

X-Band 3D-Printed Dielectric-Loaded Out-of-Band WR62 Waveguide Transmission Line for Reduced Size and Weight

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ABSTRACT: Waveguide transmission lines are known for their low insertion losses, particularly at higher GHz frequencies. However, they tend to be larger and bulkier than other transmission lines. This study investigates using dielectric-loaded waveguides to reduce the overall size and volume of waveguide-routed systems with the dielectric properties of 3D-printed plastics. Compared to other studies that use dielectric loading for size reduction, this study aims to reduce size, volume, and weight while maintaining relatively low losses. The approach is to use low-loss tangent dielectrics with a suitable dielectric constant responsible for reducing size, volume, and weight. Among 3D-printed plastics, Cyclic Olefin Copolymer (COC) is the most suitable based on its dielectric constant and loss tangent in the order of conventional dielectrics such as PTFE. The measured insertion loss of the 3D-printed dielectric-loaded waveguide was 0.22 dB, comparable to that of air-loaded WR90 waveguides of the same length (0.10 dB), while reducing the cross-sectional area by 46.35% and the weight by 26%. The advantage of 3D printing, dielectric-loaded waveguides is the ability to design less constrained waveguide profiles and intricate routing of waveguide transmission lines due to more flexible fabrication capabilities of 3D printing. Combining the benefits of 3D printing and the dielectric properties of COC, this study demonstrates the use of smaller waveguides in a higher frequency band that would be out of band in their air-loaded medium. By loading the smaller out-of-band waveguide with dielectric, the result shows that longer-wavelength signals, which would not be able to propagate without dielectric loading, can now propagate with low insertion loss. The waveguide chain is also miniaturized with a reduced size and overall volume, which would not have been possible with conventional air-loaded waveguides for a given frequency band, as it propagates below the cutoff frequency.

1. INTRODUCTION

Waveguides have been widely used because of their low insertion-loss characteristics, especially when the length of the transmission line becomes long [1–3]. The disadvantage of waveguides for Radio Frequency (RF) propagation is the size and volume of the transmission line, which is highly dependent on the operating frequency. This means that using waveguides for relatively low frequencies would be impractical, as they have the same order of magnitude as their wavelength, making them heavy and expensive [4, 5].

In this study, the frequency of interest is in the X-band, where WR90 standard waveguides are recommended for this frequency band (8.2 GHz to 12.4 GHz). To demonstrate out-of-band operation in a different waveguide standard, WR62 was used to reduce size and volume. Typically, the WR62 waveguide standard is used for the Ku-band (12.4 GHz to 18 GHz), which is completely outside the intended frequency band of use. Dielectric loading allows a smaller out-of-band waveguide to accommodate a lower frequency for propagation. Fundamentally, dielectric loading in waveguides lowers the cutoff frequency at the fundamental TE_{10} mode compared to that without loading, which this study aims to use to miniaturize waveguide chains.

There has been work done on the dielectric loading of waveguides for various applications, including phase shifters [6], dielectric-loaded RF chokes [7], bandpass filters [8–10], mode converters [11–13], and inserting dielectrics into waveguides to alter their mode dispersion and propagation characteristics [14–17]. However, the authors intend to fill the waveguide cross-section with a dielectric to support RF propagation in a completely out-of-band waveguide operation. Thus, the authors propose an impedance-matching taper, which varies both the metallic waveguide and dielectric-loaded inserts, to transition the WR90 air-loaded waveguide to the WR62 dielectric-loaded waveguide. This essentially allows a smaller waveguide to support an RF signal that would otherwise require a larger waveguide. Thus, supporting the out-of-band signal to operate with a smaller waveguide allows savings in the size, volume, and weight of the overall chain of waveguide transmission lines. This is especially important when the waveguide chain becomes long due to the waveguide routing of complex RF systems, which motivates this study.

Table 1 compares other studies that use dielectric loading for size reduction [18–23]. There are also works using dielectric loading as a non-radiative dielectric waveguide and transitions [24–32], which is similar to the authors' application. The authors, however, propose a different concept of a non-radiative dielectric waveguide with low insertion loss and wide bandwidth, where the dielectric loading is bounded by metal-

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TABLE 1. Comparison with other works.

	Reference	Technology	Frequency Range (GHz)	Bandwidth (GHz)	Fractional Bandwidth	Wavelength λ_0^* (mm)	Achieved Size	Insertion Loss (dB)	% Size Reduction λ^2
Size Reduction Comparison	[18]	Floral Foam Loaded Waveguide Resonator	5.22 – 5.31	0.095	1.8%	57.6	$0.618\lambda_0$	0.24	$(31.4\%)^2 = 9.85\%$
	[19]	Anisotropic Metamaterial Liner in Waveguide	3.69 – 3.82	0.12	3.3%~	79.7	$0.188\lambda_0$	15.0	60%#
			3.55 – 3.80	0.25	7.0%~	81.4	$0.184\lambda_0$		
	[20]	Composite Coplanar Waveguide	0.42 – 1.57	0.58	57.4%~	299.8	$0.036\lambda_0$	3.8	86.9%#
	[21]	Dielectric-Loaded Filter	10.88 – 11.12	0.24~	2.2%~	27.2	$1.488\lambda_0$	1.0	$(50\%)^{2/3} = 63\%$
	[22]	Loaded Coplanar Waveguide	2.11 – 2.40	0.14	6.3%~	132.6	$0.31\lambda_0$	1.93	50%#
	[23]	Half-Mode Substrate-Integrated Waveguide	2.12 – 2.57	0.44	19.0%~	127.5	$0.058\lambda_0$	0.4	Not Given
			3.49 – 3.90	0.40	11.0%~	81.0	$0.091\lambda_0$	0.5	
			5.84 – 6.24	0.39	7.0%~	53.0	$0.140\lambda_0$	0.9	
Broadband Transmission Line Comparison	[24]	3D-Printed Chalk Non-Radiative Dielectric-Guide	11 – 18	7.0	48.2%	20.6	$0.58\lambda_0$	3.5	Not Given
	[25]	3D-Printed Dielectric-Loaded Transition	58.0 – 67.5	9.5	15.1%	4.7	$0.70\lambda_0$	1.0	Not Given
	[26]	Hollow Rectangular Dielectric Waveguide	220 – 294	74.0	28.8%	1.1	$1.20\lambda_0$	2.6	Not Given
	[27]	Microstrip Line to Dielectric Transition	85.1 – 110	24.9	25.5%	3.0	$2.11\lambda_0$	0.49 – 1.5	Not Given
	[28]	3D-Printed Dielectric Waveguide Transition	24 – 26	2.0	8.0%	12.0	$0.85\lambda_0$	2.9	Not Given
			75 – 85	10.0	12.5%	3.7	$2.78\lambda_0$	4.3	
	[29]	Microstrip Line to Dielectric Transition	79 – 108	29	31.0%	3.2	$2.02\lambda_0$	0.52 – 2.36	Not Given
	[30]	Perpendicular Microstrip Line to Dielectric Transition	80.9 – 108.6	27.7	29.2%	3.1	$2.05\lambda_0$	0.42 – 1.72	Not Given
	[31]	Non-Radiative Dielectric H-Guide	33 – 37	4.0	11.4%	8.5	$1.86\lambda_0$	1.8 – 3.1	Not Given
	[32]	Non-Radiative Dielectric Bend	93 – 98	5.0	5.2%	3.1	$2.60\lambda_0$	0.18	Not Given
	This Work	Dielectric-Loaded Waveguide	8.2 – 12.4	4.2	40.8%	29.1	$0.54\lambda_0$	0.14 – 0.22	46.35%^

* The wavelength λ_0 is based on the center frequency of the frequency range. Fractional Bandwidth also references the center frequency.

Percentage of size reduction was taken directly from the reported value in the reference.

~ Referenced from its reported value.

^ This work reduces the cross-sectional area from WR90 (22.86 mm $\times$ 10.16 mm) to WR62 (15.79 mm $\times$ 7.89 mm), giving a size reduction of 46.35%.

lic rectangular waveguide walls. Concurrently, size and weight reduction was demonstrated. This fulfils the authors' requirement that all three criteria (large bandwidth with low loss and size reduction) be optimal simultaneously, and other works perform two of the three, which is the novelty of the authors' work.

2. WORKING PRINCIPLE

The working principle relies on loading a smaller out-of-band rectangular waveguide with a dielectric material to lower its cutoff frequency, allowing a smaller waveguide to support lower frequencies that it would be otherwise unable to. The X-band RF wave was fed from an air-loaded WR90 waveguide and slowly transitioned into a smaller out-of-band WR62 waveguide loaded with dielectric. This transition smoothly converts the air-loaded WR90 TE₁₀ mode to the dielectric-loaded WR62 TE₁₀ mode by matching the wave impedance at the interface. To achieve this, a taper needs to be designed to

match the impedance to reduce reflections, as shown in Figure 1.

The variables are the waveguide transition length $L1$ and the dielectric taper length $L2$. These two parameters are critical for matching the air-loaded WR90 to the dielectric-loaded WR62 for a smooth transition and good impedance matching. Ideally, longer $L1$ and $L2$ would reduce reflections due to impedance mismatch, which improves bandwidth and reduces insertion loss due to increased transmission efficiency. However, a balance must be struck between practicality and RF performance; lengths cannot be unlimited, and the next section will optimize the variables.

The characteristics of a dielectric taper depend on the dielectric constant and loss tangent of the 3D-printed material. The dielectric constant determines the size-reduction ratio, whereas the loss tangent results in insertion losses as the RF wave propagates through the dielectric-loaded waveguide. Equation (1) governs the geometrical dimensions of the air-loaded or

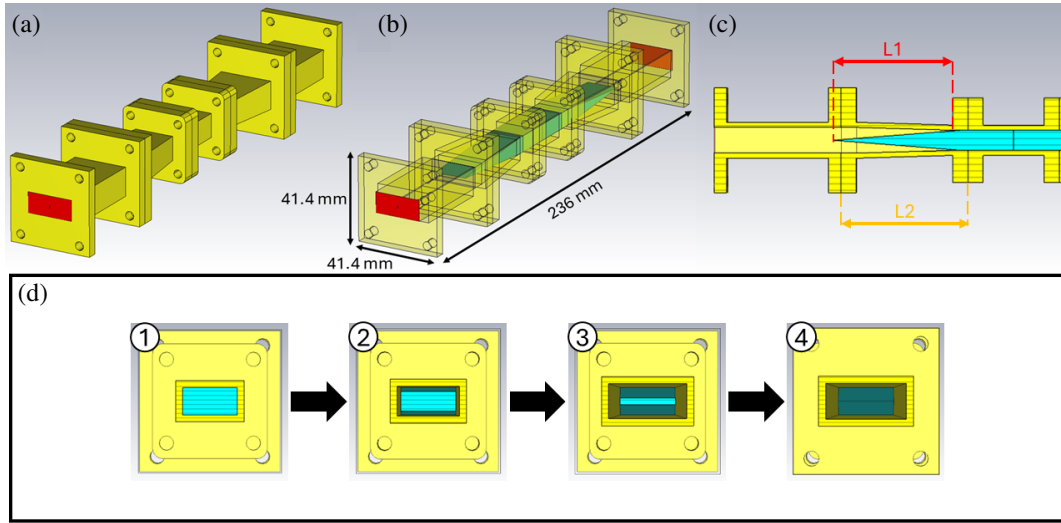


FIGURE 1. (a) Overall design of proposed waveguide component, (b) inner workings of the dielectric taper, (c) critical parameters for impedance matching, and (d) cross-sectional area for smooth tapered transition.

dielectric-loaded waveguide section by replacing the dielectric constant ε_r with 1 for air or a constant for dielectric loading [33]. Equation (3) governs the waveguide impedance based on the dielectric constant of the waveguide medium [34]. f_c is the cutoff frequency; f is the operating frequency; c is the speed of light; a is the dimension of the waveguide width; μ_0 is the permeability of free space; and ε_0 is the permittivity of free space.

$$f_{c,mn} = \frac{c}{2\pi\sqrt{\varepsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}. \quad (1)$$

$$f_{c,TE10} = \frac{c}{2\pi\sqrt{\varepsilon_r}} \sqrt{\left(\frac{\pi}{a}\right)^2}. \quad (2)$$

$$Z_{TE10} = \frac{Z_0}{\sqrt{\varepsilon_r} \sqrt{1 - \left(\frac{f_c}{f}\right)^2}}. \quad (3)$$

$$Z_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}}. \quad (4)$$

The above equations govern the WR90 waveguide, which operates in the X-band. The out-of-band part is obtained using the WR62 waveguide, which originally meant for the Ku-band, to transmit the same X-band signal. This is done by loading the WR62 waveguide with a dielectric that has an additional factor of $\sqrt{\varepsilon_r}$ in the governing equations. As shown in Figure 1(d), the transition from fully dielectric-loaded on the left to fully air-loaded on the right must be done smoothly to prevent reflections due to impedance mismatch. The cross-sectional area is governed by Equation (3), where the waveguide impedance at each cross section is determined by the ratio of dielectric loading versus the air loading at any cross-section. To transition dielectric loading to air loading from left to right, the amount of dielectric must be reduced, which decreases the effective dielectric constant term, and it disrupts the wave impedance. Thus, the

method to maintain impedance would be to increase the waveguide's height and width to reduce the cutoff frequency, essentially keeping the ratio of dielectric loading and air loading, and thus keeping the impedance matched throughout the taper transition. This forms the working principle of the dielectric-loaded taper transition for minimal reflections by impedance matching.

3. DESIGN AND SIMULATION

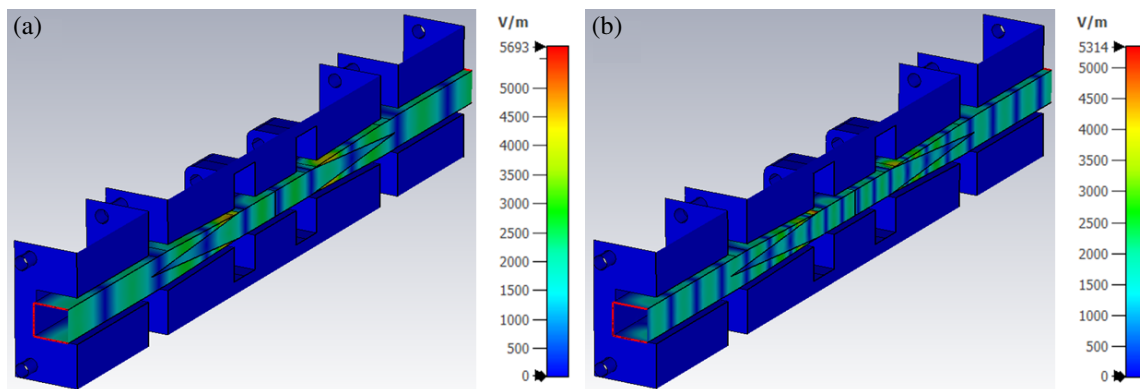
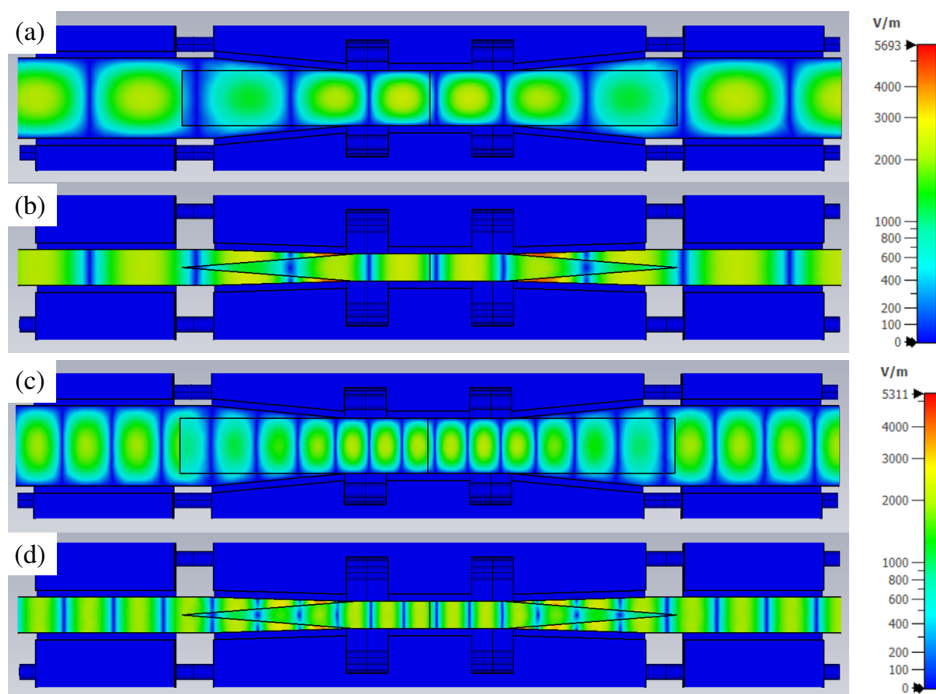
A parametric model using the governing equations was developed in Computer Simulation Technology Microwave Studio (CST MWS). The tunable parameters mentioned in the previous section were optimized to achieve efficient transmission with practicality in fabrication and its application in waveguide routing. The optimization algorithm used was the Trust Region Framework, with the goal function having the lowest insertion loss. Two WR90 waveguide ports were defined with 10 modes and excited with the fundamental TE₁₀ mode. Table 2 shows the cutoff frequency and wave impedance for the fundamental mode and possible higher-order modes. The model was run with a mesh of 27,112 tetrahedrons using the frequency-domain solver, and the materials defined are aluminum lossy metal with 3.56e7 S/m conductivity and the COC definition in the next paragraph. Figures 2, 3, and 4 show simulated electric field and S -parameter results for the optimized design.

For accuracy, the electric-field contour plot and S -parameter plot have selected only the fundamental TE₁₀ mode. The electric-field contour plot in Figures 2 and 3 shows the uniform periodicity of the RF wave, indicating a travelling RF wave inside the waveguide component, as opposed to an undesired standing wave, which would be represented as a mix of colors with significantly high electric-field amplitudes. The operational bandwidth of this optimized waveguide component can cover the entire X-band, from 8.2 to 12.4 GHz, even with a significantly smaller waveguide section.

The simulated dielectric material was a low-loss tangent material, Cyclic Olefin Copolymer (COC), whose dielectric con-

TABLE 2. Cutoff frequency of each mode at the WR90 port.

Mode		WR90 Air Cutoff Frequency (GHz)	WR90 Air Wave Impedance (Ohm)	WR62 Dielectric Cutoff Frequency (GHz)	WR62 Dielectric Wave Impedance (Ohm)
1	TE ₁₀	6.55	488.51	6.33	319.36
2	TE ₂₀	13.11	478.01	12.67	350.69
3	TE ₀₁	14.75	367.34	12.67	350.62
4	TE ₁₁	16.14	312.09	14.17	266.18

**FIGURE 2.** Simulated *E*-Field inside the component at (a) 8.2 GHz and (b) 12.4 GHz with an input power of 0.5 W.**FIGURE 3.** (a) Simulated *H*-plane cross-section *E*-Field and (b) *E*-plane cross-section *E*-Field at 8.2 GHz with an input power of 0.5 W, and (c) Simulated *H*-plane cross-section *E*-Field and (d) *E*-plane cross-section *E*-Field at 12.4 GHz.

stant and loss tangent are referenced in [35–39]. This material is available as a 3D-printing filament, which is desirable because it allows the free-form fabrication of customized waveguide chains and waveguide routing. A 10 mm-thick sample

block of COC was 3D-printed with parameters and print orientation elaborated in Section 4 for material characterization. The used method was the Nicolson-Ross-Weir method, and the used equipment was the WR90 waveguide calibration kit in Section

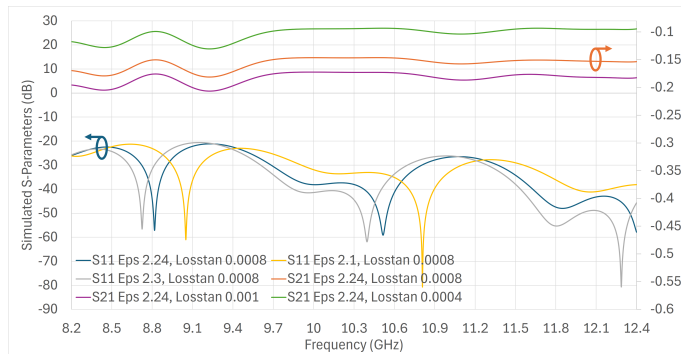


FIGURE 4. Simulated S_{11} and S_{21} performance in CST MWS including permittivity and loss tangent variations.

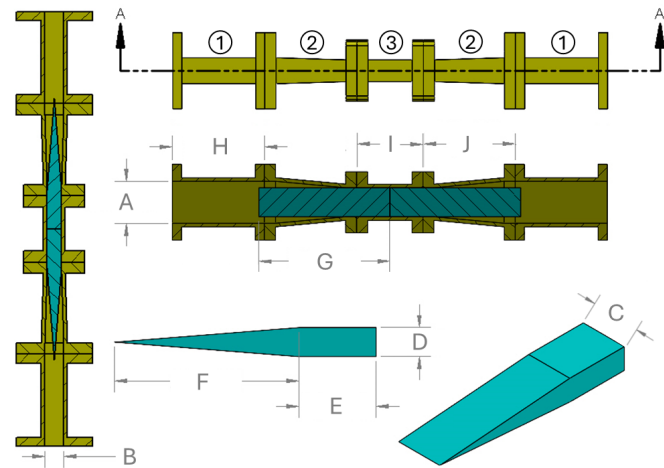


FIGURE 5. 2D Mechanical drawing for part fabrication. All dimensions are in mm. The numbers in red brackets are the measured values of the fabricated parts.

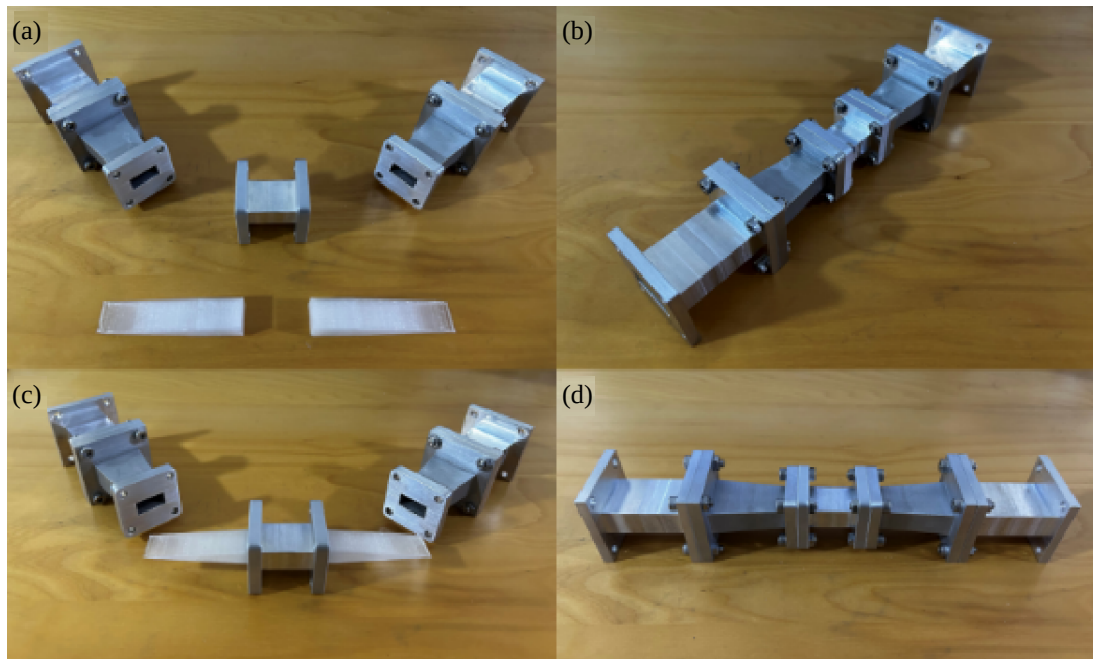


FIGURE 6. 3D-printed part fabricated using Cyclic Olefin Copolymer with machined aluminum waveguides.

5. The permittivity and loss tangent values varied across a range from 2.1 to 2.3 for the dielectric constant and 0.0004 to 0.001 for the loss tangent due to measurement uncertainty of low-loss materials. Considering the experimental values along with the literature references, a dielectric constant of 2.24 and a loss tangent of 0.0008 were used in CST MWS for the 3D-printed COC. However, a sensitivity analysis was also performed for various dielectric constants and loss tangent values in the simulated S -parameter plot in Figure 4. Figure 4 shows the result for dielectric constant of 2.24 and loss tangent of 0.0008, where S_{21} value is close to 0 dB (orange plot), indicating high transmission. Meanwhile, a good S_{11} value is achieved and better than -20 dB (dark blue plot), which indicates that most of the power is accepted into the waveguide component.

4. MECHANICAL FABRICATION

The CST model was then transitioned to mechanical modelling, as shown in Figure 5. The yellow parts were fabricated by machining aluminum (Al6061-T6), while the cyan parts were 3D printed with Creamelt® COC using the Fused Deposition Modelling (FDM) method, as shown in Figure 6. The parameters of the FDM process used were a layer height of 0.2 mm, printing temperature of 245°C, bed temperature of 90°C, 100% infill, printing speed of 30 mm/s, and retraction speed of 25 mm/s. The part was printed vertically upwards on its WR62 rectangular base. The fabricated COC part via 3D printing has a measured weight of 10.4 grams, which gives a measured density of 0.948 g/cm³ as compared to the Creamelt® data sheet of



FIGURE 7. VNA measurement setup of the waveguide component. The red line indicates the calibration plane of the WR90 calibration kit.

	Design Dimensions (mm)	Measured Dimensions (mm)
A	22.86	22.78
B	10.16	10.10
C	15.80	15.70
D	7.90	7.80
E	21.00	20.50
F	50.00	48.50
G	71.00	69.00
H	50.00	49.95
I	36.00	35.95
J	50.00	49.95

TABLE 3. Geometry dimensional parameters.

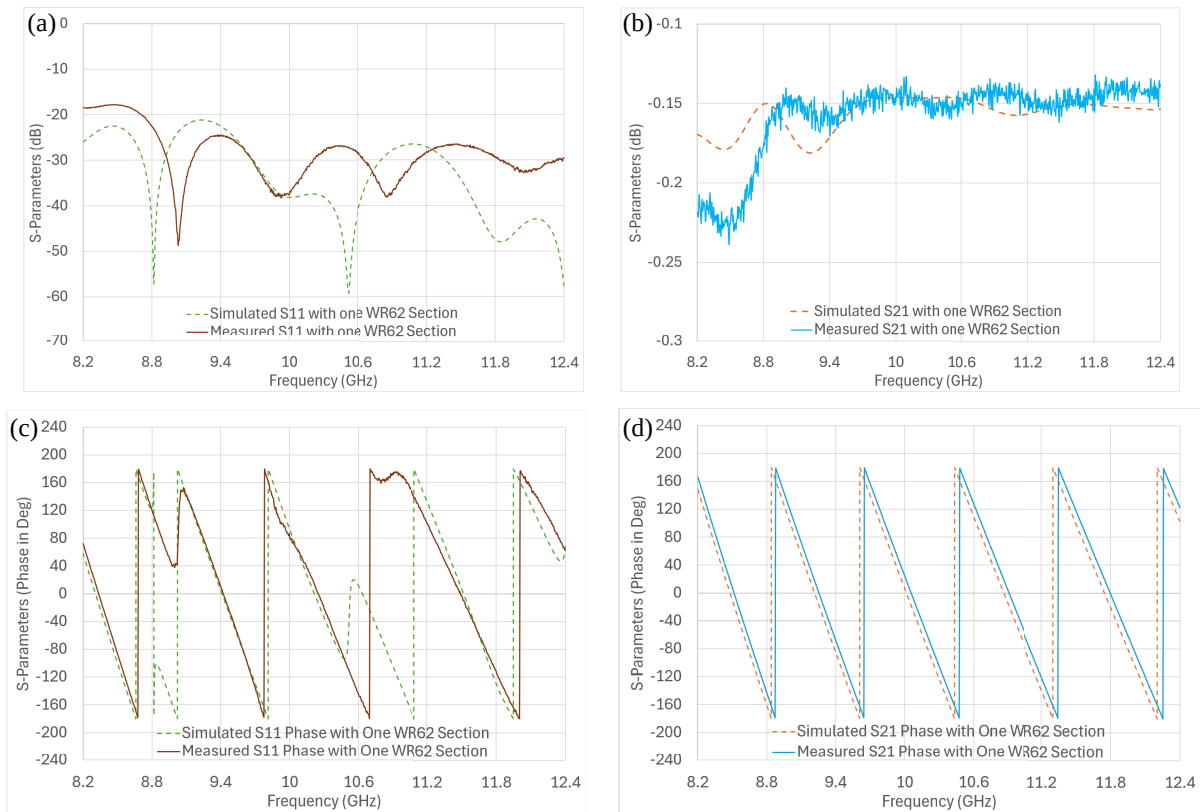


FIGURE 8. Overlaid plots of simulated and measured (a) S_{11} , (b) S_{21} , (c) S_{11} phase, and (d) S_{21} phase for one WR62 section.

1.01 g/cm³. This means that there are potential voids or porosity in the 3D-printed part due to layer-by-layer 3D printing, resulting in a decrease in the effective dielectric constant. The aluminum parts have a wall thickness of 2 mm and weigh a total of 246.6 grams, consisting of 54 grams for part 1, 51 grams for part 2, and 36.6 grams for part 3. For scalability, only part 3 is required to be repeated to extend the waveguide chain, while parts 1 and 2 do not need to be repeated. Thus, the total weight of the waveguide component in this case is 256.8 grams, and the incremental weight is due to adding more part 2 waveguide sections and the corresponding COC material loading.

Additionally, the largest tolerances come from COC parts, as 3D printing tends to give poorer tolerances than traditional aluminum machining, as seen in Table 3. As such, this could be potentially one of the main factors in deviations of the measured RF performance from simulations.

5. MEASUREMENT RESULTS

The equipment setup consisted of a Keysight Fieldfox N9918A Vector Network Analyzer and a Keysight X11644A WR90 Waveguide Mechanical Calibration Kit, as shown in Figure 7. Through-Reflect-Line (TRL) calibration was used to calibrate

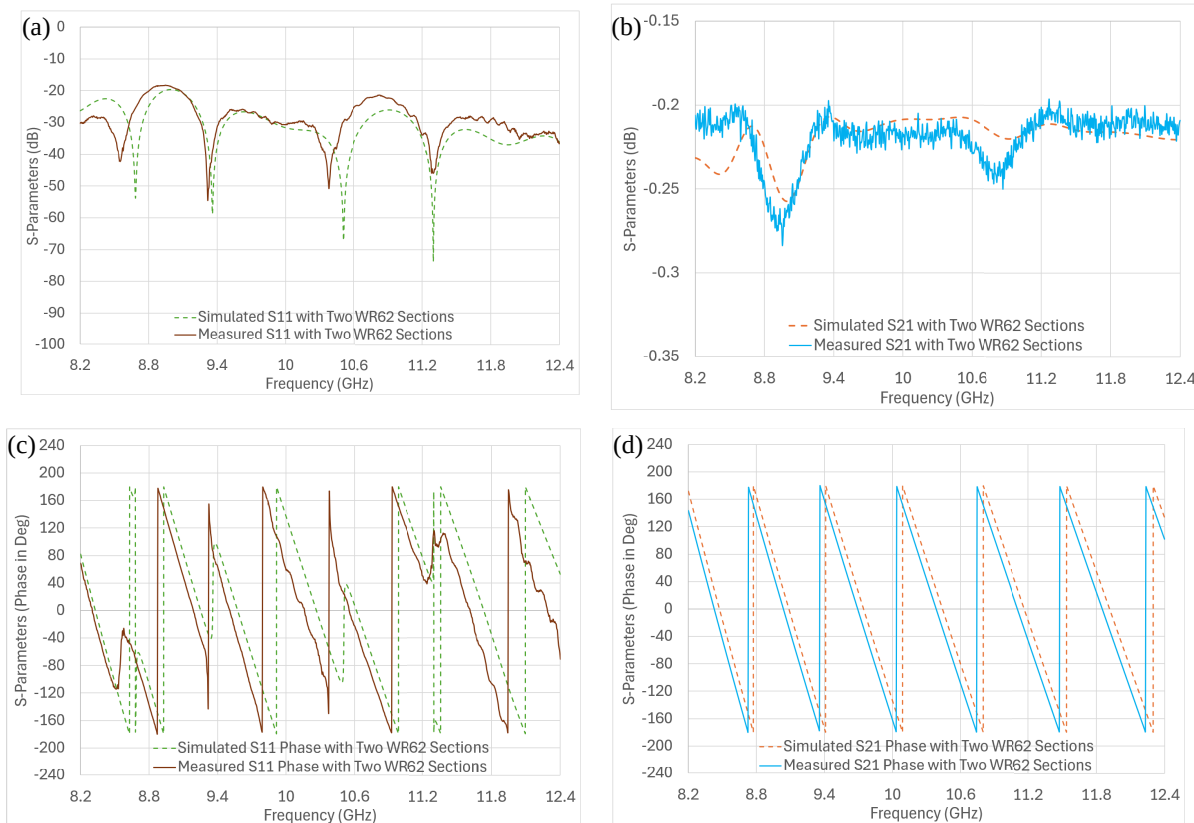


FIGURE 9. Overlaid plots of simulated and measured (a) S_{11} , (b) S_{21} , (c) S_{11} phase, and (d) S_{21} phase for two WR62 sections.

the waveguide plane to the Device Under Test (DUT), which is equivalent to the definition of the waveguide port in the CST MWS simulation.

Figure 8 shows overlaid plots of comparing S_{11} and S_{21} performance in dB and phase for one of the dielectric-loaded WR62 waveguide sections. The frequency bandwidth covered the entire X-band from 8.2 to 12.4 GHz. One possible explanation for slight deviations in S -parameters is the effect of dimensional tolerances, which result in detuning the dielectric taper transition and affect the effective dielectric constant due to smaller measured dimensions. At the center frequency of 10.3 GHz, the S_{21} was -0.15 dB while the S_{11} was -27.88 dB. The highest insertion loss of the dielectric-loaded waveguide was 0.22 dB, which was still comparable to air-loaded WR90 waveguides, where the same length of said WR90 waveguides is 0.10 dB, yet reducing the cross-sectional area by 46.35%. To compare weight fairly, a 36 mm WR62 flanged straight section weighs 41.4 grams, while an air-loaded 36 mm WR90 flanged straight section weighs 55.6 grams, which is a 26% weight reduction. To make it a fair comparison, the flange interfaces and transitions are included in the measurement for both cases, with the same calibration data, and measured several times for repeatability. This is done by keeping the same calibration plane demarcated by the red line in Figure 7 for both cases.

To determine the waveguide's RF losses in dB per meter, an additional dielectric-loaded WR62 waveguide section was added to the waveguide chain. The measurement was then repeated with the longer WR62 waveguide section, and Figure 9 shows the measured results. At the center frequency of

10.3 GHz, the S_{21} is -0.21 dB, whereas the S_{11} is -36.32 dB. The difference between the results in Figures 8 and 9 shows the addition of another repeated WR62 section of 36 mm, which corresponds to an insertion loss of 0.0591 dB at 10.3 GHz. The results suggest that even with longer waveguide sections, the insertion loss remains relatively low. The equivalent loss per meter of these dielectric-loaded WR62 waveguide sections was 1.64 dB/m, with a cross-sectional area reduction of 46.35% due to shrinking the cross-sectional waveguide area. With this, the scalability of these dielectric-loaded waveguide routings was demonstrated, which led to reducing size, volume, and weight while maintaining low RF losses. 3D printing also allows less constrained waveguide routing, such as including bends or twists, instead of just straight waveguides.

6. CONCLUSION

This study explores a dielectric-loaded out-of-band waveguide design to reduce the size, volume, and weight of long waveguide routings. A much smaller WR62 dielectric-loaded waveguide was used to propagate an RF signal that would have required a larger air-loaded waveguide to operate and yet maintain relatively low RF losses of 0.22 dB, which is still comparable to the RF losses of air-loaded WR90 waveguides (0.10 dB), while reducing the cross-sectional area by 46.35% and weight by 26%. The use of 3D printing enables the fabrication of customized waveguide chains and allows the use of the dielectric material Cyclic Olefin Copolymer in its 3D-printing-ready form. The low-loss tangent of COC allows low-loss RF trans-

mission, along with size and weight reduction of the waveguide chain, which is a key highlight of this study.

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