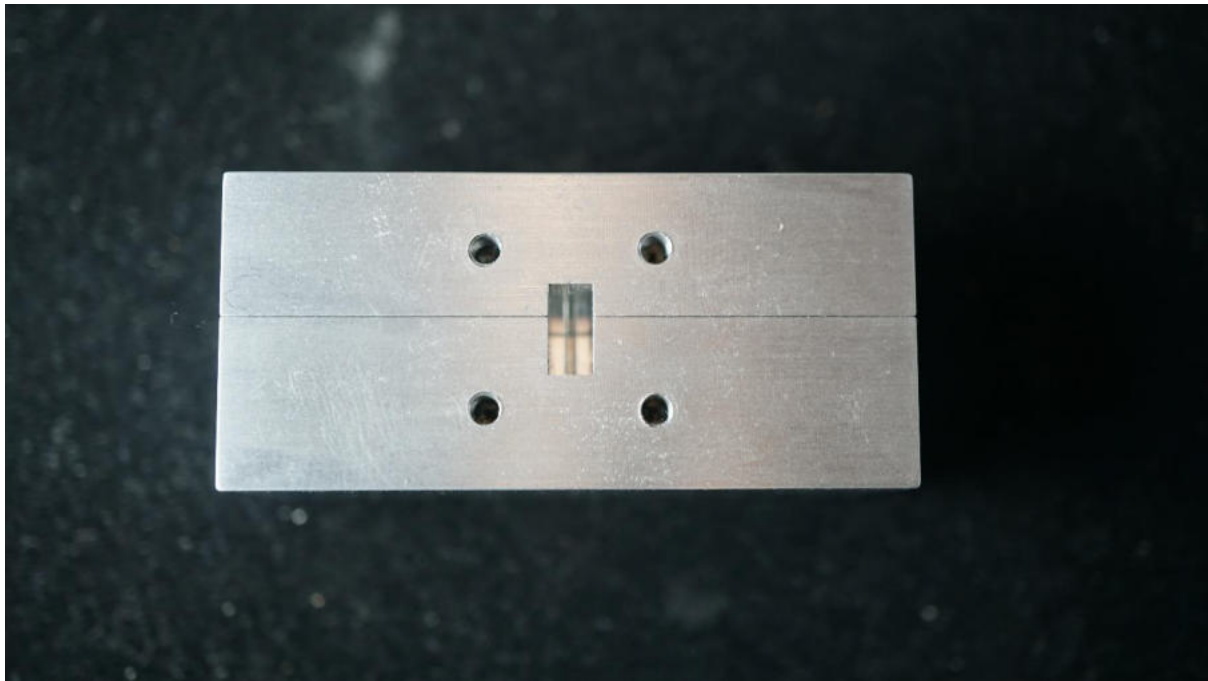


Figure 140 - Input Waveguide port (Front side) – Lancaster University



Figure 141 - Output Waveguide port (Front side) – Lancaster University

The most critical parts are the subassembly containing the T-junction with the central septum and the other subassembly corresponding to the 90-degree rotated waveguide-to-waveguide transition, where the microstrip is located.

The rotated waveguide transition represents a delicate part of the device, since it provides the interface between orthogonal waveguide sections while preserving field continuity and minimizing parasitic effects. For this reason, both subassemblies required careful fabrication and precise mechanical integration.

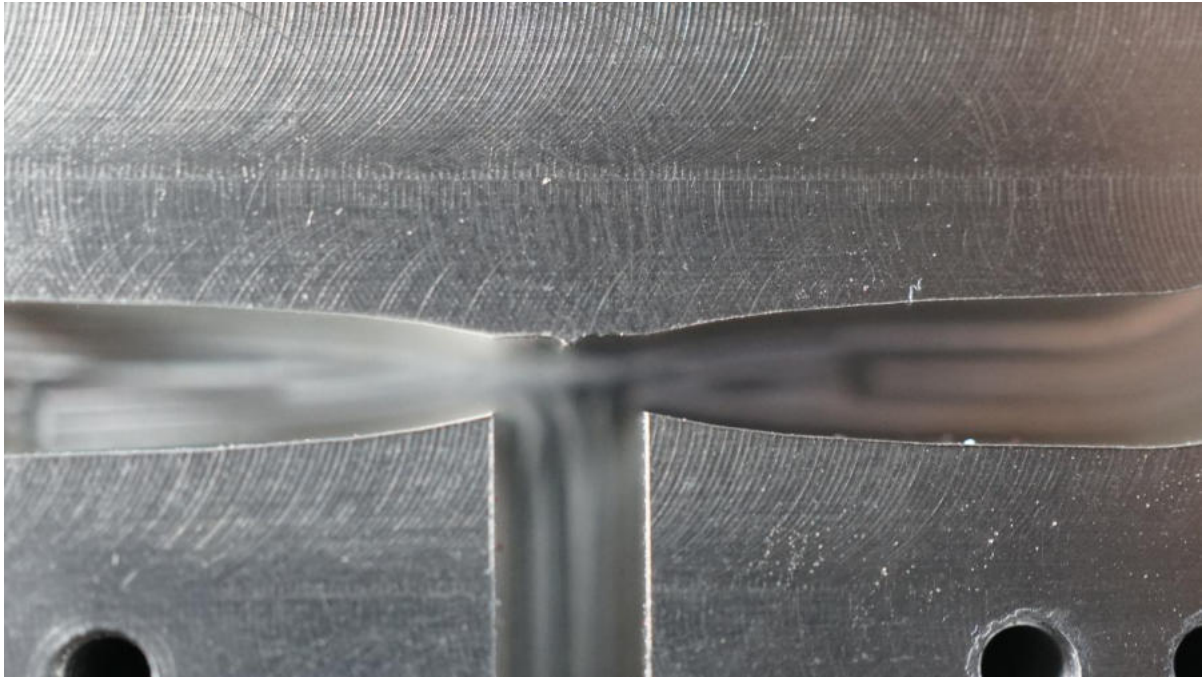


Figure 142 – Lancaster University proto (1)

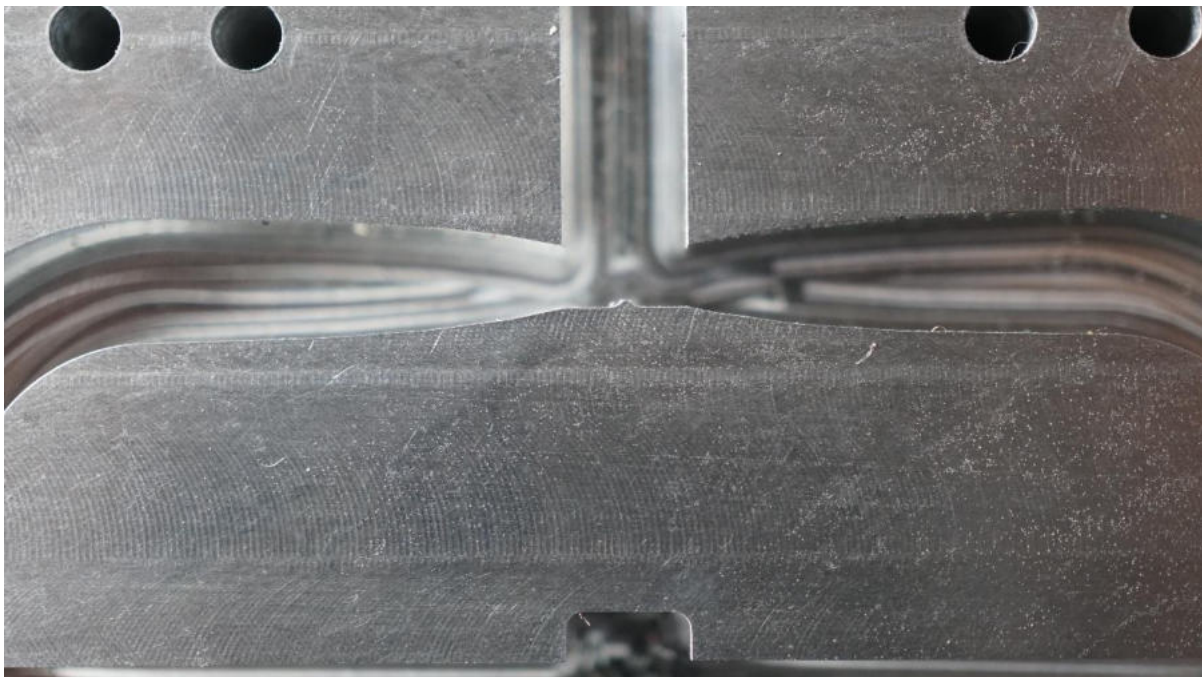


Figure 143 – Lancaster University proto (2)

The rotated waveguide transition represents a delicate part of the device, since it provides the interface between orthogonal waveguide sections while preserving field continuity and minimizing parasitic effects. For this reason, both subassemblies required careful fabrication and precise mechanical integration.

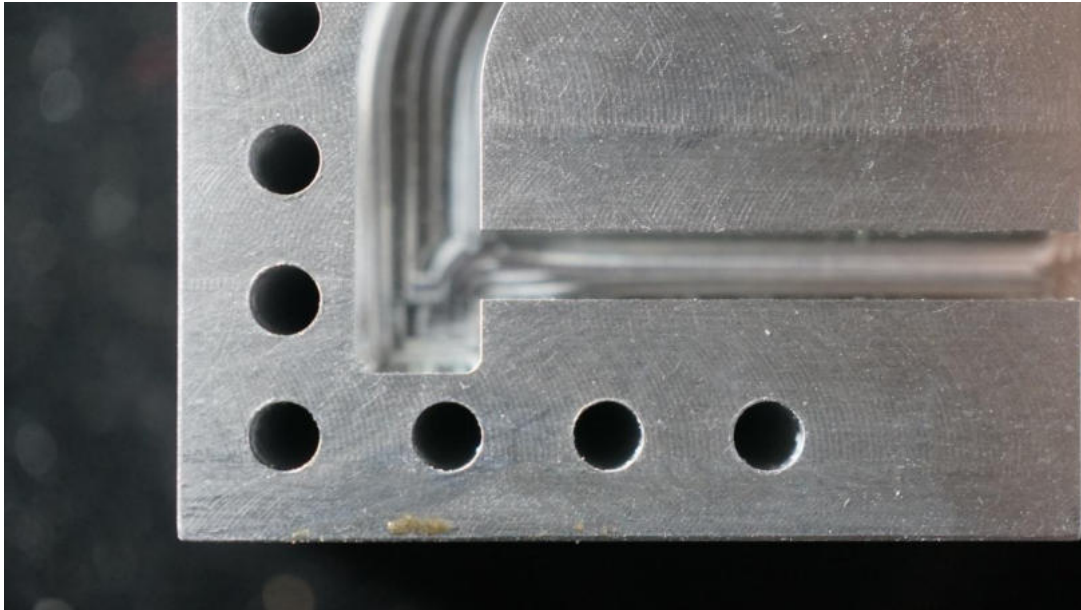


Figure 144 - Waveguide transition (Top View) – Lancaster University proto (3)



Figure 145 – Lancaster University proto (4)

Another subassembly that requires particular attention is the combining section, where the two microstrip lines are inserted. In this region, additive mode recombination takes place, whereby the field lines add in a constructive way and the amplified signal is propagated toward the output port.

As can be observed, several holes have been arranged around the perimeter of the model. These holes are designed to accommodate metal alignment rods that mechanically couple the top and bottom sizes of the assembly model.

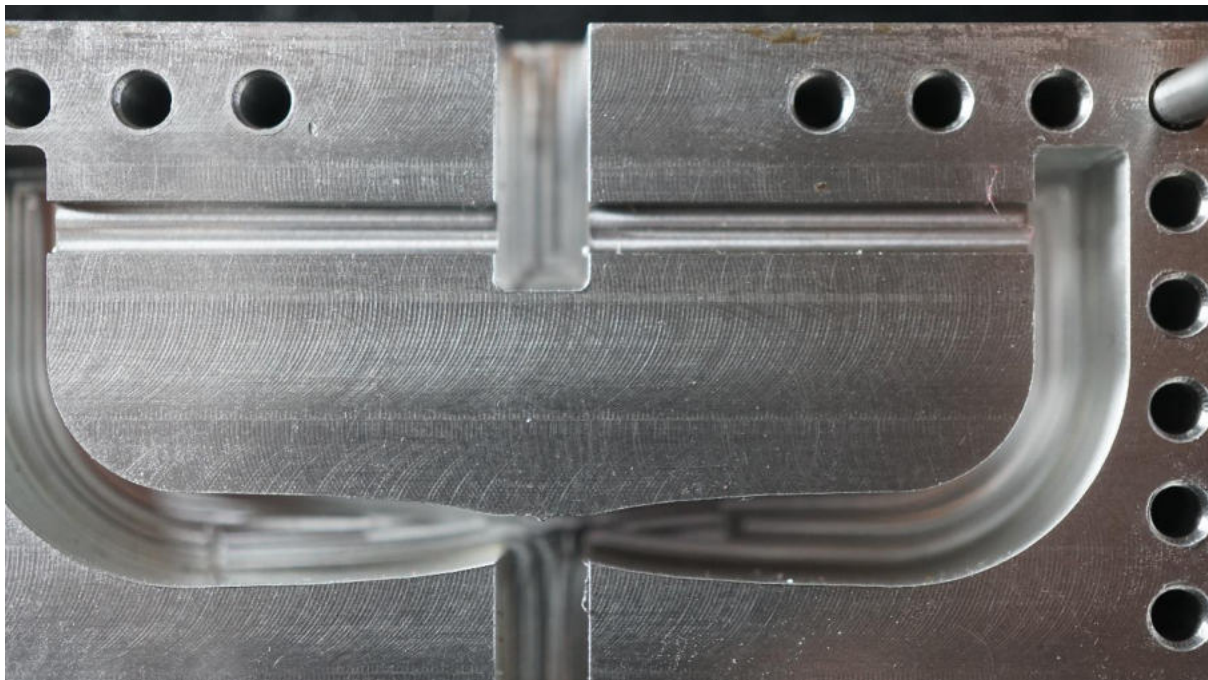


Figure 146 – Lancaster University proto (5)

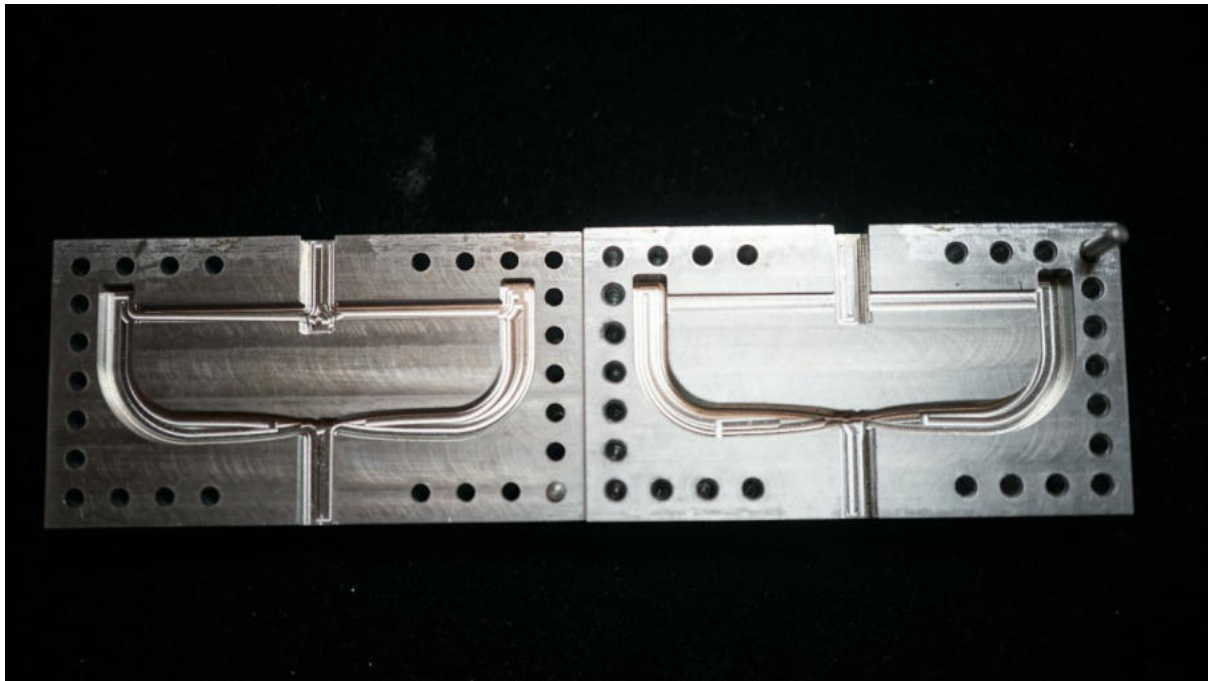


Figure 147 – Lancaster University proto (6)

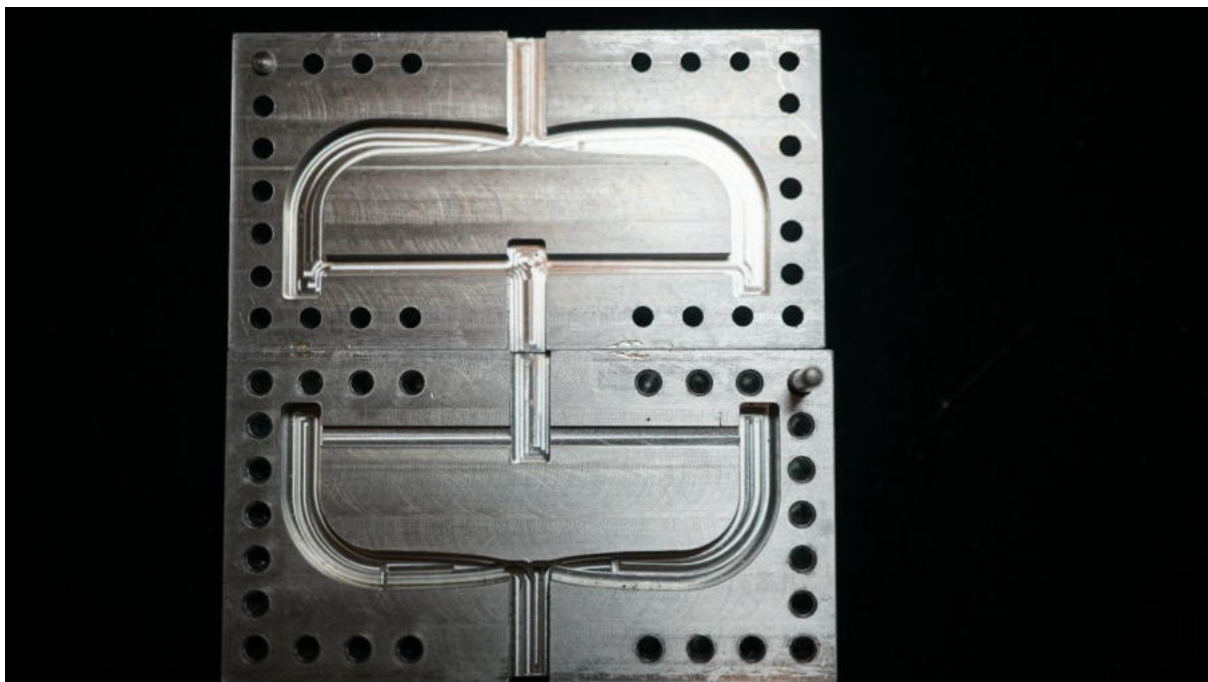


Figure 148 – Lancaster University proto (7)

The use of these alignment pins ensures accurate positioning of the two main blocks, minimizing misalignment errors that could otherwise disturb the field distribution and degrade the matching and isolation performance of the combiner.

Therefore, the alignment rods play an important role in order to guarantee the correct electromagnetic behaviour of the entire structure during testing. In Figures below, the assembly model under the microscope is shown.

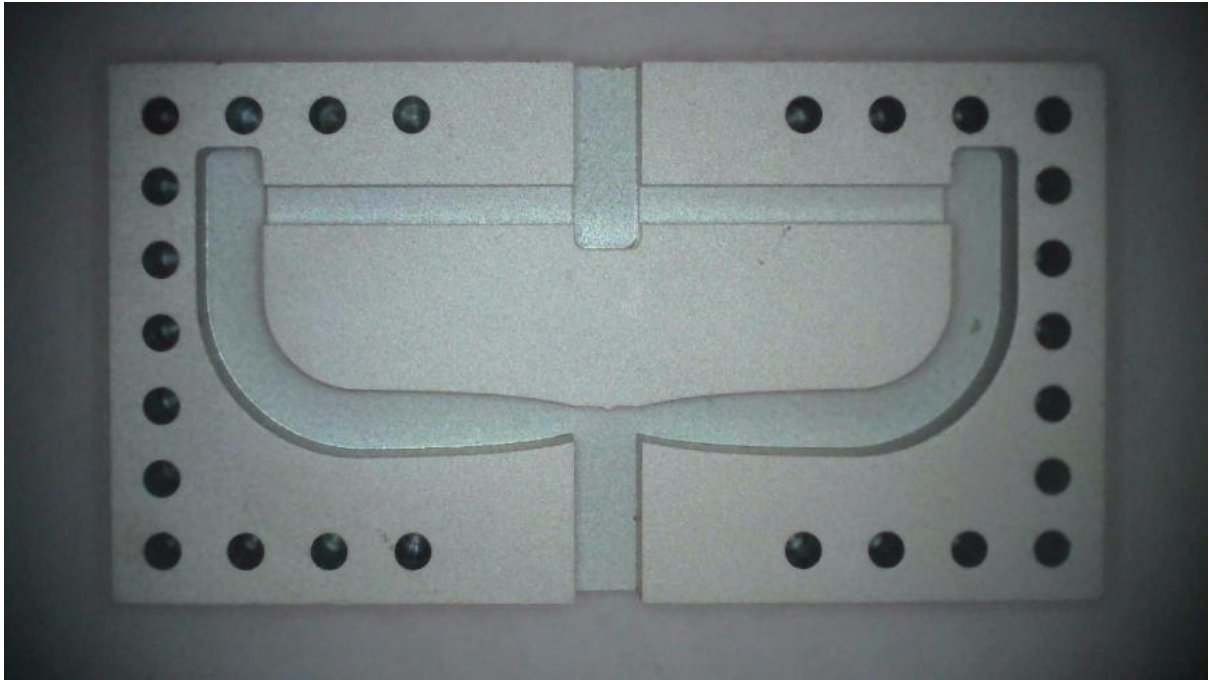


Figure 149 – Lancaster University proto (8)



Figure 150 – Lancaster University proto (9)

Once the model has been assembled, as shown in Figure 151, the characterization phase can begin. The structure proves to be very compact in terms of length, width, and height, with overall dimensions of $54.2\text{ mm} \times 29\text{ mm} \times 24.1\text{ mm}$. This compact size represents a significant advantage for integration into modular microwave systems and for applications where volume and weight constraints are critical. The overall size of the power combiner is shown.

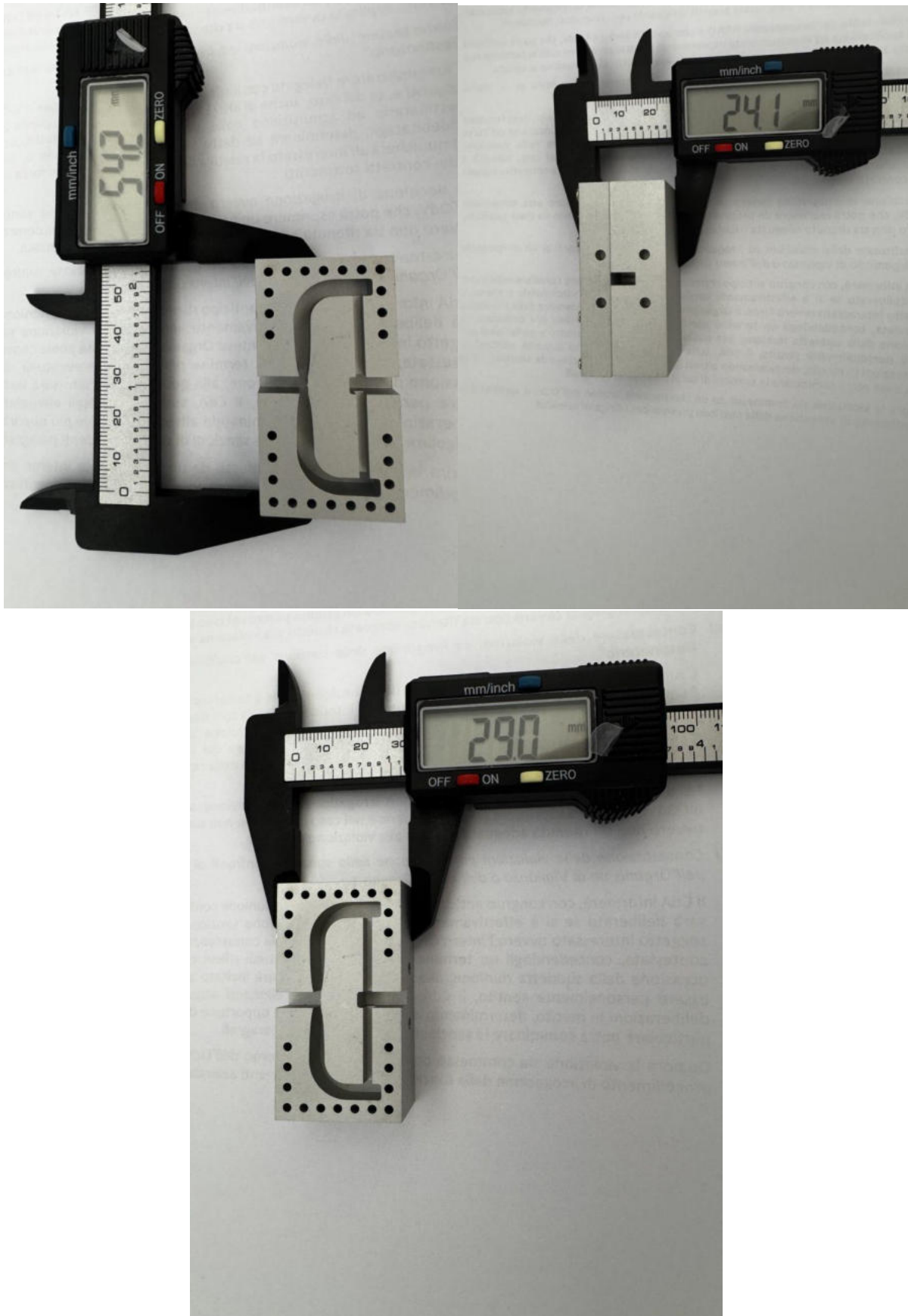


Figure 151 - Two way power combiner assembly mode

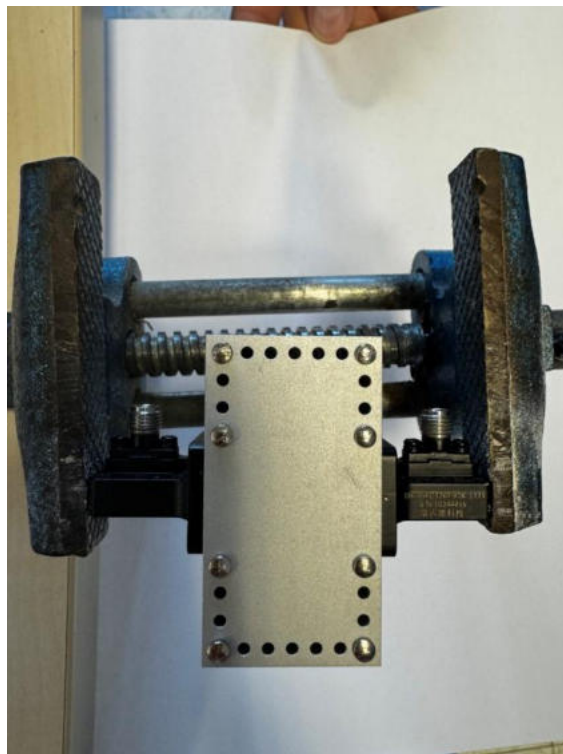
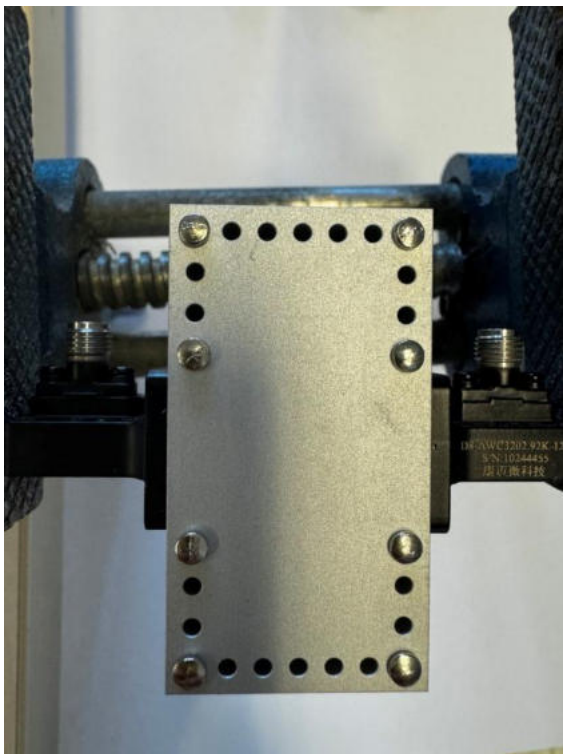
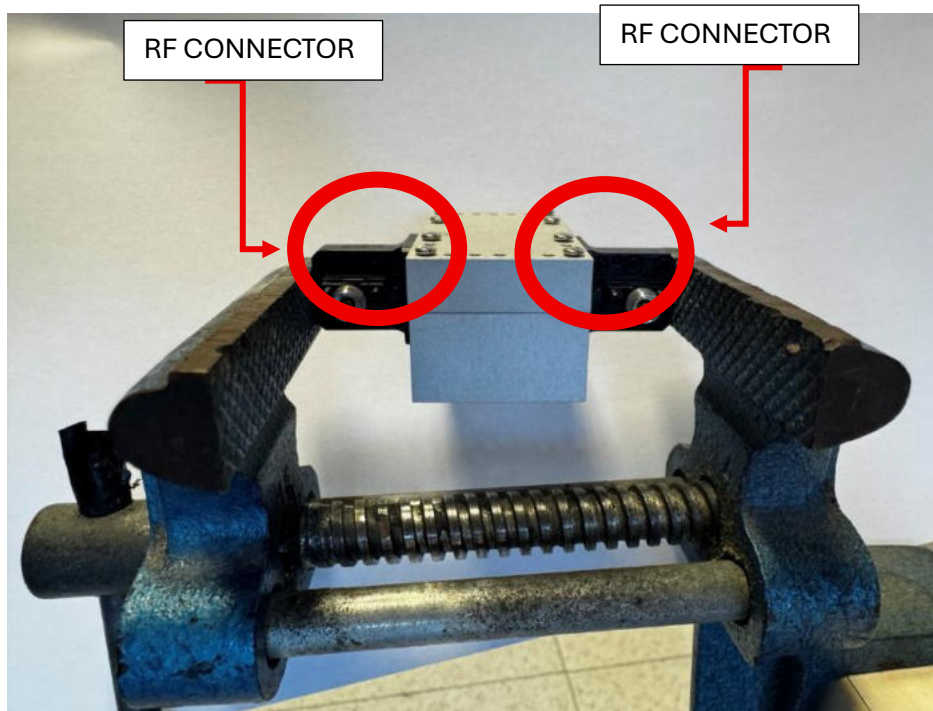


Figure 152 - Fabricated PC - overview

WR-28 Square Cover Flange to 2.92mm Female Waveguide to Coax Adapter Operating from 26.5 GHz to 40 GHz

CAD Model

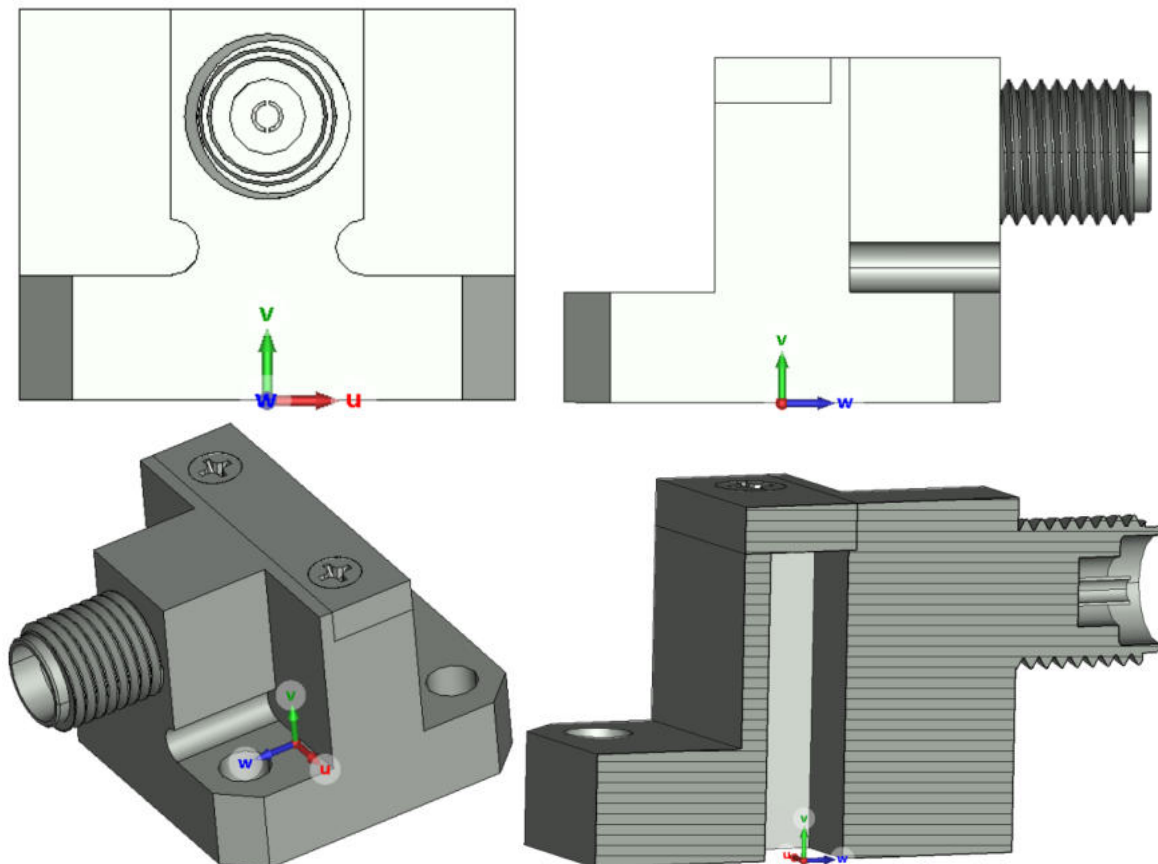


Figure 153 - Waveguide to coaxial transition

Description

The PE9826 is a waveguide to coaxial adapter operating in the Ka band with a 26.5 to 40 GHz frequency range. This adapter offers a WR-28 waveguide interface size coupled with a precision tolerance square cover flange. The PE9826 is constructed of aluminum to ensure durability and repeatable RF performance. The coaxial connector offered by this adapter uses a 2.92mm

geometry endorsed by International Electrotechnical Commission (IEC). The 2.92mm female connector mates with commercially available SMA and 3.5mm connectors. The package is REACH and RoHS compliant.

Configuration

Waveguide Size: WR-28

Flange: Square Cover

RF Connector: 2.92mm Female

Impedance: 50 Ohms

Body Geometry: Right Angle

Description	Minimum	Typical	Maximum	Units
Frequency Range	26.5		40	GHz
VSWR			1.4:1	
Insertion Loss			0.5	dB
Input Power (Peak)			20	Watts

Mechanical Specifications

Size

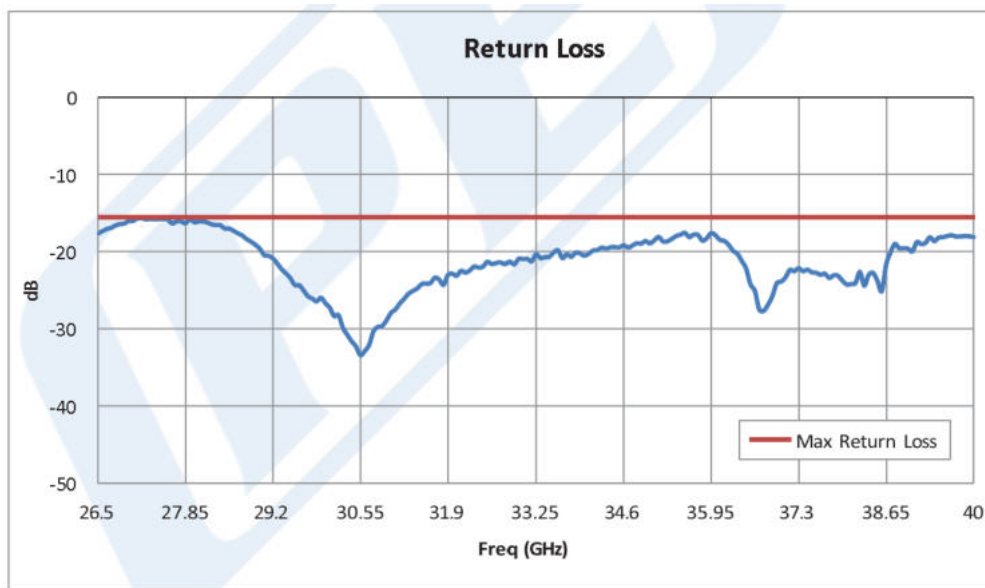
Length	1.01 in [25.65 mm]
Width	0.59 in [14.99 mm]
Height	0.75 in [19.05 mm]
Weight	0.042 lbs [19.05 g]

RF Connector

Connector Type	2.92mm Female
Body Material and Plating	Stainless Steel

Waveguide Interface

Waveguide Type	WR-28
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PE9826 CAD Drawing

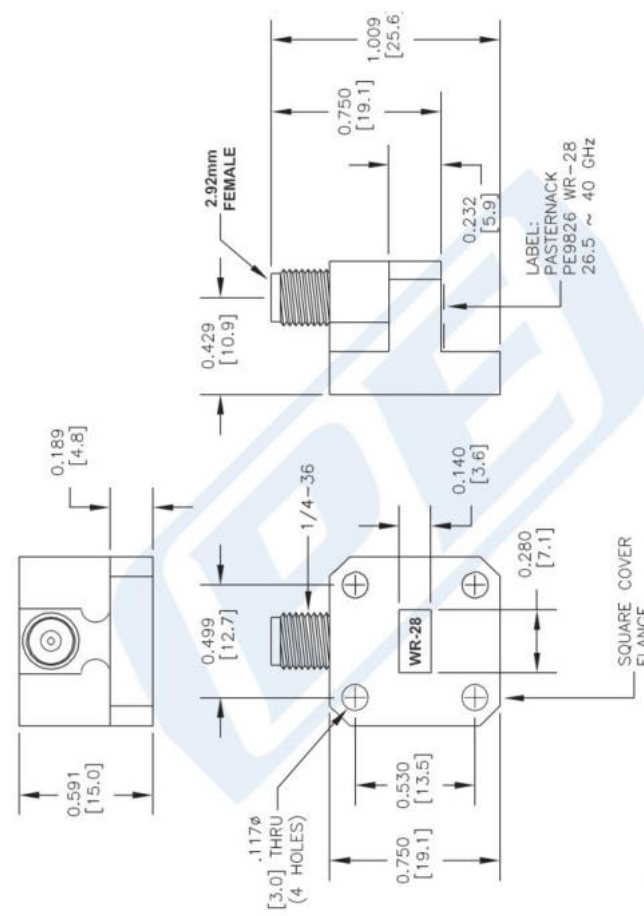


Figure 154 - Mechanical overview

Features

- Ka Band, 26.5 to 40 GHz Frequency Range
- WR-28 Waveguide Interface with Square Cover Flange
- 2.92mm Female Coaxial Interface
- Comprehensive waveguide offerings also include E-Bends, H-Bends, Sections, Filters, Terminations, Couplers, and more.

Description

The PE9826 is a waveguide to coaxial adapter operating in the Ka band with a 26.5 to 40 GHz frequency range. This adapter offers a WR-28 waveguide interface size coupled with a precision tolerance square cover flange. The PE9826 is constructed of aluminum to ensure durability and repeatable RF performance. The coaxial connector offered by this adapter uses a 2.92mm geometry endorsed by International Electrotechnical Commission (IEC). The 2.92mm female connector mates with commercially available SMA and 3.5mm connectors. The package is REACH and RoHS compliant.

Configuration

Waveguide Size	WR-28
Flange	Square Cover
RF Connector	2.92mm Female
Impedance	50 Ohms
Body Geometry	Right Angle

Electrical Specifications

Description	Minimum	Typical	Maximum	Units
Frequency Range	26.5		40	GHz
VSWR			1.4:1	
Insertion Loss			0.5	dB
Input Power (Peak)			20	Watts

Mechanical Specifications

Size	
Length	1.01 in [25.65 mm]
Width	0.59 in [14.99 mm]
Height	0.75 in [19.05 mm]
Weight	0.042 lbs [19.05 g]
RF Connector	
Connector Type	2.92mm Female
Body Material and Plating	Stainless Steel
Waveguide Interface	
Waveguide Type	WR-28

7.6 Characterization and Measurement results

This section presents the characterization of the power combiner introduced above. Figure below reports its performance in terms of return loss and insertion loss. The measurements were carried out using a Rohde & Schwarz vector network analyzer. The resulting curves highlight the electrical behavior of the

structure over the investigated frequency range. In Figure 155, the measured Return loss of the PC is shown.

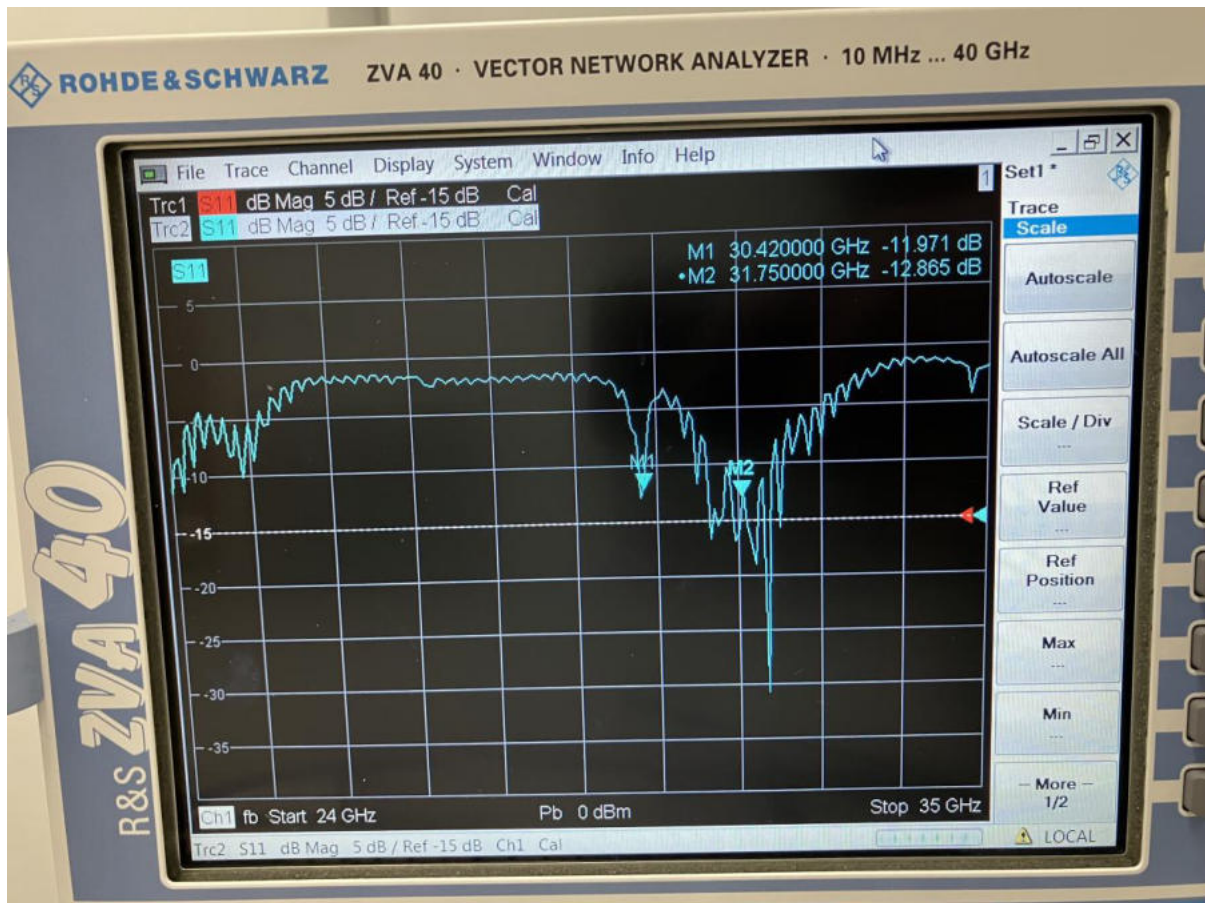


Figure 155 - Measured S11 with Rohde & Schwarz VNA

First, it can be observed that within the 24–35 GHz observation window, the structure is poorly matched below 30 GHz, where the return loss is approximately 2.5 dB. A better impedance matching condition is then achieved at higher frequencies, with a resonance of about 11.97 dB at 30.42 GHz. In addition, the structure appears to be reasonably matched over an approximate 1 GHz

bandwidth around 31.75 GHz, where S_{11} reaches 13 dB. The measured Insertion loss of the PC is shown in Figure 156.

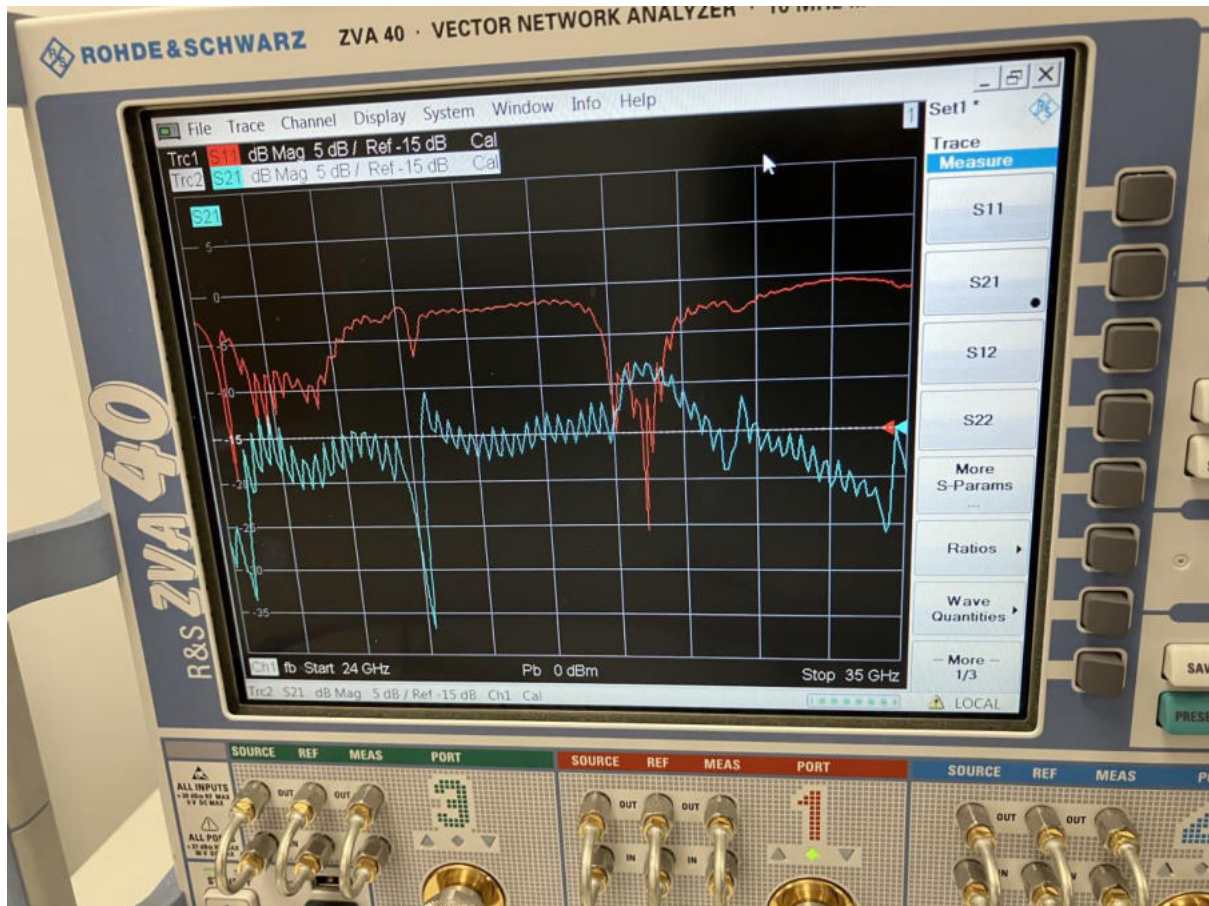


Figure 156 - Measured S_{11} and S_{21} with Rohde & Schwarz VNA

Meanwhile, by observing the measured S_{21} curve, it can be seen that the input and output ports of the structure are effectively isolated. In other words, the insertion loss over the 24–35 GHz band is very high, approximately 15 dB.

The measurements clearly highlight the difficulty of fabricating structures of such small dimensions. Several factors may account for the mismatching between the

simulated and measured performance. These include the surface roughness of the aluminum substrate, as well as imperfections in the microstrip, which plays an important role in the proper conversion of the propagating mode. Another possible factor may be the accurate placement of the microstrips within the pocket such as the spacing of 440 μm at the final recombination stage.

8 Conclusions

This thesis presented the advantages of replacing well-known vacuum tubes with much more compact and efficient structures. In particular, several power divider and power combiner solutions operating in the K-Ka frequency band have been discussed. The high efficiency guarantees reduced heat generation, which can be easily managed in the proposed combining structures. These solutions can be employed in both civil and military applications, including radar and satellite systems, where high output power is required. The solutions proposed offer an excellent way to combine very high power level provided in applications in which there is a requirement of high power handling. For instance, by the time you have the requirement to transmit power radiating into the free space. In this case, the transmitting chain should be able not only to generate a huge amount of power but also to manage the power itself. The power generation is guaranteed by placing a Hpa (High power amplifier) in each way of the power divider section.

Once the input signal gets splitted and amplified, the power combination section handles and ensures a correct way the power flows through the output port. The presented work has the aim to keep the focus on the high power structures that guarantee high power handling such as high level of miniaturization and integration in complex transmitting and receive systems.

In the future, I will be involving on the design of a complete radar system operating at high frequencies in order to demonstrate the effectiveness of having a microwave product with a high power handling, high bandwidth, low losses and a satisfying thermal management.

8.1 Courses, PhD Programs, and Training Events

- Conferenza “17th United Conference on Millimetre Waves and Terahertz Technologies (UCMMT2024)”, tenutasi a Palermo dal 21 al 24 Agosto 2024.
- PhD School Electronics for Health, SIE Genova 2024 24-26 Giugno 2024
- “Fiber optics for healthcare sensing”, Prof Campopiano, Università di Napoli

- “Electronics for medical echographic systems”, Prof Tortoli, Università di Firenze
- “Ultra low quiescent PMIC for wearable-healthcare applications”, Ing. Masici, Analog devices
- “Optical Wireless Communication Systems for Biomedical Implants: Basics and Advances of Biotelemetry”, Prof. De Marcellis, Università dell’Aquila
- “Electronic Interfaces for Biomedical Applications”, Dr. Zompanti, University Hospital of Rome
- “Neuromorphic Technology for Robotics and Healthcare, from Circuits to Applications”, Dr. Bartolozzi, Italian Institute of Technology
- “MEMS Sensor featured with a Potentiostat for Biosensing”, Ing. Borgioli, STMicroelectronics
- “Ultrasonic energy harvesting and communication for implanted medical devices”, Prof. Grasso, Università di Catania
- “Microelectronics in nucleic acid sensors (RNA, DNA). Past, present and future prospects”, Ing. Bianchessi, STMicroelectronics
- “Connecting the humans, HMIs for wearable medical electronics”, Prof. Benatti, Università di Modena e Regio Emilia
- “Neural Stimulation Devices for Implantable and Wearable Applications”, Prof. Barbaro, Università di Cagliari
- “Biomedical Applications of Microwave and Millimeter-Wave Radars”, Dr. Caridllo, Università di Messina

- “Accelerate Digital Healthcare through Edge AI”, Ing. Gardella, NXP-Semiconductor
- “A power efficient device, tailored for MEMS-speaker based hearing aids”, Ing. Fusetto, Inventum
- “How high speed communications are playing a key role in modern medical analysis”, Ing. Deapoli, Synopsis.
- Seminario Keysight "EDA CONNECT TOUR 2024" - 23 gennaio, Roma

8.2 List of Publications

- P. Livreri, A. P. Sausa and L. Miano, "Design and Experimental Validation of a Compact Eight-Way Radial Coaxial Power Combiner for K/Ka-Band Applications," in IEEE Access, doi: 10.1109/ACCESS.2026.3713422.
- Livreri P, Miano L., Sausa A. P., and Di Mitri A., Design of a Compact Top-Bottom 2-way and 4-way Slotted Waveguide Power Combiner for Ka-Band

Spatial Applications. In UCMMT 2026 (The 19th IEEE United Conference on Millimetre Waves & Terahertz Technologies)

- Miano, L., Sausa, A.P., Gulotta, A., Livreri, P. (2025). Design of compact Novel Top-Bottom 8-way waveguide to microstrip Power Combiner with 90° phase shift mode converter for Ka-Band Spatial Application. In IEEE UCMMT 2025.
- Gulotta, A., Mendola, L., Miano, L., Soruri, M., Livreri, P. (2025). A 20 W High Gain, High Efficiency L/S-Band Power Amplifier for 5G Connected Cars. In IEEE UCMMT 2025.
- Gulotta A., Miano L., Muratore A., Livreri P. (2024). Design of a 8 way Ka-band Power Combiner using Stripline Transitions. In 2024 17th United Conference on Millimetre Waves and Terahertz Technologies, UCMMT 2024 (pp. 118-121). Institute of Electrical and Electronics Engineers Inc. [10.1109/UCMMT62975.2024.10737781].
- Gulotta A., Miano L., Livreri P. (2024). Design and Simulation of Butterfly Antenna in different Lattice Configurations. In 2024 17th United Conference on Millimetre Waves and Terahertz Technologies, UCMMT 2024 (pp. 207-210). Institute of Electrical and Electronics Engineers Inc. [10.1109/UCMMT62975.2024.10737814].
- Miano L., Gulotta A., Muratore A., Livreri P. (2024). Coaxial Power Combiner with Eight Radial Coaxial Output Ports for Ka-Band Applications. In 2024 17th United Conference on Millimetre Waves and Terahertz Technologies, UCMMT 2024 (pp. 122-125). Institute of Electrical and Electronics Engineers Inc. [10.1109/UCMMT62975.2024.10737806].

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- Gulotta, A., Mendola, L., Miano, L., Soruri, M., Livreri, P. (2025). A High-Gain, High-Efficiency GaN HEMT Power Amplifier for L/S-Band 5G Vehicle Connectivity. In 2025 20th International Conference on PhD Research in Microelectronics and Electronics, PRIME [10.1109/PRIME66228.2025.11203722].
- Accursio Gulotta, Luigi Miano, Patrizia Livreri, Alfonso Farina (2022). Radar Imager for Mars' Subsurface Exploration (RIMFAX). AEIT(3/4), 64-69.

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