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DOCTORAL DISSERTATION

Title of doctoral dissertation: Novel high frequency components with non-conventional shape employing smooth geometry deformation and manufacturing by 3D printing

Title of doctoral dissertation (in Polish): Nowe układy pasywne wysokiej częstotliwości o niestandardowym kształcie wykorzystujące techniki deformacji geometrii brył oraz druk 3D

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Novel high frequency components with non-conventional shape employing smooth geometry deformation and manufacturing by 3D printing

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ABSTRACT

Additive manufacturing has emerged as a promising method for fabricating complex 3D geometries with customizable surface curvatures. These capabilities surpass the limitations of conventional techniques, allowing the realization of structures often unattainable by traditional manufacturing processes, thus enabling innovation in many areas of engineering. This doctoral thesis introduces shape deformation as a new approach to the computer-aided design of high-frequency electronic circuits. An overview of the state-of-the-art in 3D-printed passive microwave components is first provided, and, in this context, the deficiencies of existing EM design tools are exposed. On this basis, it is shown that free-form geometry deformation opens up new design possibilities in the 3D modeling of microwave structures, which helps bring the full potential of additive manufacturing to the field of microwave engineering. The proposed design framework combines shape deformation based on radial basis function interpolation with an electromagnetic field solver based on the 3D finite-element method. The described procedure is used to develop novel microwave structures with non-standard geometry exhibiting enhanced electrical performance such as high quality factor or wider spurious mode separation. In order to confirm the correctness of the proposed approach, all presented designs are experimentally verified by 3D-printed prototypes.

STRESZCZENIE

W ostatnich latach produkcja addytywna stała się obiecującą metodą wytwarzania złożonych geometrii 3D o konfigurowalnych krzywiznach powierzchni. Zastosowanie tej metody pozwala na realizację struktur, które nie są możliwe do wykonania przy użyciu tradycyjnych procesów produkcyjnych, co umożliwia wprowadzanie innowacji w wielu obszarach inżynierii. W niniejszej rozprawie przedstawiono deformację kształtu jako nowe podejście do komputerowego projektowania układów elektronicznych wysokiej częstotliwości. Zaprezentowano przegląd najnowszych rozwiązań w zakresie pasywnych komponentów mikrofalowych drukowanych w technologii 3D, zwracając uwagę na ograniczenia istniejących narzędzi projektowych. W tym kontekście wskazano, że swobodna deformacja brył otwiera nowe możliwości projektowe w modelowaniu struktur mikrofalowych, co pozwala w pełni wykorzystać potencjał druku 3D w dziedzinie inżynierii mikrofalowej. Proponowana procedura projektowa łączy deformację kształtu opartą na interpolacji za pomocą radialnych funkcji bazowych z analizą pola elektromagnetycznego opartą na metodzie elementów skończonych. Opisana metoda została wykorzystana do opracowania nowych struktur mikrofalowych o niestandardowej geometrii, charakteryzujących się lepszymi parametrami elektrycznymi, takimi jak wysoka dobroć lub większa separacja rodzajów pasywnych. W celu potwierdzenia poprawności proponowanego podejścia wszystkie przedstawione projekty zostały zweryfikowane eksperymentalnie za pomocą prototypów wykonanych w technologii druku 3D.

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LIST OF SYMBOLS AND ABBREVIATIONS

Physical quantities and mathematical expressions

$\varphi(\cdot)$	– radial basis function
f_0 (GHz)	– center frequency
Q	– quality factor, Q-factor
S_{11} (dB)	– scattering parameter, describing input power reflection from Port 1
S_{21} (dB)	– scattering parameter, describing power transmission from Port 1 to Port 2
FBW	– fractional bandwidth, defined as the bandwidth relative to its center frequency
IL (dB)	– insertion loss
Ra (μm)	– average surface roughness
RL (dB)	– return loss

Other terms and abbreviations

ABS	– acrylonitrile butadiene styrene
AM	– additive manufacturing
BPF	– bandpass filter
BW	– bandwidth
CAD	– computer-aided design
CIJ	– continuous inkjet
CNC	– computer numerical control
CSG	– constructive solid geometry
CWDM	– circular waveguide dual-mode
DLP	– direct light processing
DM	– dual-mode
DMLS	– direct metal laser sintering
DoD	– drop-on-demand
EBM	– electron beam melting

EM	– electromagnetic
ESA	– European Space Agency
FDM	– fused deposition modeling
FEM	– finite-element method
FFD	– free-form deformation
FFF	– fused filament fabrication
IEEE	– Institute of Electrical and Electronics Engineers
ILP	– integral layer peeling
LPF	– low-pass filter
LS	– level-set
MLS	– micro laser sintering
NURBS	– nonuniform rational B-splines
OMT	– orthomode transducer
OMUX	– output multiplexer
PCB	– printed circuit board
PLA	– polylactic acid
RBF	– radial basis function
RF	– radio frequency
SATCOM	– satellite communication
SLA	– stereolithography apparatus
SLM	– selective laser melting
SLS	– selective laser sintering
TE	– transverse electric
TG	– topology gradient
TM	– transverse magnetic
TZ	– transmission zero
UV	– ultraviolet
VNA	– vector network analyzer

Chapter 1

INTRODUCTION

1.1. *Motivation*

High-frequency passive components, such as filters, multiplexers, power dividers, feeding networks or antennas, are an important part of every modern telecommunication system. The role of these components is to transmit modulated electromagnetic (EM) wave signals from the generator into free space, receive the incoming signals at the receiver, and ensure proper separation of neighboring channels. When designing electronic circuits for high frequencies such as microwaves, the difference compared to low-frequency electronics is that the electrical response of the device depends greatly on its physical realization (i.e., its shape and size). The relation between circuit geometry and its transfer function depends on the EM field distribution across the component. This relation is particularly important when considering metal waveguide components, such as those used in high-power satellite and radar systems. Therefore, in order to properly design an electronic circuit at microwave frequencies taking into account the field effects, advanced software tools for EM field simulation are commonly used. The design procedure involves preparing a computer model of the device, executing a series of computationally intensive EM simulations, and optimizing the geometry to achieve the requested electrical specification.

The ongoing expansion of modern wireless communication systems is driving the need for higher data throughput, lower latency, and greater reliability, while also requiring networks to support extremely dense channels with improved spectral and energy efficiency. As system requirements continue to grow, so does the demand for new high-frequency electronic circuits that provide enhanced electrical performance, greater miniaturization, and better integration. Meeting these requirements inevitably increases design complexity, making the development of advanced microwave components an increasingly challenging task. Since conventional fabrication methods may no longer be sufficient in fulfilling these objectives, advanced manufacturing technologies are being explored as a way to overcome current limitations.

The rapid development of additive manufacturing (AM) and its introduction to the RF and microwave industry has brought a groundbreaking change in paradigm related to the design and fabrication of high-frequency components [1–4]. By employing 3D-printing techniques, the creation of 3D structures of almost arbitrary shapes and surface curvatures becomes possible, opening a range of new possibilities in the attainment of complex radio frequency (RF) components. The modeling flexibility offered by AM is a great advantage over conventional fabrication technolo-

gies, as it lifts the limitations of established processes such as computer numerical control (CNC) milling. AM enables the production of components with geometries and functional integration that are challenging to achieve with conventional manufacturing. As a result, AM is increasingly viewed as an enabling technology for many microwave engineering applications, including mobile and satellite communications. In this context, 3D-printing techniques can play a pivotal role in developing a new generation of microwave devices and in opening the way towards design solutions that have not yet been explored. In particular, as recognized in [5], additive manufacturing of novel RF components has become one of the leading research and development directions in the space communications sector, and this trend is expected to gain importance in the coming years.

Although AM has unlocked the potential for monolithic fabrication of complex 3D structures, the limitations of conventional modeling tools still remain a significant constraint in the design of such components. This is because, until recently, the prevailing paradigm for 3D modeling of RF components in professional design tools relied on constructing models from a library of predefined solid geometries. Such an approach was adequate given the capabilities of conventional fabrication methods, such as CNC machining. However, this is no longer the case with the emergence of advanced manufacturing technologies such as AM. Therefore, the development of new EM computer-aided design (CAD) tools and design methods is essential in order to allow for the design of microwave components with nearly arbitrary shapes and help shift the design paradigm towards a truly AM-oriented one. To this end, a new design optimization framework based on the free-form shape deformation technique is proposed. The new method is applied to design novel microwave components benefiting from non-conventional, optimized shape and manufacturing by 3D printing, which show improved RF performance over state-of-the-art solutions.

1.2. Goals and claims of the work

The primary objective of this work is to overcome the existing limitations of EM modeling by enhancing conventional EM CAD tools by incorporating the free-form shape deformation of 3D solids and to demonstrate the opportunities offered by this technology. Within the scope of this work, a new method for design and geometric optimization of passive microwave structures is proposed and closely studied. The integration of the aforementioned modeling technique with a full-wave EM field simulation solver is shown to enable the potential brought by 3D printing technology. The proposed design optimization toolkit is then used to develop new AM-oriented non-planar microwave components with unique shape and smooth surface profile. New designs for microwave resonators and filters are presented, along with the detailed description of the design procedure, with a special emphasis on shape optimization. Each design presented in this dissertation is experimentally validated using a 3D-printed prototype. Selected topics regarding 3D printing processes for RF electronic components are also discussed. It is shown that the manufactured prototypes exhibit improved RF performance over conventional designs.

In essence, the claims of the thesis are:

1. A free-form shape deformation technique based on radial basis function interpolation opens up new design possibilities and offers additional degrees of freedom in the 3D modeling of microwave electronic components,
2. Using shape deformation in EM design allows for exploiting the potential of additive manufacturing processes in the fabrication of microwave devices with complex shapes and smooth profiles,
3. Implementing free-form deformation in the design process, combined with fabrication by additive manufacturing, can lead to high-frequency components with improved RF performance,
4. By combining advanced 3D-printing processes with appropriate post-processing, it is possible to produce microwave hardware whose performance is comparable to that of components manufactured using established processes.

1.3. Thesis outline

This dissertation is based on the research findings described in six original publications [MB1–MB6], included herein in their entirety. The thesis is divided into five chapters. Chapter 1 presents the motivation and defines the goals of this work. Chapter 2 provides basic considerations for microwave electronics design, reviews the latest developments in passive 3D-printed microwave solutions, and in this context identifies the shortcomings of current EM design tools. The main contribution of this work is the introduction of the shape deformation method to EM design along with its integration into a design optimization framework, as described in detail in Chapter 3. The remaining part of this dissertation is devoted to the application of this method to design novel microwave resonators and filters with superior performance, which has been reported in five international journal publications, and presented in Chapter 4. Part of the research presented in publications [MB1–MB3] was conducted within the framework of a project coordinated by the European Space Agency (ESA), in collaboration with external partners; ESA's involvement highlights the strong potential of this work for practical application in satellite communication systems. Finally, the conclusions and outlook for future research is given in Chapter 5.

Chapter 2

OVERVIEW OF THE STATE-OF-THE-ART

This chapter discusses the current paradigm for EM design and 3D modeling of high-frequency electronic components. The capabilities of currently available design tools are carefully examined in the context of new and emerging fabrication technologies such as AM. A detailed overview of the latest developments in passive 3D-printed microwave components is presented. Based on the solutions analyzed, the main limitations of the design process and opportunities for improvement are identified.

2.1. Basic principles regarding high-frequency electronic design

The design of microwave components differs significantly from the design of low frequency electronic devices. In low frequency electronics, the design process is based on the analysis of a circuit model governed by Kirchhoff's laws. The designer uses lumped elements, such as resistors, capacitors or inductors as well as semiconductors such as diodes or transistors. An example of a lumped-element circuit is shown in Fig. 2.1a. However, at higher frequencies such as microwaves, the wavelength of the guided signal becomes comparable to the physical dimensions of the structure, which means that the electrical response of the circuit depends on the spatial distribution of the EM field inside the component. This implies that at RF, lumped components have to be replaced often entirely with distributed elements, such as short or open sections of transmission line, resonating structures, etc. Consequently, RF electronic circuits can be designed using distributed elements only, i.e., by determining and tuning the physical geometry of the structure, as shown in Fig. 2.1b. Therefore, in order to properly design an electronic circuit at microwave frequencies taking into account the field effects, advanced numerical methods for EM field simulation need to be used. In practice, the design procedure involves building a realistic device model in a full-wave simulation software, and performing a series of computationally intensive EM simulations. Obtaining an electrical response that will fulfill the requested specification often requires a costly optimization of the design.

Since the geometry of an RF component directly affects its electrical performance, full control over the shape and size of the circuit is essential for a precise design process. However, one of the weakest points of that process is the limited ability of EM-oriented CAD environments to represent and manipulate geometrically complex structures. Although the majority of currently used EM software tools are generally very robust and propose many advanced utilities, the main

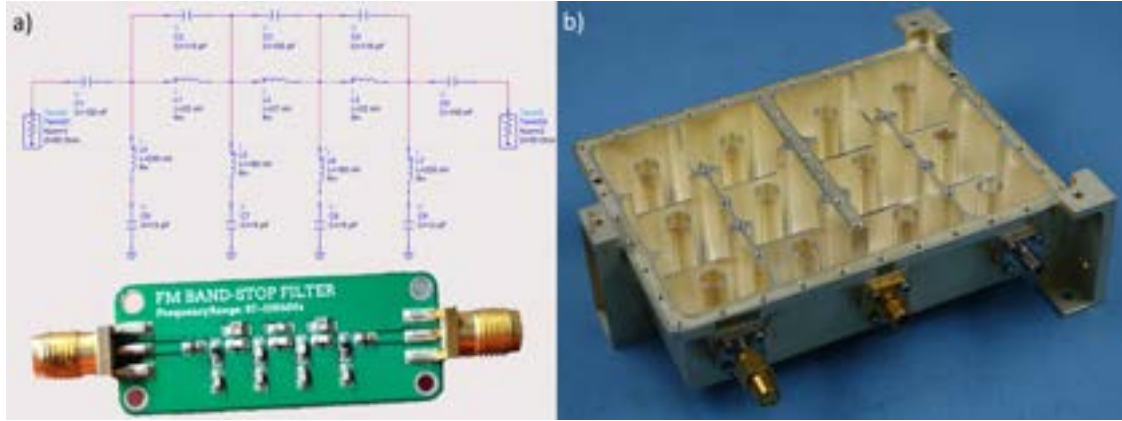


Fig. 2.1: Examples of lumped and distributed electronic circuit realizations: a) A band-stop filter designed using lumped elements (inductors and capacitors): equivalent circuit and physical realization on a printed circuit board (PCB) [6]; b) A coaxial cavity diplexer designed with distributed elements – combine resonators [7].

limitation of all these tools lies in the lack of full 3D modeling freedom, i.e., due to the fact that they rely on the constructive solid geometry approach.

2.2. 3D modeling with Constructive Solid Geometry

Constructive solid geometry (CSG) is the prevailing paradigm for 3D modeling of EM structures, widely used among all major CAD tools for microwave components design. CSG is based on using simple, predefined geometrical primitives (such as boxes, cylinders, spheres) and applying Boolean operations to them, such as union, subtraction, intersection, or other relatively simple shape modifications, e.g. edge filleting or chamfering, as shown in Fig. 2.2. In addition, some software tools allow the creation of arbitrarily shaped surfaces based on parametric, equation-based objects (e.g., B-Splines [9], or Non-Uniform Rational B-Splines (NURBS) [10]), which are often considered the industry standard for modeling curves and surfaces in CAD and computer graphics. While parametric surface description is more general than CSG, it may be cumbersome, especially when optimizing complex structures. This is because such objects (curves or surfaces) are defined by a large number of control parameters, which ultimately amounts to a large number

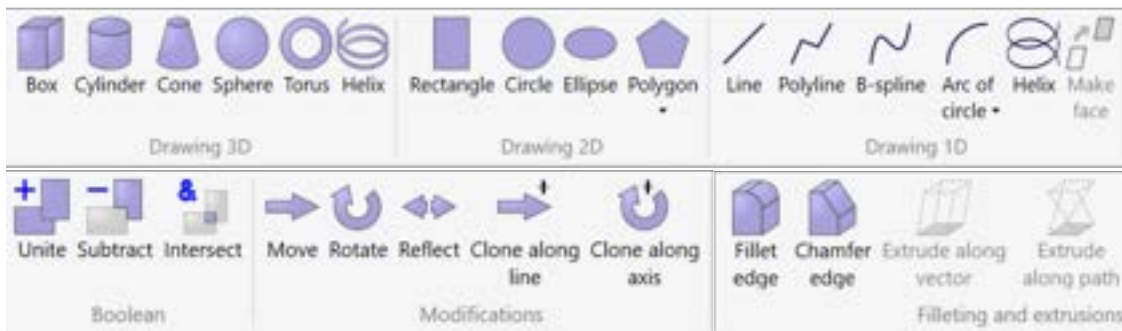


Fig. 2.2: Some of the CSG operations available in InventSim 3D FEM electromagnetic field solver, ver. 5.1 [8].

of variables, making the optimization process all the more challenging. Therefore, the design of more complex geometrical models which are not predefined or built up from building blocks is at least very demanding, and beyond a certain level of complexity, simply not feasible. Alternatively, certain practitioners developing microwave components oriented for AM, in the absence of specialized EM design tools, need to rely on external software for shaping the structure (e.g., Autodesk Fusion 360 [11]), and import/export the 3D models to analyze them in a separate EM simulation environment. Naturally, such an approach is extremely inefficient and requires an iterative design procedure involving multiple import and export operations, and as such, does not allow for integrated EM tuning or design optimization. Considering the EM CAD solutions, either based on CSG and parametric curves (such as CST Microwave Studio [12], or Ansys HFSS [13]), or using a library of predefined shapes (e.g., Microwave Wizard [14]), all the widely used industry tools impose severe constraints on the modeling capabilities for EM design.

For decades, the approach based on CSG was considered satisfactory because the design of microwave electronics was constrained by the capabilities of available manufacturing technologies, such as CNC milling. However, this is no longer the case due to the increasing popularity and availability of advanced manufacturing techniques, such as repeatable and accurate 3D printing. The currently available AM processes have reached a level of precision sufficient not only for rapid prototyping, but also for the production of specialized high-frequency electronic systems [4]. One of the key advantages of the 3D-printing process is its manufacturing flexibility – the limitations of conventional techniques, such as CNC milling, no longer stand in the way. In many cases, the AM process allows for the manufacturing of a device not feasible using any other manufacturing method. In this context, the conventional 3D modeling based on CSG has become insufficient when the development of new, AM-oriented electronic hardware is considered.

2.3. Additive manufacturing of passive microwave components

Additive manufacturing or 3D-printing, as one of the advanced manufacturing processes, has been identified as a key enabling technology by the European Union [15] and is considered an important part of the fourth industrial revolution (Industry 4.0) [16]. In the area of high-frequency electronics, utilizing AM for novel, highly integrated solutions has become one of the key research and development directions in modern telecommunication systems, such as the space microwave sector, and this trend is expected to gain importance in the coming years [5].

Over the past decade, AM has emerged as a promising method for the fabrication of passive microwave components such as filters, couplers, polarizers, power splitters and antennas [17–23], as well as highly integrated system parts such as RF front-ends and antenna feeding networks [24–28]. The ability to produce complex structures offered by AM processes opened a range of possibilities for new designs which would not be feasible otherwise. Components developed using AM can make use of custom, non-standard profiles and monoblock fabrication, which can

lead to improved RF performance, such as low loss, wide spurious mode separation, high power handling capability and thermal stability, to name a few. Besides, AM can also contribute to the miniaturization of systems by integrating several functional parts into one physical device [18,29]. In order to accurately assess the potential of this fabrication technology along with its limitations, selected 3D-printing processes often used for microwave electronics are presented, followed by a critical assessment of the state-of-the-art.

2.3.1. Additive manufacturing processes used in microwave electronic design

Although commonly referred to collectively as AM or simply 3D printing, there is in fact a wide variety of additive processes, differing in terms of printing method, materials used, precision and, therefore, application. Among the processes used specifically for fabricating RF and microwave components, Tomassoni *et al.* [4] distinguished four categories of methods: material extrusion, material jetting, vat photo-polymerization, and powder bed fusion. They have been summarized in Table 2.1.

AM processes based on material extrusion, such as fused deposition modeling (FDM) — also known as fused filament fabrication (FFF)— are among the most commonly used 3D printing techniques in many application areas, due to their wide accessibility and low manufacturing costs. In this process, the desired 3D object is built layer by layer by extruding a thermoplastic material through a heated nozzle. Typical materials used in extrusion processes are polymer filaments, such as acrylonitrile butadiene styrene (ABS), polylactic acid (PLA), or nylon. Because these filaments are not conductive, usually surface metallization is required after printing to obtain the desired EM operation. Although FDM offers low-cost fabrication, FDM printers typically have limited precision; therefore, their use in RF applications is mainly restricted to prototyping. Nevertheless, several practical RF components have also been attempted [30–32].

Another category of 3D printing methods used for RF applications is material jetting. This process is similar to a standard 2D inkjet printer; droplets of a build material are deposited by a moving nozzle and solidified by exposure to ultraviolet (UV) light. Unlike material extrusion printers, this process can produce multi-material parts by jetting two or more materials simultaneously. Typically, there are two movable nozzles: the first deposits the build material and the second places the support material, which is used to support overhanging features and is dissolved post-printing. Depending on the way the material is deposited, the process is usually classified as either continuous inkjet (CIJ) or drop-on-demand (DoD) printing [33]. The materials used as ink are typically wax-like photopolymers with dielectric properties; however, conductive inks such as silver nanoparticle ink can also be used, enabling direct printing of electronic circuits on dielectric substrates [34,35]. Inkjet printing and material jetting techniques have also been used to produce antennas and metasurfaces [36]. The main limitation of this method is the required viscosity of the material used so that it can be deposited in droplet form through the nozzle.

Vat photo-polymerization is an alternative AM process that creates 3D objects by selectively curing layers of liquid photopolymer resin with a UV light source such as a laser or digital projector. The liquid resin is contained in a vat, where it is progressively solidified layer by layer using a UV source. After printing, the part typically undergoes washing and additional UV curing to enhance its mechanical and thermal properties. In order to obtain conductive surfaces, the 3D-printed structure has to be metal coated (e.g. copper or silver plated) after printing. Based on how the layers are scanned, two photo-polymerization methods are distinguished. Stereolithography (SLA) is a process which implements vector scanning by a focused UV laser source. The laser traces the shape of each layer on the surface of the resin, causing the exposed areas to solidify. Conversely, in direct light processing (DLP) the entire layer is cured simultaneously using light from a projector such as a UV lamp. SLA offers higher resolution than DLP, at the expense of printing speed. Vat photo-polymerization processes, especially SLA, have been used for fabricating a wide variety of microwave components such as waveguides, filters, antennas with high precision and surface finish [20, 37, 38]. The average surface roughness (R_a) of SLA 3D-printing can reach 5-10 μm , so that SLA-printed high-frequency components operating at W-band (reaching 100 GHz) have been successfully demonstrated [39].

Another type of 3D printing technology which is often considered for RF applications are AM processes based on powder bed fusion, such as selective laser sintering (SLS), selective laser melting (SLM), and electron beam melting (EBM). The common denominator of these methods is their principle of operation: a thin layer of powder material is spread across a building platform and selectively fused by a high-energy source such as a laser or an electron beam. This process is repeated layer by layer until the desired 3D model is formed. The material used as powder can be plastic (as in SLS) as well as metal (as in SLM or EBM) or ceramic [40]. The method which is particularly useful for microwave applications is SLM, also referred to as direct metal laser sintering (DMLS), because it allows for monolithic (single-piece) fabrication of complex structures directly in metal [2, 3]. This process results in fully metallic parts, so no metal coating or electroplating of the surface is necessary, although post-treatment by surface polishing may be required to achieve a satisfactory surface finish. With SLM, one can produce microwave components using various metal alloys, such as stainless steel [41], aluminum (AlSi10Mg) [2, 42], titanium (Ti6Al4V) [2], Invar [43] or bronze (Cu-15Sn) [44]. The dimensional accuracy of SLM printers is usually in the range of 30–100 μm , and the obtained average surface roughness (R_a) is about 5–30 μm [45]. While this level of accuracy and surface quality is deemed sufficient for RF applications up to ca. 30 GHz, it becomes a serious limitation for higher frequencies such as mm-wave and sub-THz. For such high-end products, a high-precision SLM process named micro laser sintering (MLS) can be utilized. By using extremely fine metal powders, highly focused laser beams and thin layer thickness, MLS reaches the dimensional accuracy of $\pm 15 \mu\text{m}$, which allows the production of advanced microwave components operating at frequencies up to 300 GHz [46, 47].

Tab. 2.1: Comparison of additive manufacturing processes for passive microwave components

AM process type	Representative techniques	Typical materials	Typical RF application	Limitations
Material extrusion	FDM	ABS, PLA, nylon	Prototyping, dielectric supports	Surface roughness, limited resolution
Material jetting	CIJ, DoD	Polymer and conductive inks	Planar RF circuits, antennas, metasurfaces	Material viscosity, ink conductivity
Vat photo-polymerization	SLA, DLP	UV-curable polymer resins	Waveguides, filters, antennas & more	Metal coating required for conductivity
Powder bed fusion	SLS, SLM, EBM	plastic, ceramic, metal alloys (Al, Ti, other)	Waveguides, filters, RF front-ends & more	Surface roughness, polishing may be required

FDM: fused deposition modeling, ABS: acrylonitrile butadiene styrene, PLA: polylactic acid, CIJ: continuous inkjet, DoD: drop-on-demand, SLA: stereolithography apparatus, DLP: direct light processing, SLS: selective laser sintering, SLM: selective laser melting, EBM: electron beam melting, Al: aluminum, Ti: titanium.

While AM opens up a wide range of possibilities for EM design of complex 3D structures, it comes with its own limitations and manufacturing restrictions that need to be considered when developing high-frequency components for 3D-printing. These restrictions imply a specific design for manufacturing (DfM) approach, with certain rules and guidelines to follow. For instance, although the range of allowable shapes which can be printed is extremely large, the 3D object is built sequentially layer by layer along a given printing direction, and thus each part of the structure must be supported (no overhanging parts are allowed). This means that if the printed model contains surfaces inclined above a certain angle (usually 45°), these surfaces must be supported by additional posts. As long as these supports are easily accessible, they can be removed in a post-printing step. However, for certain RF components, such as metal waveguide structures or integrated RF front-ends, support removal might be difficult or even impossible, as the interior of the structure is inaccessible after printing. On the other hand, leaving such supports inside the component is not an option, as they would change the EM response of the device and hence degrade its operation. Another limitation critical to RF engineers, which has already been mentioned above, concerns the surface roughness and effective electrical conductivity of the 3D-printed parts. This is because, although they are developing rapidly, AM processes still have not reached the accuracy and repeatability of conventional manufacturing techniques such as CNC milling or micromachining. Therefore, in most cases, post-processing treatment such as surface polishing, metal plating is necessary to reduce the losses resulting from poor conductivity. Although there are very precise AM processes such as MLS, which can achieve higher accuracy, they are still very rare and costly. Considering all the above, new ways of applying AM technology to the RF sector and improving the fabricated hardware are still being explored by academia and industry alike.

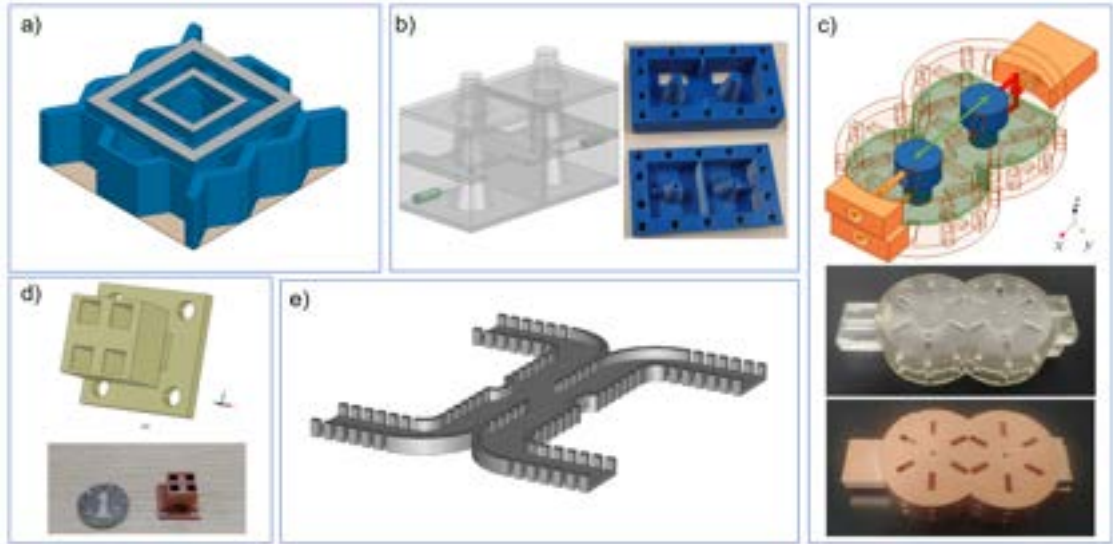


Fig. 2.3: Some realizations of 3D-printed microwave components based on CSG: a) honeycomb metamaterial absorber fabricated with FDM [17], b) microwave filter with conical posts printed using material jetting [53], c) coaxial diplexer printed using SLA [22], d) filtering aperture antenna [18], e) E-band hybrid coupler fabricated with SLA [20].

2.3.2. Review of the state-of-the-art 3D-printed RF components

Over the course of the past few years, the question of how to fully exploit the geometrical freedom offered by AM in the design process has become a pressing subject among researchers and engineers in the field of microwave engineering. Referring to the state-of-the-art, numerous publications on various 3D-printed passive microwave components have been reported in recent years, including transmission lines [19, 38, 48], filters [42, 49–53], couplers [3, 20], polarizers [19, 23], antennas [3, 18, 38, 54], as well as microwave absorbers [17, 55] and metasurfaces [36]. While it is clear that the earlier studies (2014-2015) focused primarily on examining the viability of the new manufacturing technology [30, 48, 56], it was shortly followed by a search for new solutions enabled by AM [3–5, 29, 51, 57]. A few examples of such advanced 3D-printed passive components developed in recent years are given in Fig. 2.3. Although AM brings out the potential for creating complex and intricate structures, the presented literature survey reveals that the limitations of conventional modeling tools still remain a significant constraint in the design of such components. When examining the 3D-printed structures presented in Fig. 2.3, it becomes apparent that all these designs are constructed from relatively simple building blocks (e.g., cones in Fig. 2.3b, cylinders in Fig. 2.3c or cubes in Fig. 2.3d), effectively using only a fraction of the capabilities that AM provides. For instance, in metallic waveguide technology, AM has opened a wide range of attainable shapes and smooth-profiled surfaces, and the possibility of their efficient single-piece fabrication. Even so, the most frequently reported 3D-printed waveguide devices are still limited to spherical, hemi-spherical or ellipsoidal models [37, 43, 51, 58], or other regularly shaped structures [19, 21, 50], as shown in Fig. 2.4. There are numerous related publications presenting novel components manufactured by AM, which all follow that pattern and thus show that RF design is still bounded by the limitations of EM software tools and the conventional CSG approach.

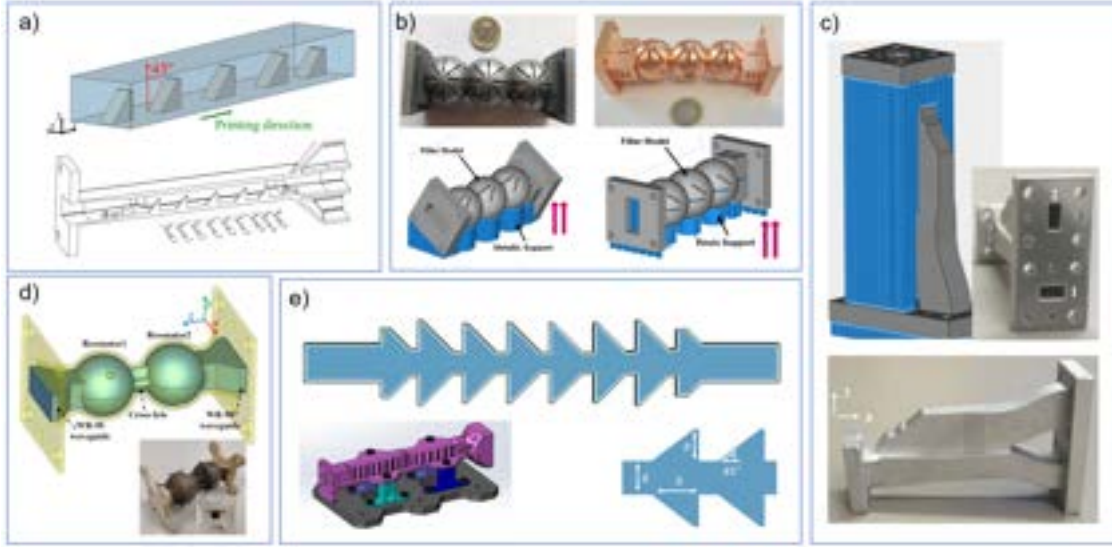


Fig. 2.4: Realizations of the state-of-the-art 3D-printed waveguide components: a) evanescent mode filter printed with SLM [50], b) slotted spherical resonator filters (fabrication by SLM and SLA) [58], c) Ka-band orthomode transducer (OMT) printed with SLM [19], d) Invar spherical dual-mode filter printed with SLM [43], e) SLM-printed corrugated low-pass filter [21].

The consideration of 3D structures of a higher level of complexity in the design of AM-oriented microwave devices has been addressed in only a few publications, which are still rather limited. Examples include proposals of twisted or bent waveguides [29, 59], or the introduction of certain shape optimization of metal or dielectric posts placed inside a resonant cavity [52, 60–62]. Clearly, the full potential of AM for RF has not yet been fully discovered and exploited.

2.4. Overview of shape optimization methods suitable for AM-oriented microwave design

In order to overcome the inherent limitations of the CSG approach, several advanced strategies for shape optimization have been proposed to date for application in RF design and integrated into EM CAD [63–66]. These methods aim to exploit the free-form geometries beyond the well-established modeling bases, while maintaining compatibility with full-wave EM simulation and optimization. Among the most prominent are: parametric geometry optimization, which uses analytically defined or spline-based shapes; gradient-based shape optimization, which enables continuous boundary deformation driven by sensitivity analysis [67]; and topology optimization, which allows the material distribution itself to evolve and thus discover new structural configurations [68].

The first approach is based on using NURBS, B-splines, and other equation-driven geometric models as an extension to the classical parametric modeling available with CSG. In this framework, the geometry of a microwave component is described using analytic or spline-based functions, such as B-Spline curves and NURBS surfaces [9, 10] and other equation-driven shapes (e.g., polynomial tapers, sinusoidal corrugations), whose defining parameters become the optimization variables. This enables highly expressive geometric representations that are particularly

advantageous for AM, since NURBS inherently enforce smooth, continuous boundaries and precise control over the modeled geometry. The approach can be easily integrated with gradient-based or global optimization methods, enabling efficient refinement of complex free-form shapes. However, it also has significant limitations: because the geometry is tied to a predefined parametric model, the optimizer cannot introduce topological changes. Furthermore, because NURBS describe surfaces rather than volumes, parameterizing a 3D object requires enforcing continuity along surface edges (a watertight shell), which may be a major practical difficulty. Moreover, constructing a well-parameterized NURBS-based model requires careful setup and often involves a large number of design variables, which can lead to ill-conditioned optimization.

Another approach relies on gradient-based shape optimization, and is most known as the level-set method [67]. In this technique, the geometry of the analyzed object is represented implicitly as a zero-level contour of a higher-dimensional scalar function, named the level-set (LS) function, typically denoted as $\phi(x)$. The shape of the analyzed structure is defined by the function so that $\phi < 0$ represents one material or region (e.g., the interior of the object) and $\phi > 0$ represents the other material or region (e.g., the exterior). To address high-frequency EM problems, the LS method can be integrated with 3D finite-element method (FEM). In such an implementation, the modeled geometry is described by a continuous scalar field $\phi(x)$ which is stored and updated on the discretized finite-element mesh. Therefore, because the FEM domain is discretized, the LS function values are sampled at the mesh nodes and interpolated within each element. During optimization, the boundary of the structure is adjusted in directions that reduce the objective function, allowing the shape to evolve without requiring an explicit geometric parameterization of the model. To determine how the boundary should move, the method evaluates the objective function by applying a perturbation of the boundary. This requires computing the corresponding shape gradient, which indicates the appropriate direction for updating the boundary and improving the design [67]. In practice, the evolving object can represent a variety of microwave materials, e.g., a ferromagnetic waveguide [69], or a dielectric resonator [63, 64], as shown in Fig. 2.5. In this example, the zero-level contour of the LS function describes the boundary of the dielectric material. The computational domain is discretized by a 3D mesh used in FEM. The cost function to be

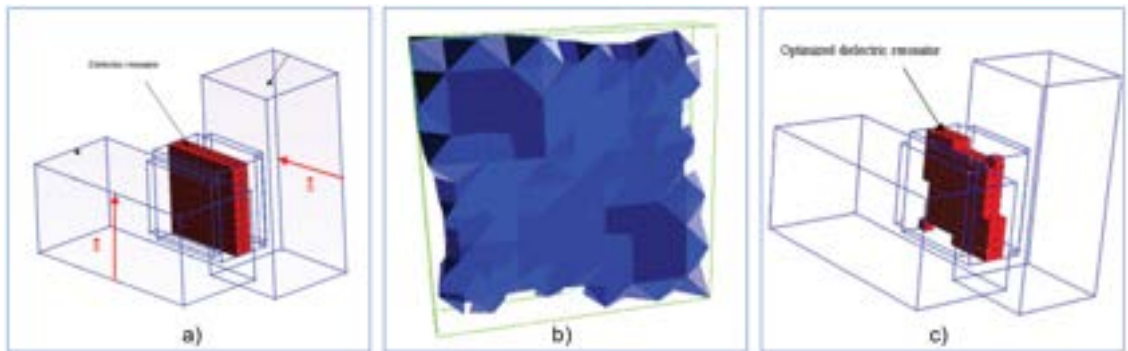


Fig. 2.5: A dielectric resonator developed using the level-set optimization method [64]: a) initial shape, b) optimal shape obtained with LS method, c) optimal shape after approximation (pixelation).

minimized is defined for the scattering parameters, S_{11} and S_{21} , obtained from EM simulation. The shape of the resonator is deformed iteratively by evolving the boundary. This example illustrates how the LS method enables versatile geometric evolution on a 3D FEM mesh and supports changes in topology. However, this technique also exhibits several limitations that must be acknowledged in practical RF design. Because the geometry is controlled implicitly by a scalar field, enforcing manufacturing constraints—such as minimum feature size, smooth surfaces, or AM-specific overhang restrictions—often requires additional filtering or regularization steps, which can complicate the procedure. The method is also sensitive to numerical discretization: maintaining a well-behaved LS function typically requires periodic re-initialization, and the accuracy of boundary motion depends strongly on mesh quality. As the optimal boundary cannot be captured with infinite precision, the final shape is approximated by the resolution of the grid, as shown in Fig. 2.5b–c. Furthermore, although the level-set framework can handle topological changes, it does not inherently guarantee smooth or physically meaningful boundaries. As a result, the obtained shape may require further processing to ensure manufacturability. The corrected or approximated geometry may then exhibit a different EM response, since it is no longer an exact one-to-one representation of the original solution found by optimization. Finally, this approach can be computationally demanding, since the entire LS field must be updated at each iteration, including the evaluation of the shape gradient on the boundary.

The last of the aforementioned techniques for shape optimization, which can be useful in AM-oriented microwave design, is the topology gradient (TG) method [68]. It is a sensitivity-based framework for topology optimization that evaluates how the objective function responds to a small change in the computational domain, that is, to selective addition or removal of material. The method evaluates the topological derivative, which measures the change in the objective function resulting from the insertion or inclusion of material at a given point in the domain. The TG method, when coupled with a 3D FEM solver, is likewise constrained by the discretized domain. The sensitivity field is evaluated on the finite-element mesh, meaning that the resolution and quality of the mesh directly influence the smallest features that can be created or removed during optimization. As with the level-set method, the achievable topological detail is therefore limited by the underlying FEM discretization. The TG method has been applied to the optimization of microwave components, e.g., for redistribution of metal on microstrip structures [70] or for shaping dielectric resonators [71]. However, the TG optimization also faces practical limitations related to discretization and practical implementation. Because the topological derivative is evaluated point-wise, its accuracy depends strongly on the underlying mesh resolution. From a computational standpoint, evaluating the derivative across the entire domain can be expensive, especially in large high-frequency EM problems. Moreover, the point-by-point evaluation of the sensitivity field often leads to scattered or highly fragmented (pixelated) material patterns unless additional conditions are applied, making post-processing or subsequent shape refinement steps essential for obtaining physically meaningful designs. In order to overcome the limitations of each approach applied separately, a combined approach based on level-set and topology gradient methods has been proposed

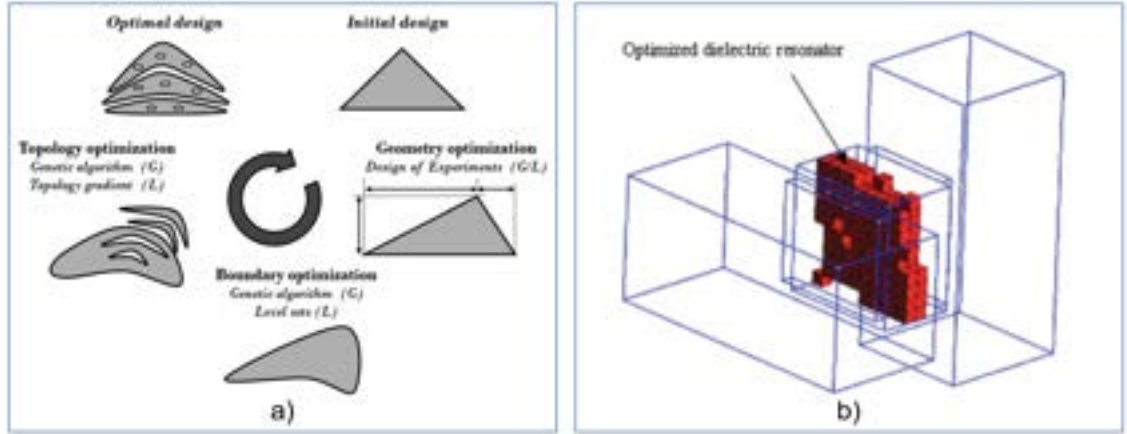


Fig. 2.6: Advanced strategies for topology and geometry optimization: a) illustration of the combined shape optimization approach in [65], b) a dielectric resonator filter optimized using the combined LS+TG approach [64].

in [64] and later enhanced with genetic algorithms [65], as illustrated in Fig. 2.6. The TG is first used to identify regions where introducing or removing material would improve the objective, enabling controlled topological changes that the LS method alone cannot generate. Once the holes in the topology are created, the shape gradient evolution refines the resulting boundaries, overcoming the tendency of the TG approach to produce irregular or fragmented shapes. In this way, the hybrid scheme supports both topological modification and boundary refinement. Nevertheless, a major limitation that remains is the restricted control over the final geometry and the continued dependence on the underlying mesh discretization. In view of the limitations associated with the described shape optimization methods, as well as the constraints imposed by parametric modeling approaches, a conclusion is reached that existing techniques still restrict geometric freedom in 3D EM design. These shortcomings highlight the need for new EM CAD tools and dedicated optimization strategies that offer greater flexibility in 3D modeling, as pursued in this thesis.

Chapter 3

FREE-FORM SHAPE DEFORMATION

The foregoing discussion revealed that the design paradigm based on CSG is still the prevailing one in microwave engineering, despite the ongoing development of new manufacturing technologies such as AM. To overcome the resulting design limitations and enable the realization of microwave components with nearly arbitrary geometries, this thesis introduces a novel approach that integrates free-form shape deformation techniques into conventional CSG-based 3D solid modeling tools. A key feature of the proposed technique is its ability to transform an initial CSG-based geometry into free-form shapes through a simple and compact deformation framework characterized by a low number of control parameters. In combination with 3D finite-element method (FEM), this approach enables efficient shape optimization during the design process without resorting to excessively complex parameterizations, such as those employed in spline-based methods (e.g., B-spline, or NURBS approaches). In comparison with methods based on shape gradients and topology optimization described in Section 2.4, the presented approach can reshape the structure in a smooth, global manner without producing pixelated or discretized geometries.

Shape deformation methods originate from the field of computer graphics, where they emerged as powerful tools for manipulating complex 3D objects. One of the first geometric modeling techniques for deforming 3D solids was the free-form deformation (FFD) technique, proposed by Sederberg and Parry in 1986 [72]. The principle of operation behind FFD involves placing the object inside a flexible, mathematically defined lattice. Adjusting the control points of that lattice modifies all enclosed geometry in a smooth and coordinated manner, enabling efficient and intuitive shape modification without the necessity of direct vertex-level manipulation, i.e., regardless of how the digital object had been initially defined using schemes such as CSG. Subsequently, other methods for geometric manipulation were developed, including an extended FFD based on adaptive lattice deformations [73], a modified FFD based on nonuniform rational B-splines (NURBS) [74], Laplacian mesh editing [75], as-rigid-as-possible surface modeling [76], and others [77]. Several deformation methods have proposed using multivariate radial basis function (RBF) interpolation as the mathematical kernel for performing the relocation of points defining the 3D model [78–81].

Outside the field of computer graphics, several of these methods have recently been applied to various areas of engineering, such as computational simulations of fluid dynamics [78, 82, 83] and low-frequency electromagnetics [84]. However, no attempts have yet been made to apply these free-form modeling techniques to the field of high-frequency electronics. Therefore, to

the best of the author's knowledge, this work constitutes the first documented contribution to the introduction of shape deformation techniques into the design of complex microwave components.

The remainder of this chapter is organized as follows: Section 3.1 describes the operating principles of the shape deformation method based on RBF interpolation, which has been extensively used throughout this work, accompanied by several illustrative examples. Next, the general framework for integrated EM design and optimization using the proposed modeling technique is then presented in Section 3.2.

3.1. Radial Basis Function interpolation

Shape deformation operations are implemented in this work using an interpolation scheme based on radial basis functions, as reported in [78–81]. Multivariate RBF interpolation is well-suited for free-form deformation because it yields smooth, continuous mappings in any number of dimensions, ensuring globally coherent results without sharp artifacts. The deformation is defined analytically over $\mathbb{R}^3$ using only control point coordinates and displacements, enabling consistent application across all types of grids: structured, unstructured, or irregular point sets.

In general, an RBF is a real-valued function φ whose value is only determined by the distance from the origin, so that $\varphi(\mathbf{x}) = \hat{\varphi}(\|\mathbf{x}\|)$. The reshaping operation is generally characterized by a mapping function $\mathcal{D}$:

$$\begin{aligned} \mathcal{D}(\mathbf{x}) : \mathbb{R}^3 &\rightarrow \mathbb{R}^3 \\ \mathbf{x} \mapsto \mathcal{D}(\mathbf{x}) &= \mathbf{x} + \mathbf{d}(\mathbf{x}) \end{aligned} \tag{3.1}$$

where $\mathbf{d}(\mathbf{x})$ is the deformation field applied to the 3D model. In the case of RBF-based deformation, it takes the following form:

$$\mathbf{d}(\mathbf{x}) = \sum_{i=1}^N \gamma_i \varphi(\|\mathbf{x} - \mathbf{x}_i\|) + p(\mathbf{x}) \tag{3.2}$$

where N is the number of control points in the defined deformation, γ_i are the weights corresponding to the radial basis functions φ defined at the positions of the control points $\mathbf{x}_i$ (in 3D $\mathbf{x}_i \in \mathbb{R}^3$, $\mathbf{x}_i = [x_i, y_i, z_i]$), and $p(\mathbf{x})$ is a low-order polynomial (defined optionally).

The choice of the basis function determines how the mesh nodes are interpolated to conform to the updated positions of the control points, it is therefore a crucial part of the procedure in order to obtain satisfying results. There are many different RBFs which may be used for multivariate function interpolation [85], such as:

- Gaussian basis function:

$$\varphi(\|\mathbf{x}\|) = \exp\left(-\left(\frac{\|\mathbf{x}\|}{r}\right)^2\right), \tag{3.3}$$

- multiquadric basis function [86]:

$$\varphi(\|\mathbf{x}\|) = \sqrt{\|\mathbf{x}\|^2 + r^2}, \quad (3.4)$$

- inverse multiquadric basis function:

$$\varphi(\|\mathbf{x}\|) = \frac{1}{\sqrt{\|\mathbf{x}\|^2 + r^2}} \quad (3.5)$$

- thin plate spline [87]:

$$\varphi(\|\mathbf{x}\|) = \left(\frac{\|\mathbf{x}\|}{r}\right)^2 \ln\left(\frac{\|\mathbf{x}\|}{r}\right) \quad (3.6)$$

- Beckert-Wendland C^2 basis function [88]:

$$\varphi(\|\mathbf{x}\|) = \left(1 - \frac{\|\mathbf{x}\|}{r}\right)^4 + \left(4\frac{\|\mathbf{x}\|}{r} + 1\right) \quad (3.7)$$

- polyharmonic spline [87]:

$$\varphi(\|\mathbf{x}\|) = \begin{cases} \left(\frac{\|\mathbf{x}\|}{r}\right)^k & \text{if } k = 1, 3, 5, \dots \\ \left(\frac{\|\mathbf{x}\|}{r}\right)^k \ln\left(\frac{\|\mathbf{x}\|}{r}\right) & \text{if } k = 2, 4, 6, \dots \end{cases}, \quad (3.8)$$

where r and k are the radius and power of the RBF, respectively, usually known as shape or scaling parameters. An illustrative representation of the selected functions is provided in Fig. 3.1. The effect of changing the shape parameter r can be observed on a Gaussian RBF in Fig. 3.2.

In order to perform the deformation, the mapping $\mathcal{D}$ needs to be specified. The procedure begins by specifying a set of N control points together with their associated displacements, which define both the direction and magnitude (length) of the offset. Subsequently, an appropriate RBF and its parameters (e.g., radius, exponent) are selected. When the basis function is chosen and the mapping function $\mathcal{D}$ defined, a linear equation system has to be solved to find the unknown weights γ_i given the coordinates of the original and displaced control points [80]. The matrix

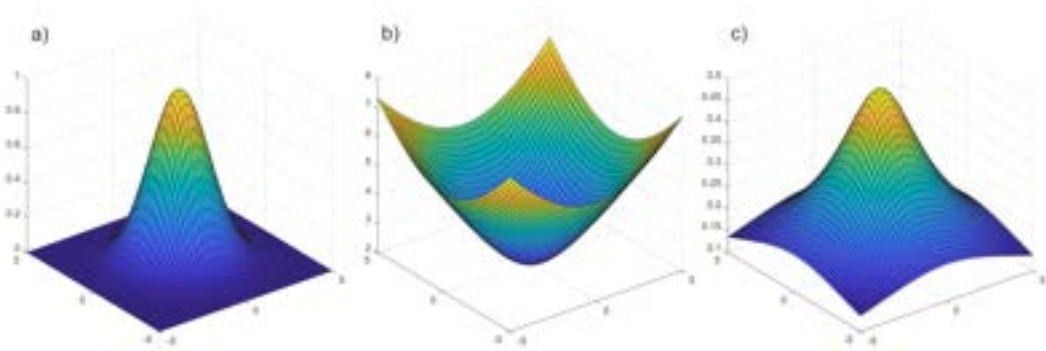


Fig. 3.1: Selected radial basis functions in 3D, with radius $r = 2$: a) a Gaussian RBF as in eq. (3.3), b) a multiquadric RBF as in eq. (3.4), c) an inverse multiquadric RBF as in eq. (3.5).

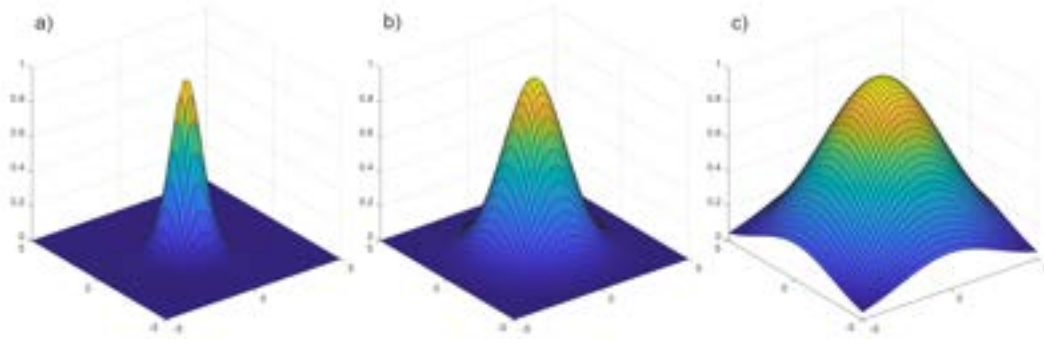


Fig. 3.2: A Gaussian RBF defined by equation (3.3), plotted in 3D for different values of the radius, r : a) $r = 1$, b) $r = 2$, c) $r = 4$.

representation of this equation system is as follows:

$$\Phi \gamma = B \quad (3.9)$$

where γ is the weight matrix of dimension $N \times 3$, corresponding to N control points defined in a 3D coordinate system. Φ is an $N \times N$ matrix containing RBF values computed for the Euclidean distances between the corresponding control points, $\Phi_{i,j} = \varphi(\|x_i - x_j\|)$, and B is an $N \times 3$ right-hand side matrix with the new (displaced) coordinates of the control points. The system is dense, symmetric, positive definite, and its conditioning depends on selected RBF and control points distribution. Once the weights have been computed, the deformation field $d(x)$ is constructed using (3.2) and can be applied to continuously displace all mesh points of a 3D object. The method is generally applicable to 3D objects described by surface or volume meshes (treated as grids of points); in the present work, a tetrahedral mesh is used to enable further application and integration with 3D FEM simulations.

The operating principle of this technique is demonstrated on the example of a cube subjected to squeezing, as illustrated in Fig. 3.3. To perform the deformation, a set of control points is defined and evenly distributed around the object, as shown in Fig. 3.3b. Note that, in general, control points can be defined around, on, or inside the object, depending on the need; what is crucial for defining the deformation is the displacement of these points, not only their initial position. In this example, the control points are defined on a box enclosing the cube. There are 26 control points in total: eight points at the vertices, six points at the center of each face, and twelve points at the midpoints of each edge. The displacement vectors are then defined for some of these points; that is, the points either remain unchanged or are moved to certain new coordinates, which causes the object to deform and results in the creation of a complex geometry shown in Fig. 3.3c. It should be noted that the deformation scheme does not limit the number of control points. More complex operations can easily be employed by adding more control points with movement in arbitrary directions. Such definition of object manipulation by moving user-defined control points is intuitive, and its results easy to predict, as the deformation follows the direction of the displacements. The operation is performed in such a way as to preserve the continuity of the object's surface (topology)

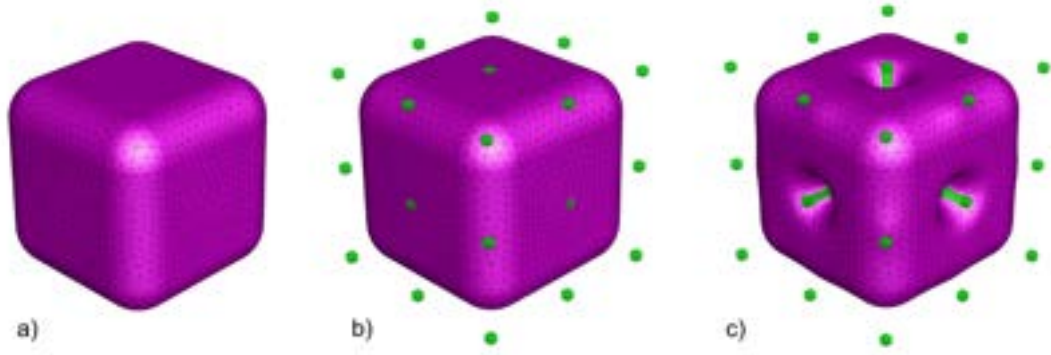


Fig. 3.3: Example of a free-form shape deformation operation: a) basic object - a cube (3D mesh view), b) a lattice of control points (green dots) defined around the object, c) deformed object (offset vectors represented by green arrows).

and its smoothness.

It can be seen that, by starting from a basic geometric shape and simply moving the control points in an arbitrary way, various complex geometries can be obtained. The influence of changing the RBF parameters on the final shape has been demonstrated in Fig. 3.4. Depending on the radius, the effects of the deformation with the chosen RBF will be more or less pronounced; the smaller the radius, the more local the deformation, as shown in Fig. 3.4(a–c). The k parameter, defined for functions such as polyharmonic spline (eq. (3.8)), determines the smoothness of the performed interpolation, as can be observed in Fig. 3.4(d–f). It follows that the selection of the most suitable RBF and its parameters will depend on the specific model and is subject to manual adjustment by the operator; no single RBF will provide satisfactory interpolation in every case.

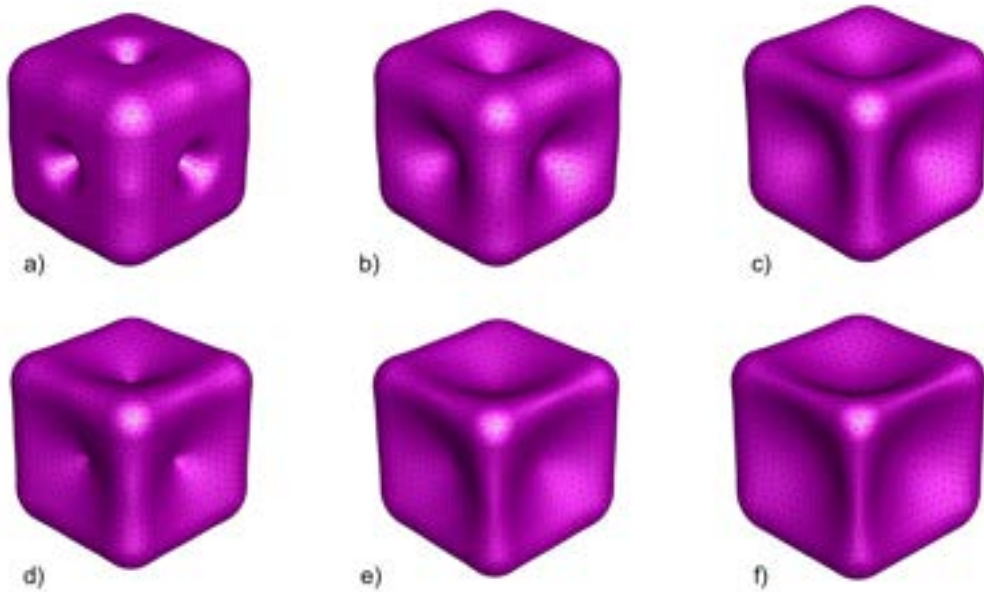


Fig. 3.4: Results of applying different RBF parameters to the cube deforming operation from Fig. 3.3: a) Gaussian (eq. (3.3)) with radius $r = 1$, b) Gaussian with radius $r = 2$, c) Gaussian with radius $r = 3$, d) polyharmonic (eq. (3.8)) with power $k = 1$, e) polyharmonic with power $k = 2$, f) polyharmonic with power $k = 4$.

Examples in Fig. 3.3 and Fig. 3.4 show that the mapping function can be easily defined with a limited number of control points, and then applied to create complex 3D objects. RBF-based shape deformation can also be applied to perform more advanced operations, such as bending or twisting complex objects, as shown in the example of a waveguide filter in Fig. 3.5. To define an operation such as rotation or bending, a set of control points in original location $P_A(x_A, y_A, z_A)$ has to be displaced to a desired new location $P_B(x_B, y_B, z_B)$. That is, the trajectory of the displacement has to be provided by assigning offset vectors $\vec{v} = (v_x, v_y, v_z)$ to the corresponding control points, such that $v_x = x_B - x_A, v_y = y_B - y_A, v_z = z_B - z_A$. What is more, one can also define a series of deformations that can be executed one after another, such as a combination of squeezing, twisting, and bending, as shown in Fig. 3.5e. In such cases, the 3D mesh of the object undergoes the first deformation, after which the resulting mesh is passed as input to the subsequent operation, and so forth. This sequential approach enables the decomposition of a complex modeling task into a series of simpler deformation steps, each of which can be more easily defined and controlled.

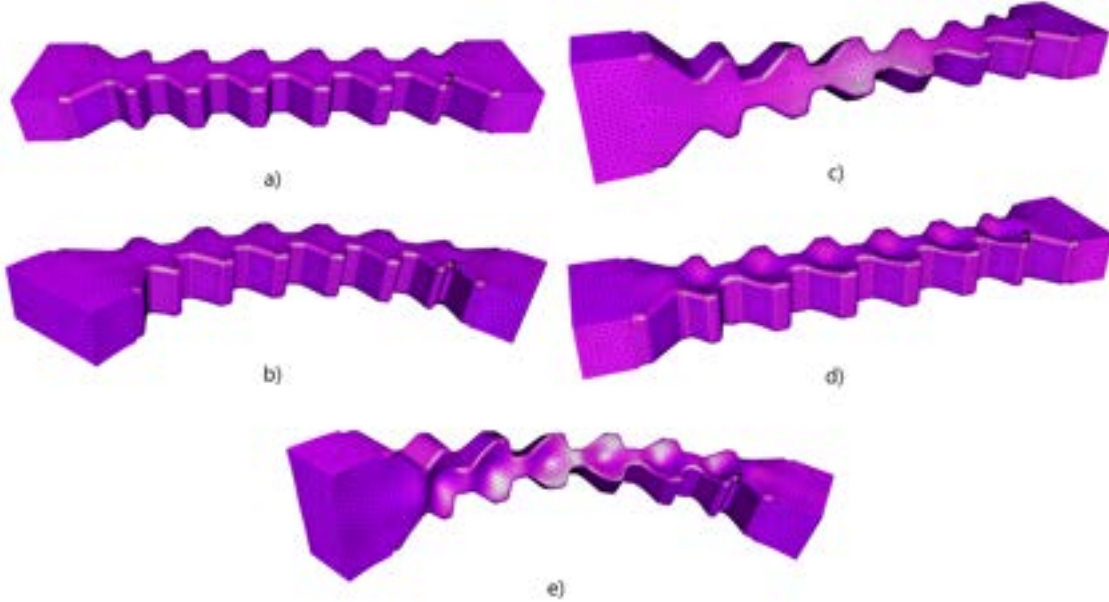


Fig. 3.5: Example of some more complex shape deformations: (a) original structure, (b) bending, (c) twisting, (d) squeezing resonant cavities vertically, and (e) squeezing + twisting + bending.

3.2. Integration with EM computer-aided design and optimization

The main contribution of this work lies in the incorporation of the proposed shape deformation technique into the design of high-frequency electronic circuits. The proposed framework employs advanced 3D modeling, enriching the standard design parametrization with a new way of object manipulation, controlled by an independent set of variables that define the shape deformation. The deforming operations utilize multivariate RBF interpolation as described in Section 3.1. The next section explains how mesh deformation is integrated with a full-wave EM field simulator, allowing for a semi-automated design of various passive microwave components. The deformation is applied directly to the volumetric (3D) mesh of the analyzed structure, which is generated prior

to executing the EM simulation using 3D FEM. What follows is a presentation of an integrated optimization framework combining RBF-based geometry modification with a fast FEM solver, as a result offering a powerful design optimization tool for microwave engineering applications. The proposed framework allows for simultaneous control over two sets of design variables, namely, the parameters of the structure modeled with CSG, and those that determine the shape deformation.

3.2.1. Integration with a full-wave EM simulator using 3D finite-element method

In order to enable simultaneous object modification and direct EM field analysis of the modeled structures, the developed shape deformation technique has been integrated with a full-wave EM field simulator based on 3D FEM [8]. The integration is based on the idea that the FEM solver operates with a volumetric tetrahedral mesh, which needs to be generated every time the analyzed geometry is changed. Therefore, it could be beneficial to combine the proposed modeling based on RBF interpolation with the inherent part of each full-wave simulation, i.e., the mesh generation step. This way, the geometric deformation can be performed directly on the 3D mesh that is first created for a basic model that was formed using CSG. The deformed mesh is then directly passed to the FEM solver, seamlessly merging the proposed method with the EM simulation workflow without introducing significant additional complexity. Such an approach sets up shape deformation as the second level of object modification, on top of the standard CSG operations available in every CAD tool. Furthermore, the computational cost of the deformation step is also not significant, as the size of the linear problem to be solved for RBF interpolation is usually small, especially compared to the cost of solving EM problems with 3D FEM. As an illustrative example, consider a single full-wave EM simulation of the sixth-order filter from [MB1]. That design is a Ku-band receive (Rx) bandpass filter, developed by shaping a straight section of a WR-75 waveguide, as shown in Fig. 3.6a. The overall simulation process can be divided into three stages: mesh generation, shape deformation, and FEM solving. The corresponding runtimes for these stages, computed on an Intel Core i9-10900X workstation with 64 GB of RAM, are listed in Table 3.1. The

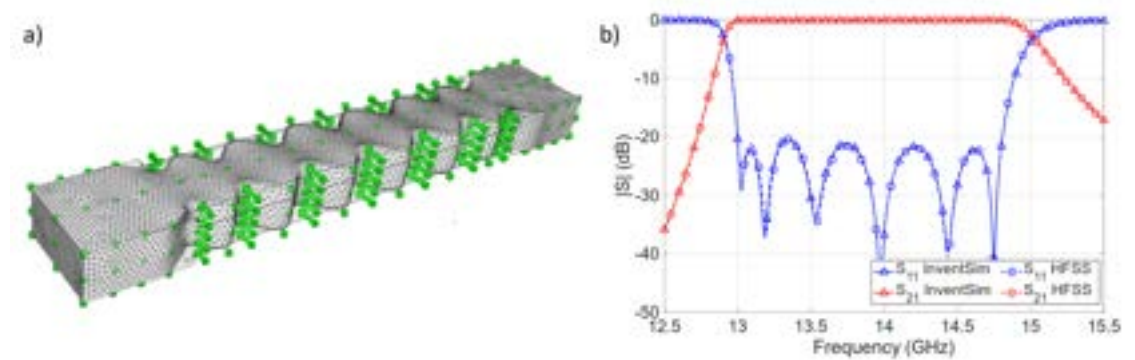


Fig. 3.6: Example of a free-form shape deformation operation performed on a sixth-order BPF from [MB1]: a) EM model of the filter (3D mesh view) with control points (green dots) and displacement vectors (green arrows), b) full-wave EM simulation result (S-parameter response), compared for direct solution using deformed mesh (InventSim) and for newly generated mesh (imported model, HFSS).

Tab. 3.1: Breakdown of simulation runtime for the sixth-order BPF [MB1] shown in Fig. 3.6.

FEM problem size	task duration (mm:ss)			
	mesh generation	shape deformation	FEM solver	total Σ
405,684	00:09	00:01	01:55	02:05
1,197,406	00:19	00:03	03:45	04:07
1,929,194	00:36	00:04	06:31	07:11

RBF used in this design is a thin plate spline with $r = 1$ as in eq. (3.6). There are $N = 196$ control points in total used to perform the deformation, which results in an $N \times N$ linear system (i.e., 196×196) to solve in the RBF interpolation problem. The runtimes are compared for different FEM problem sizes (defined as the number of unknowns in the linear system), ranging from 400,000 to nearly 2,000,000. The runtime for FEM is given for the fast frequency sweep from 12.5 to 15.5 GHz. Analyzing these results reveals that the computational cost of the shape-deformation step is negligible compared to the overall simulation time, amounting to only about 10% of the mesh-generation time and less than 1% of the total execution time. Although this varies with the specific design and its complexity, more advanced designs generally lead to FEM solving times on the order of tens of minutes to hours, while the RBF interpolation step remains consistently on the scale of seconds. This is because, for a standard RBF interpolation problem, the key cost driver is the number of control points, N , rather than the size of the mesh which is transformed. As described in the previous section, building the RBF interpolation (solving for coefficients γ) boils down to solving a dense linear system of size $N \times N$, resulting in the time complexity on the order $O(N^3)$. However, in the case of most practical designs such as the example presented in [MB1], the number of control points N is usually within the range of several tens to hundreds, whereas the size of the FEM problem reaches hundreds of thousands to millions, as shown in Table 3.1.

In addition to the discussion on the computational effort, the filter model in Fig. 3.6a illustrates how to properly define the deforming operation to affect only selected regions of the structure (here: the irises and cavity walls), while keeping the remaining parts of the model unchanged (here: the input/output waveguide sections). Since the proposed shape deformation approach is always applied to the entire mesh of the model, it is a global operation, i.e., it may alter any point in the deformed mesh. In order to ensure that some parts of the model are not affected by the operation, two actions should be undertaken. First, additional, immovable control points must be specified at the desired positions to preserve the mesh in their vicinity—as it was done in Fig. 3.6a for the short waveguide sections at the input and output. Second, the parameters of the RBF interpolation are adjusted to ensure that the declared operation has a localized effect, so that it only influences the mesh in close proximity to the displaced control points.

Because shape deformation is performed on the 3D mesh of the analyzed object, the FEM solver performs calculations of the EM field directly on the deformed mesh. While such an integrated approach significantly reduces the complexity of the whole process and brings new mod-

eling capabilities to EM design, one may raise questions about the quality of the processed mesh, and consequently, its influence on the reliability of the obtained results. It is therefore important to underline that the proposed operation preserves the continuity of the object's surface (topology) and its smoothness, and thus the accuracy of the simulation results depends primarily on the mesh density and mesh element type, as for all solutions obtained with 3D FEM. To demonstrate this, the response of the sixth-order filter example from [MB1] has been computed twice: first, in InventSim [8], using the deformed mesh; and second, in Ansys HFSS [13], where an entirely new, undeformed mesh was generated for an imported model of the designed component. The results of the simulations performed with the two software tools are compared in Fig. 3.6b, showing excellent convergence, thus confirming the reliability of the obtained solution. Furthermore, the correctness of the proposed shape deformation framework has been repeatedly confirmed with the numerous designs presented in this thesis, each of which was experimentally verified using fabricated and measured prototypes. It should also be mentioned that, in the proposed approach, the mesh used for the deformation—generated in InventSim using the third-party software Netgen [89]—is a second-order tetrahedral grid. This choice allows for a better representation of curved shapes, making it particularly well-suited for complex, arbitrarily shaped geometries.

3.2.2. General framework for EM design optimization with shape deformation

Once the RBF-based shape deformation feature has been successfully integrated with the full-wave EM simulation environment based on 3D FEM, the natural next development focused on extending this tool into a fully automated optimization framework. The main objective of the proposed toolkit, demonstrated in Fig. 3.7, is to empower the available EM design utilities with the new way of object manipulation, thus expanding the scope of attainable shapes and complex structures that can be designed. Therefore, to enable automated design of microwave components with complex shapes, it was necessary to integrate the presented geometric deformation technique with fast and efficient optimization methods, especially the ones tailored for specific microwave engineering applications, such as the zero-pole method for filter optimization [90]. The introduced framework for design optimization provides an all-in-one integrated toolkit, allowing model modification on two levels: the standard modeler based on constructive solid geometry operations, and the new shape deformation feature, available as a separately defined mesh operation. The optimization itself involves managing two sets of design variables: geometric parameters altering the basic object constructed with CSG, as well as those controlling the positions and displacements of FFD control points. Such a configuration allows simultaneous control over both parametrization levels within a single optimization.

A complete step-by-step description of the optimization process is as follows. First, a 3D mesh is generated for the defined baseline structure. This mesh is then directly modified by the defined shape deformation operation. The deformed object (represented with the processed mesh) is then processed by the EM field simulator, as shown in the block diagram in Fig. 3.7. Depending

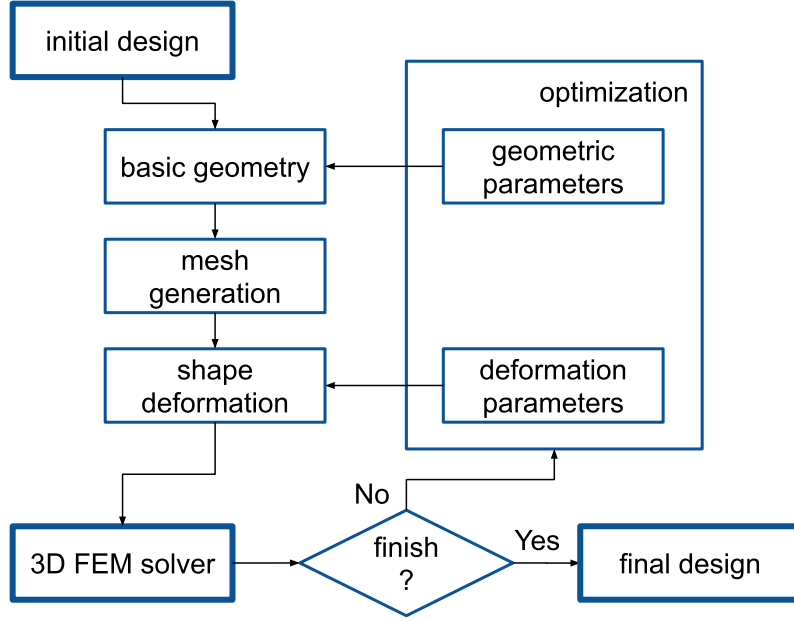


Fig. 3.7: A flowchart representing the proposed design optimization workflow with integrated shape deformation feature.

on the chosen solution type, the solver can compute the system response in the frequency domain (modal analysis) or determine the resonance frequencies and quality factors of the modes (eigenmode analysis). In each iteration, the optimizer evaluates the goal function according to the specified criteria, and performs a search of the parameter space to make another step. If the optimization method is gradient-based, this stage usually involves calculating the gradients of the goal function to determine the direction of the next step. The optimization proceeds until the specified termination conditions are met, that is, when the goal function value falls below a predefined threshold or when the maximum number of iterations has been reached. The resulting mesh of the deformed object can subsequently be converted into a 3D model file format, such as STL, and further processed using external geometry modeling software, e.g., to prepare production files for AM.

Most importantly, the optimization schemes are enhanced with fast sensitivity analysis to quickly evaluate the gradients of the defined system-response-based objective function. This approach can also be extended to design variables that control the shape deformation. This is because the deformation procedure operates on the same 3D mesh that is generated once for the baseline model. No re-meshing is required; therefore, the matrix structure and the problem size of the FEM system are preserved. To evaluate the sensitivity of the response with respect to the design parameters, the displacement of mesh nodes caused by a small perturbation of each design variable is first computed. Using this information, the corresponding perturbation of the global FEM matrix can then be determined. Since the mesh is perturbed rather than regenerated at this stage, the mesh topology and the structure of the system matrix remain unchanged after perturbation. This enables faster computation of the perturbed system solution, either by applying the adjoint sensitivity approach [91] or by using model order reduction techniques [92]. As a

result, rapid gradient evaluation is possible using the same precomputed symbolic factorization of the FEM system. The preservation of the matrix structure during sensitivity analysis is what makes gradient computation in the proposed optimization scheme highly efficient.

In summary, the advantages of the proposed design methodology are the following:

- unlocking a vast space of complex shapes, previously unexplored in microwave engineering: the presented approach allows for modeling and analysis of novel designs, which can now be fabricated using advanced manufacturing methods such as AM,
- simple and intuitive definition and control: shape deforming operations are managed by arbitrarily moving user-defined control points, a concept which is intuitive and does not require specialized knowledge of the underlying technique,
- a small number of design variables: the proposed FFD approach enables the creation of very complex models using only a fraction of parameters that would be required with other techniques such as parametric object representation (e.g., B-Splines or NURBS), thereby facilitating the optimization process,
- straightforward integration with EM simulation procedures: the employed RBF interpolation can be effectively performed on the same 3D mesh that is generated for FEM solvers,
- minimal computational cost: compared to the usual computational effort required for FEM solving, the cost of performing the RBF interpolation is negligible.

It should be emphasized that the outcomes of this work, in addition to being used in academic research on innovative microwave component designs, have led to the commercial implementation of shape deformation into the professional EM simulation software InventSim [8], where the shape deformation is available as an additional feature in versions 5.1.0 and later. Furthermore, the efficiency of the described optimization framework has been demonstrated on a waveguide bandpass filter design from Fig. 3.6 in a conference publication:

[MB1] **M. Baranowski**, Ł. Balewski, A. Lamecki and M. Mrozowski, "Fast Design Optimization of Waveguide Filters Applying Shape Deformation Techniques," in *Proceedings of the 24th International Microwave and Radar Conference (MIKON)*, September 2022, doi: 10.23919/MIKON54314.2022.9924969,

which was presented by the author at the 24th International Microwave and Radar Conference (MIKON) in Gdańsk, Poland, 2022. The full content of this contribution is provided in the following pages. In this paper, it is shown that the proposed framework can be efficiently employed to design microwave components with enhanced performance and exploit the potential of AM techniques. The paper demonstrates how the described integrated toolchain was used to optimize an inline sixth-order filter design, reaching a high performance from a poor starting point in just 10 iterations. It should be emphasized that this design utilizes two types of optimization

variables on two levels of object parametrization: geometric, and deformation, as explained above and visually summarized in Fig. 3.7. Moreover, the optimized design was successfully confirmed by measurement of a fabricated prototype, monolithically 3D-printed in metal from aluminum powder. The reported study can be treated as a first successful test of the new design concept, which was subsequently followed by many more advanced designs, all of them described in the following chapter of this thesis.

**[MB1] Fast Design Optimization of Waveguide Filters
Applying Shape Deformation Techniques**

M. Baranowski, Ł. Balewski, A. Lamecki and M. Mrozowski

24th International Microwave and Radar Conference (MIKON), Gdańsk, Poland, 2022

Author's contribution:

Implementation of the described design optimization framework: writing MATLAB scripts controlling geometry deformation operations, design & optimization of the presented waveguide filter, preparation of the model for 3D printing, prototype measurements.

Fast Design Optimization of Waveguide Filters Applying Shape Deformation Techniques

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Abstract—This paper presents an efficient design of microwave filters by means of geometry optimization using shape deformation techniques. This design procedure allows for modelling complex 3D geometries which can be fabricated by additive manufacturing (AM). Shape deforming operations are based on radial basis function (RBF) interpolation and are integrated into an electromagnetic field simulator based on the 3D finite-element method (FEM). The optimization controls two sets of design variables, i.e. the geometrical parameters of the structure and the parameters that determine the shape deformation. This approach provides great design flexibility at marginal additional cost since shape deformations require very little computational effort. A complete design procedure of a Ku-band RX waveguide bandpass filter is shown to verify this approach. The designed filter is fabricated with selective laser melting (SLM) technology. Measurements are presented in comparison with simulation.

Index Terms—waveguide filters, computer-aided design, design optimization, shape deformation, radial basis functions.

I. INTRODUCTION

Recent advances in microwave components design, among others, indicate a strong interest in new fabrication technologies such as additive manufacturing, also known as 3D printing [1]–[4]. AM enables the fabrication of complex 3D structures of almost arbitrary shapes, thus opening up new design possibilities for RF and microwave engineers. The introduction of more complex, smooth geometries enables new designs with enhanced electromagnetic (EM) performance, such as e.g. higher quality factor Q and low loss, wider higher-order mode separation in microwave filter design, or higher power handling. However, most of the widely used CAD software for high-frequency EM design does not offer sufficient tools to meet these new design opportunities. This is because the currently prevailing paradigm for geometry manipulation in EM CAD tools is constructive solid geometry (CSG), which is based on performing operations on simple building blocks such as boxes, cylinders, and cones. Structures which cannot be constructed from these geometrical primitives are not allowed, which can be a major limitation in the design procedure. In order to overcome these design restrictions and make greater use of the advantages offered by 3D printing

technology, the solution is to give RF designers more degrees of freedom in the modelling of microwave components. One of the ways to achieve this goal is to extend the conventional CAD based on CSG with shape deformation techniques [5].

Shape deformation of 3D geometries can be performed utilizing several techniques. This concept was first brought into use in computer graphics applications with the introduction of the free-form deformation (FFD) method in 1986 [6]. This approach, along with several other techniques which have been developed afterwards [7], [8], has then been applied to multiple areas of engineering [9]–[11]. Recently, shape deformations based on RBF interpolation have been introduced for RF and microwave components design [12].

The aim of this paper is to present fast and efficient optimization of waveguide filters using shape deformations in the design process. The new functionality of InventSim software combines RBF-based geometry modification with fast EM simulation with 3D FEM, as a result offering an all-in-one integrated optimization tool [13]. The design can be modified with two sets of optimization variables, namely the geometrical parameters of the structure and the variables controlling the deformation. The optimization procedure uses the zero-pole method, which has proven to be very efficient for microwave filter design [14], [15]. This strategy estimates the error using the location of zeros and poles of the filter's transfer and reflection functions. The proposed optimization procedure is applied to the design of a waveguide bandpass filter for satellite communications. The complete optimization process is described in Section IV-A. A prototype filter is fabricated using SLM process from aluminum alloy AlSi10Mg. The simulated model is verified by measurements in Section IV-B.

II. SHAPE DEFORMATION WITH RBFs

The shape deformation method used in this study is based on RBF interpolation, as shown in [9]. The principle of this operation is based on the definition and modification of some arbitrarily chosen control points. The deformation is performed by mapping the structure (defined by a surface or volume

mesh) according to the displacements of the control points. The mapping operation is represented in general by a mapping function $\mathcal{D}$:

$$\mathcal{D}(\mathbf{x}) : \mathbb{R}^3 \rightarrow \mathbb{R}^3$$

$$\mathbf{x} \mapsto \mathcal{D}(\mathbf{x}) = \mathbf{x} + \mathbf{d}(\mathbf{x}) \quad (1)$$

In this notation, $\mathbf{d}(\mathbf{x})$ is a vector displacement field which is applied to the 3D model. If radial basis functions are used in the process, the deformation field takes the following form:

$$\mathbf{d}(\mathbf{x}) = \sum_{i=1}^N \gamma_i \varphi(\|\mathbf{x} - \mathbf{x}_i\|) + p(\mathbf{x}) \quad (2)$$

where N is the number of control points defining the deformation, γ_i are the weights of radial basis functions φ , defined in the positions of the control points $\mathbf{x}_i$, and $p(\mathbf{x})$ is a low-order polynomial (optional). To define the mapping, one needs to define a set of N control points and their displacements, i.e., their initial and new (shifted) coordinates. Next, the linear system is solved to calculate the unknown weights γ_i . The constructed deformation field is then applied to smoothly deform the object.

III. GENERAL FRAMEWORK FOR DESIGN OPTIMIZATION WITH RBF INTERPOLATION

The proposed framework for the design optimization is shown in Fig. 1. The optimization loop consists of two design stages, each controlling separate design variables. The first stage involves managing the conventional geometrical parameters which are used to alter the basic shape of the structure that is constructed using CSG. The second stage is performed after mesh generation and comprises the shape deformation, which is controlled by dedicated parameters, i.e. the positions and displacements of the control points. Then, the deformed structure (volumetric mesh) is analyzed by an EM simulator to calculate its response in the frequency domain (S-parameters). Based on the frequency response of the filter, the optimizer evaluates the goal function, which is defined by the designer. We can use either the zero-pole based optimization scheme developed for filter design, or a general approach based on allowed and forbidden areas (masks) defined for transfer function. The optimization of the structure continues until certain termination conditions are met (e.g. error value below a preset threshold or maximum number of iterations reached). The proposed scheme is empowered with fast sensitivity analysis to evaluate the gradient of the cost function defined in the optimization, which also takes into account the design variables that define shape deformation. Such a low level integration significantly speeds up the optimization process.

IV. EXAMPLE OF FILTER DESIGN

The verification of the developed optimization procedure is carried out on an exemplary design of a Ku-band RX bandpass filter in waveguide technology. The specifications to be met by the filter are a wide passband of 13 – 14.8 GHz (center

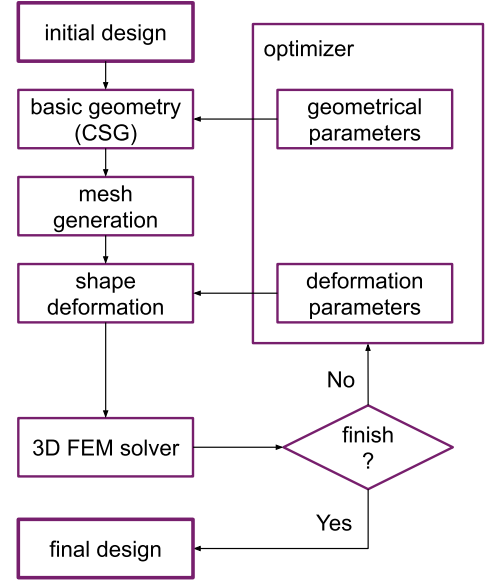


Fig. 1. A flowchart representing the general design optimization workflow.

frequency $f_0 = 13.9$ GHz, fractional bandwidth FBW = 13%) and an in-band return loss (RL) over 20 dB. The topology chosen for this design is a simple inline sixth order structure. To present the versatility of this design procedure, the initial structure constructed with CSG is a short section of a WR-75 waveguide with rounded corners, as shown in the upper left of Fig. 2.

A. Design Optimization

In this filter design, the inductive couplings between the cavities are realized by squeezing a straight waveguide section from both sides. The design is controlled by a total of 13 parameters: $L_1 - L_6$ describing the lengths of the cavities and $a_1 - a_7$ standing for the deformation depth of each iris section.

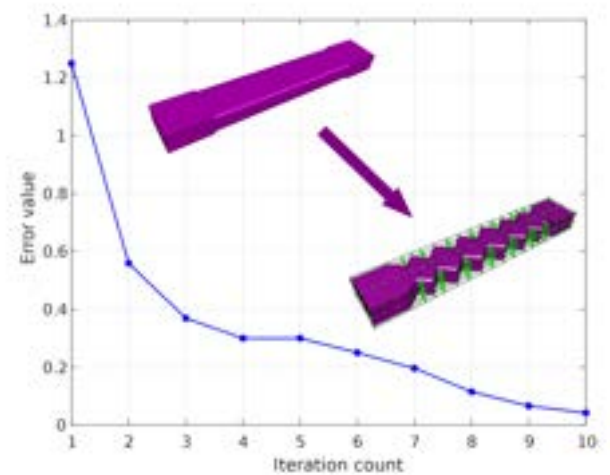


Fig. 2. Optimization error per iteration step. The basic (non-deformed) and deformed structure of the filter is also provided.

The length of each inductive iris section separating the cavities was set to a constant value of $d = 4$ mm. The positions of the control points were parameterized in a way the points are always placed in the middle of each iris section, following the changes of the resonators' lengths.

As mentioned previously, in this study the goal function compares the locations of zeros and poles extracted from the filter's transfer and reflection functions with reference values of the prototype filter [14]. Here, the starting point for the optimization task gave an error value of 1.25, whereas the goal was to reduce the error value below 0.05. The optimization was performed on Intel Core i9-10900X workstation with 64 GB RAM. The task took only 10 iterations and was completed in 47 minutes, resulting in a final error value of 0.042, as shown in Fig. 2. The S-parameters response of the initial and the optimized design is provided in Fig. 3. It can be seen that the framework is exceptionally efficient, leading to an optimized design even from a poor starting point in a very short time.

B. Fabrication and Measurement

The optimized design was prepared for fabrication and 3D-printed in one piece with AlSi10Mg alloy using SLM technology, as shown in Fig. 4. The fabricated component was measured and the results were compared with the simulations. Based on several previous prototypes manufactured using SLM, the effective conductivity of the metal, due to surface roughness, was estimated as $\sigma_{eff} = 4.2 \cdot 10^6$ S/m [12]. As shown in Fig. 5, the measurements show a very good agreement with the simulated model. This is a very successful result since no tuning was applied to obtain such a response. The passband is shifted ca. 50 MHz above the specified band, which is less than 3% of the bandwidth and could be corrected by tuning screws. The in-band RL of the filter is over 20 dB, and the insertion loss is 0.17 dB at f_0 and varies from 0.1 to 0.3 dB over the passband, which corresponds to a Q-factor of 1800. When compared to a classic sixth-



Fig. 4. A 3D-printed prototype of the filter.

order inductive waveguide filter design with the same effective conductivity σ_{eff} , the proposed design would provide a 20% higher Q-factor over the reference with only ca. 1.2% increase of total volume (17221 mm³ vs. 17011 mm³). To achieve a better effective conductivity and obtain a high Q-factor, the model could undertake surface polishing or silver plating. Nonetheless, this design shows the possibility of obtaining an improved EM performance by design based on smooth geometry deformations.

V. CONCLUSION

A fast design optimization of waveguide filters using shape deformations was presented. The proposed all-in-one integrated CAD tool gives RF and microwave engineers an easy and efficient way to design novel microwave filters with enhanced electromagnetic performance. This design framework can be used to exploit the full potential of additive manufacturing techniques. In the future, geometry deformation

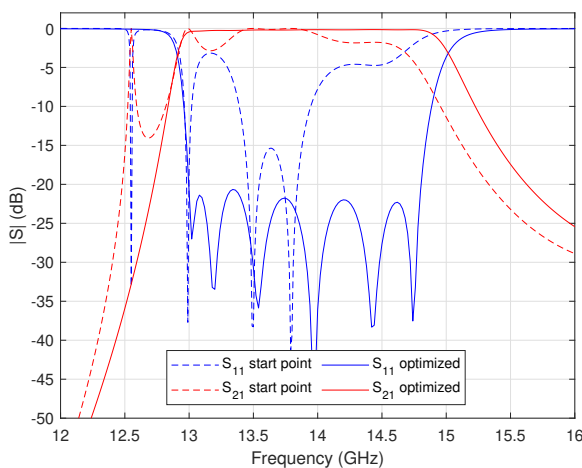


Fig. 3. S-parameter response of the filter. Dashed lines - initial design, solid lines - optimized design.

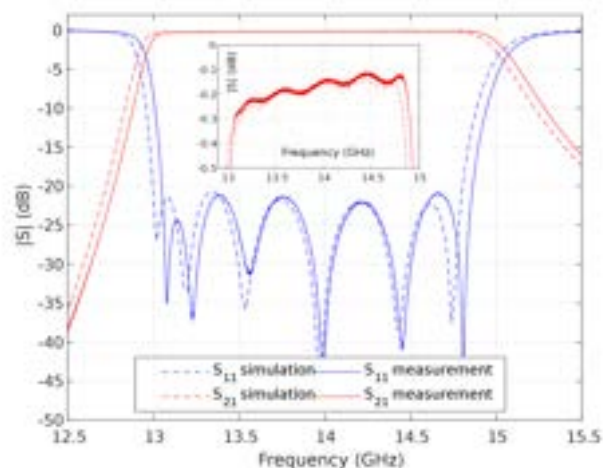


Fig. 5. Comparison of simulation and measurement results. Dashed lines - simulation, solid lines - measurement.

will be applied to design more complex passive components, such as waveguide filters with more stringent specifications, complex topologies, additional transmission zeros, or couplers and multiplexers.

ACKNOWLEDGMENT

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Chapter 4

APPLICATION OF SHAPE DEFORMATION TO EM DESIGN

This chapter describes the main achievements of this study, namely the outcomes of applying the proposed shape deformation approach to the design of novel microwave components with enhanced RF performance. The validity of the presented design methodology, and thus the fulfillment of the claims identified in Section 1.2, has been demonstrated in five original scientific articles published in international journals of the Institute of Electrical and Electronics Engineers (IEEE). A complete list of these publications, [MB2, MB3, MB4, MB5, MB6], and their brief descriptions can be found below. The findings of each of the five publications have been summarized and supplemented in the subsequent sections of this chapter—each section devoted to the research related to a specific article. A portion of the reported work, documented in publications [MB2, MB3], was conducted within a research project funded by the European Space Agency, carried out in collaboration with external partners. The project entitled "*RF filters with Enhanced Electromagnetic Performance for Space Applications Exploiting Shape Optimization with Free-Form Deformation and Additive Manufacturing Techniques*," (contract No. 4000128097/19/NL/CBi) focused on the joint application of shape deformation and 3D-printing techniques. This research area is identified as strategically important, particularly for advancing new solutions for satellite communication systems.

[MB2] M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski, and J. Galdeano, "The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques," *IEEE Transactions on Microwave Theory and Techniques*, vol. 71, no. 7, pp. 3065-3074, July 2023, doi: 10.1109/TMTT.2023.3239363,

This article introduces and experimentally verifies shape deformation as a new approach to computer-aided design of passive RF components. It is demonstrated that geometry deformation techniques provide additional degrees of freedom in the 3D modeling of microwave structures, which is highly desirable in order to exploit the full potential of AM of high-frequency electronic devices. The proposed framework is used to design novel components with enhanced RF performance: cavity resonators with high quality (Q) factor, bandpass filters with improved spurious mode separation, and a compact, 90° twisted waveguide filter. Three filter designs with complex geometry were verified by 3D-printed prototypes fabricated in metal with SLM. The measurement results confirmed the effectiveness of the new design approach and demonstrated the possibilities offered by 3D printing.

[MB3] M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication Systems," *IEEE Microwave and Wireless Components Letters*, vol. 32, no. 12, pp. 1403-1406, Dec. 2022, doi: 10.1109/LMWC.2022.3193722,

This paper presents a novel microwave filter design utilizing a modified circular waveguide dual-mode (CWDM) cavity shape, applicable in satellite communication systems. The modification leads to a wide spurious-free stopband, which is highly desirable for channel separation in waveguide contiguous output multiplexers (OMUXs). The new resonant cavity design is a result of applying shape deformation to the basic cylindrical cavity in order to improve its out-of-band performance. A highly selective narrowband fourth-order Ku-band channel filter is designed, 3D-printed in metal as a single piece and measured. The fabricated prototype demonstrates a very wide upper stopband, which would allow the realization of a wideband manifold-coupled OMUX without the insertion of any additional lowpass filters.

[MB4] M. Baranowski, Ł. Balewski, A. Lamecki and M. Mrozowski, "Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators," *IEEE Access*, vol. 12, pp. 78275-78285, 2024, doi: 10.1109/ACCESS.2024.3408671,

This paper presents a new smooth-profile waveguide filter design based on specially formed dual-mode (DM) cavity resonators. The proposed DM cavity is designed using the shape deformation method. The coupling between the two orthogonal cavity modes is achieved by breaking the symmetry of the structure, thereby eliminating the need for additional coupling elements. Two filter prototypes employing the designed cavities are developed: a fourth-order bandpass filter (BPF) with a 90° rotated output and a transmission zero (TZ), and an eighth-order BPF with four TZs. The designs are developed taking into account the limitations of 3D printing to enable their single-piece fabrication without internal supports. Both filter designs were experimentally verified through prototypes manufactured using the SLM process. The two reported structures exhibit low insertion loss, relatively high Q and a smooth profile, making them potentially useful for low-loss, high-power applications.

[MB5] M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Álvarez-Botero, A. Lamecki, M. A. G. Laso, and P. Martín-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," *IEEE Microwave and Wireless Technology Letters*, vol. 35, no. 9, pp. 1292-1295, Sept. 2025, doi: 10.1109/LMWT.2025.3578244,

This letter reports on a novel design for an air-filled, self-supported coaxial-line filter featuring a smooth profile, oriented to 3D printing in metal. The filter is intended for Earth observation data downlink systems, where it must effectively reject signals in a wide frequency range. The design incorporates a smooth-profile low-pass filter (LPF) structure, integrated with a short X-band BPF section with four quarter-wavelength short-circuited stubs. The optimization of the LPF is attained by means of shape deformation and leads to a wide rejection band up to around 40

GHz, to suppress the third harmonic and other undesired out-of-band frequencies. A very compact coaxial-line filter prototype was manufactured in one piece using SLM. The fabricated component meets the stopband demanded by current Earth observation satellite systems and exhibits the widest stopband rejection among the related state-of-the-art solutions.

[MB6], J. Sorocki, I. Piekarz, **M. Baranowski**, A. Lamecki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024, doi: 10.1109/TMTT.2024.3361976,

In this work, a novel and cost-effective polishing technique is introduced for reducing the internal surface roughness of additively manufactured components, specifically developed for waveguide circuits operating in the millimeter-wave frequency range. Waveguide structures produced by powder bed fusion processes such as SLM typically exhibit surface roughness on the order of 10–50 microns, which significantly increases conductor losses associated with surface irregularities and thereby degrades their electrical performance. To mitigate this effect, a low-cost method using glass micro-beads as an abrasive medium in combination with a rotary tumbler is proposed. This approach enables efficient penetration of the abrasive medium into internal channels and cavities with cross-sectional dimensions reaching sub-millimeter scale. In contrast to electro-polishing, neither chemicals nor sophisticated equipment is necessary, while similar benefits of good internal channel penetration and polishing are still provided. An experimental study, carried out by principal investigators from AGH University of Krakow, was conducted on several waveguide sections and bandpass filters fabricated through SLM, operating within the 8.2–40 GHz range. Investigation of polishing impact on both mechanical and electrical properties showed a 10%–20% loss reduction within tested waveguide sections and finally a 10%–40% insertion loss reduction for the tested filters. Within this study, the author's contribution focused on benchmark waveguide filter designs optimized for additive manufacturing, which served as test units in the experimental investigation.

4.1. *The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques*

The first article in this compilation, published in *IEEE Transactions on Microwave Theory and Techniques*, describes the fundamentals of the proposed shape deformation method for microwave component design. In this paper, shape optimization based on FFD is presented for the first time, and integrated into the design process of cavity resonators and filters. Mesh deformations are coupled with EM field simulations, as described in Section 3.2. The aim of this work is to show that the proposed approach can overcome current limitations in RF design, and thereby improving key parameters of the developed electronic circuits, such as unloaded quality factor, Q_u , spurious mode separation, compactness and mass reduction. Such features are highly desirable in the design of microwave filters for space applications, such as output multiplexers or transmit and receive signal paths in satellite communications systems. The methodology is validated by three experiments. First, an investigation of cavity resonators was conducted, exploiting shape deformation to improve the Q-factor and spurious-free performance. Then, the concept was extended to the design of waveguide cavity filters. Two designs were examined: a third-order and sixth-order BPF, utilizing vertically squeezed cavities to achieve wider out-of-band performance. Finally, the FFD approach was used to design an integrated filtering waveguide twist component, that is, a 90°-twisted waveguide BPF. Three prototypes were manufactured in a single mechanical part in aluminum alloy with SLM to validate the presented designs. The findings presented in the article are summarized and expanded with supplementary information in the remainder of this section.

4.1.1. *Optimization of cavity resonators using shape deformation*

The first practical application of FFD to EM design considered the modification of the shape to design cavity resonators with enhanced RF performance. In particular, shape deformation techniques were used to optimize the geometry of a rectangular waveguide resonant cavity in order to investigate the possibility of maximizing the quality factor as well as higher-order mode separation. To this end, an in-depth study involving a series of optimization tasks was carried out. In each iteration, the resonance frequencies and Q-factors of the modes excited in the analyzed cavity were calculated by performing an eigenmode simulation. In order to perform the optimization, the following three objectives were defined:

1. maintaining the resonance frequency of the fundamental mode, f_0 (with given accuracy),
2. maximizing the quality factor Q of the fundamental mode, and
3. maximizing the spurious-free range, Δf , defined as the absolute difference between the first higher-order mode resonating at f_1 and the fundamental one at f_0 .

The experiment focused on obtaining the highest possible separation between the modes (Δf),

while keeping the resonance frequency of the operating mode fixed at a given f_0 and also achieving a desired Q . The value of the finite electric conductivity chosen for the simulations is equal to that of aluminum ($\sigma = 3.8 \cdot 10^7$ S/m), and the desired resonance frequency of the cavity was set to $f_0 = 10$ GHz. The studied resonant cavity, shown in Fig. 4.1a, was modified with the aid of FFD. In order to alter the geometry of the structure, a set of control points, 26 in total, was defined, as shown in Fig. 4.1b. The vertical squeezing operation was declared by assigning two offset vectors at the central points of the cavity's upper and lower faces, directed toward the cavity interior. Given the control points and displacement vectors, the deformation can be performed as described in Section 3.1, resulting in a smoothly deformed structure shown in Fig. 4.1c. The transformed 3D mesh is then used for the eigenmode analysis. The design variables used in the optimization include three geometrical parameters, i.e., cavity height, length, and width, as well as one deformation parameter controlling the squeezing operation.

The optimization procedure was executed multiple times with the objective of targeting different values of Q and Δf , yielding a set of resonator geometries that all maintained the same resonance frequency of $f_0 = 10$ GHz. The outcomes are presented in a chart in Fig. 4.2, demonstrating a clear trade-off between the achievable Q and the mode separation $\Delta f = f_1 - f_0$. Two cavity geometries representing the extreme cases from this chart are also depicted in Fig. 4.2. The results indicate that increasing the higher-order mode separation can be achieved at the expense of a reduced Q -factor. Thus, depending on the requirements of a given application, an appropriate resonator geometry providing the desired combination of Q and Δf can be selected—by controlling the deformation of the cavity, one gains the flexibility in the choice of the desired $Q/\Delta f$ ratio.

The same optimization was run one more time for an extreme scenario, targeting the maximum achievable mode separation at the expense of Q , but maintaining the resonance at f_0 and a smooth cavity surface. The solution which was obtained is shown in Fig. 4.3, for which the distance between the fundamental mode and the first spurious one is 14.4 GHz, i.e., the f_1/f_0 ratio equals 2.44. The Q of the fundamental mode, considering the finite conductivity of aluminum, is equal to 4148. As can be inferred from the results presented in Fig. 4.2 and Fig. 4.3, there is a

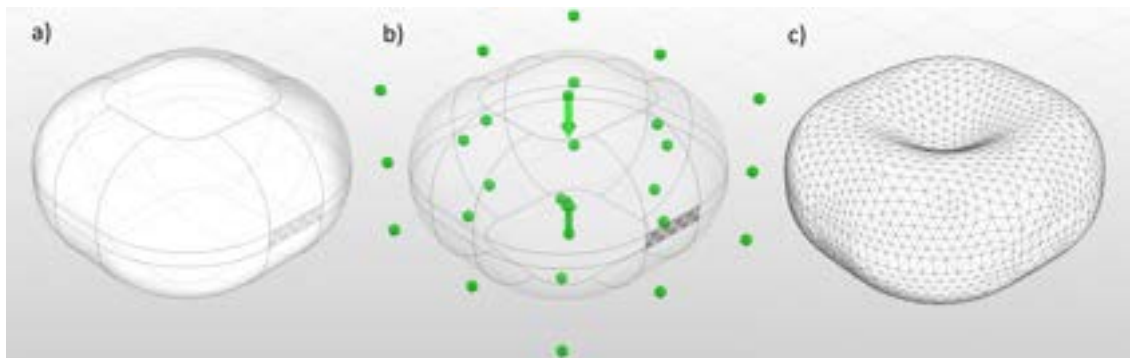


Fig. 4.1: Shape deformation of the cavity: a) basic rectangular cavity with rounded edges, b) definition of the control points (green dots) and vertical squeezing operation (green arrows), c) the resulting 3D mesh.

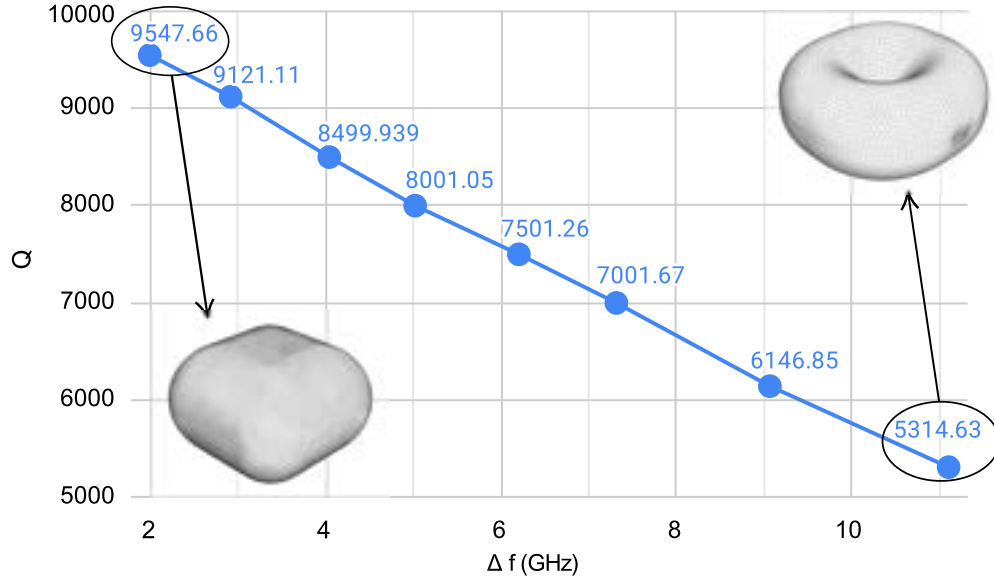


Fig. 4.2: Trade-off chart: resonator Q -factor vs. higher-order mode separation for $\sigma = 3.8 \cdot 10^7$ S/m. Two cavity models are shown for extreme cases: highest Q (bottom left) and widest mode separation (upper right).

clear dependence of the cavity shape on Q and Δf ; when aiming for high Q , the geometry of the cavity approaches a spheroid, while when targeting a maximum Δf , it evolves into an incomplete toroid, or a pinched sphere with an hourglass-shaped cross-section. Consequently, the use of FFD facilitates efficient exploration of the design space, allowing the selection of geometries that best satisfy specific performance requirements.

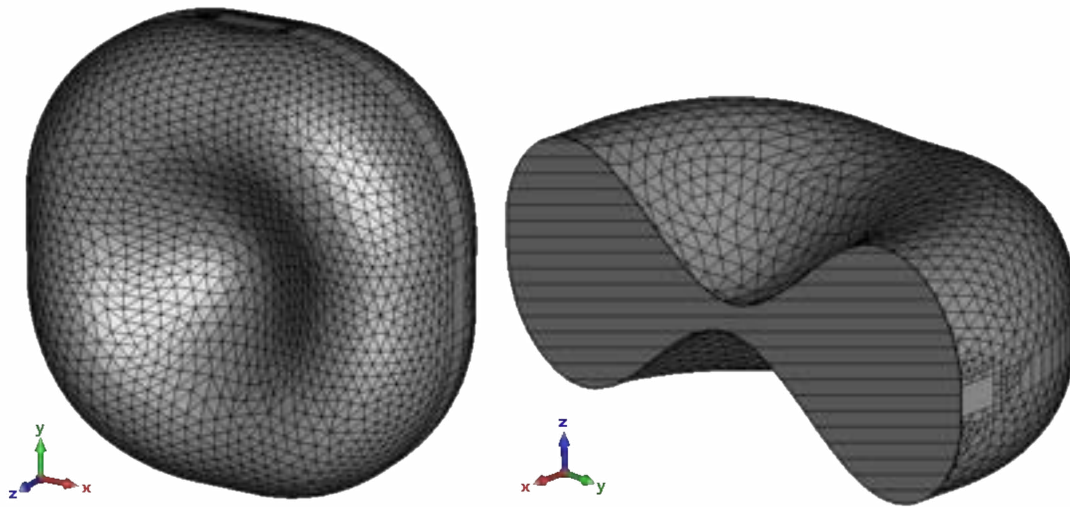


Fig. 4.3: Geometry of the resonant cavity (left) and its cross-section (right), exhibiting the widest achieved spurious-free range, $\Delta f = 14.4$ GHz, with $f_0 = 10$ GHz and Q equal to 4148 (assuming $\sigma = 3.8 \cdot 10^7$ S/m).

4.1.2. Design of inline rectangular waveguide cavity filters using shape deformation

The cavity resonator analysis presented in the previous section has proven that shape deformation can be successfully implemented in a practical scenario to enhance the capabilities of computer-aided EM design. Naturally, what follows is the adoption of these outcomes to design cascaded resonator filters based on shape-modified rectangular waveguide cavities. Two filter designs were developed, aiming for the requirements of Ku-band broadband input filters for satellite communication systems [93]: center frequency $f_0 = 14.25$ GHz, bandwidth (BW) equal to 500 MHz, which gives a fractional bandwidth (FBW) of 3.5%, and a return loss level greater than $RL \geq 20$ dB. Both filter designs were experimentally verified by prototypes which were additively manufactured in metal with SLM. The main objective of this study was to apply shape deformation techniques to achieve an improvement in terms of out-of-band rejection of the filters. Although microwave filters with superior spurious-free performance have previously been reported in other technologies (e.g., combline, ceramic, and coplanar waveguide filters [94–97]), this work emphasizes the improvements attainable through a simple geometric deformation, enabling fabrication of the final structure as a single 3D-printed piece without internal supports.

At this point, the specific considerations of the 3D printing process itself should once again be emphasized. Although AM opens up new design possibilities in microwave engineering, as every other technology, it comes with its own limitations which need to be accounted for, as discussed in Section 2.3.1. When considering 3D printing directly in metal, the main challenges concern the attainable surface roughness and the ability of single-piece fabrication along a given printing direction. The former will influence greatly the effective metal conductivity and hence the insertion loss (IL), and can be treated during post-processing—as will be studied and described in detail in Section 4.5. The latter case, however, has to be taken into consideration very early on, at the design stage, as it imposes additional constraints on the allowable geometry of the model. Therefore, in both designs presented herein, the shape of the cavities was adapted so as to allow for single-piece AM of the prototypes. That is, to avoid overhanging walls, all resonant cavities used in the following filter designs have been modeled to have a quasi-triangular shape, chamfered along the printing direction.

The first filter circuit which was developed using FFD techniques is a third-order inline BPF. The initial design, which serves both as a reference and a starting point in the design process, is a conventional inductive iris H -plane waveguide filter, which is well known in the literature [98]. The coupling-routing diagram, the synthesized coupling matrix of the circuit and its corresponding ideal filtering characteristic are given in Fig. 4.4, along with the EM models of the designed filters. The inductive iris filter, presented in Fig. 4.4d, was implemented in the WR-75 rectangular waveguide standard for Ku-band operation. The design was subsequently adapted for single-piece fabrication via SLM by reshaping the cavities, as depicted in Fig. 4.4e. Finally, vertical deformation of the cavities was introduced to improve out-of-band performance and move away the higher-

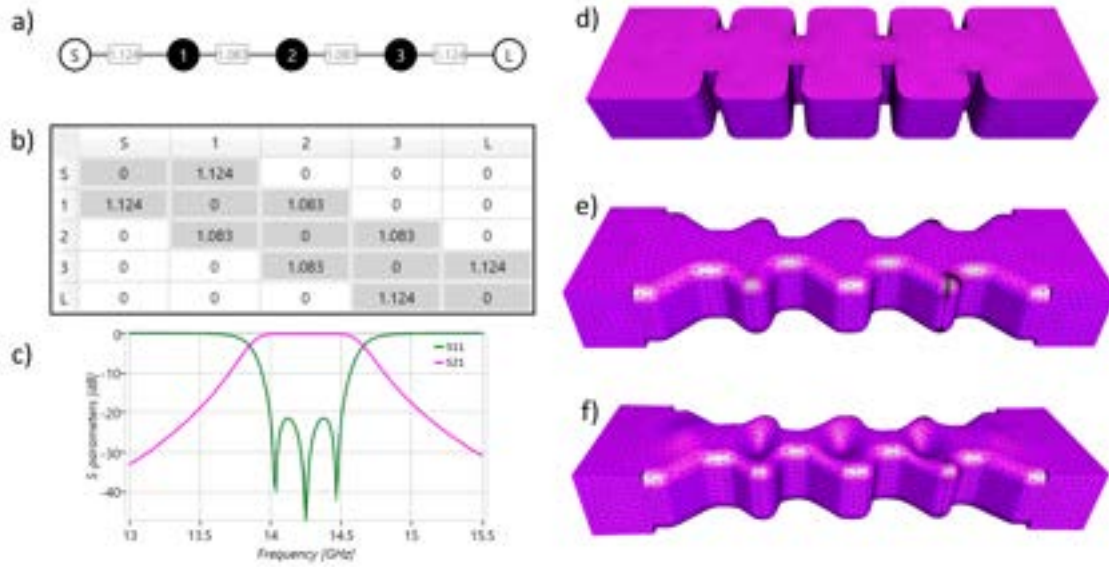


Fig. 4.4: Design of the third-order filter: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) EM model of the standard iris filter, e) EM model of the modified AM-compliant filter, f) final design with deformed cavities.

order modes. The deformation depth, the same for each cavity, was equivalent to 2.5 mm, and the RBF used for interpolation was a Gaussian basis function as in eq. (3.3) with radius $r = 4$ mm.

The design process involved geometric optimization, incorporating the parameters of the basic structure—namely, resonator lengths and iris widths—together with simultaneous shape deformation. The optimization was performed using the zero-pole method implemented in InventSim software, which has proven very efficient for automated design tuning of microwave filters [99]. In short, the method compares the locations of zeros and poles of the optimized filter with the zeros and poles corresponding to the ideal filter characteristic. The whole procedure was carried out uti-

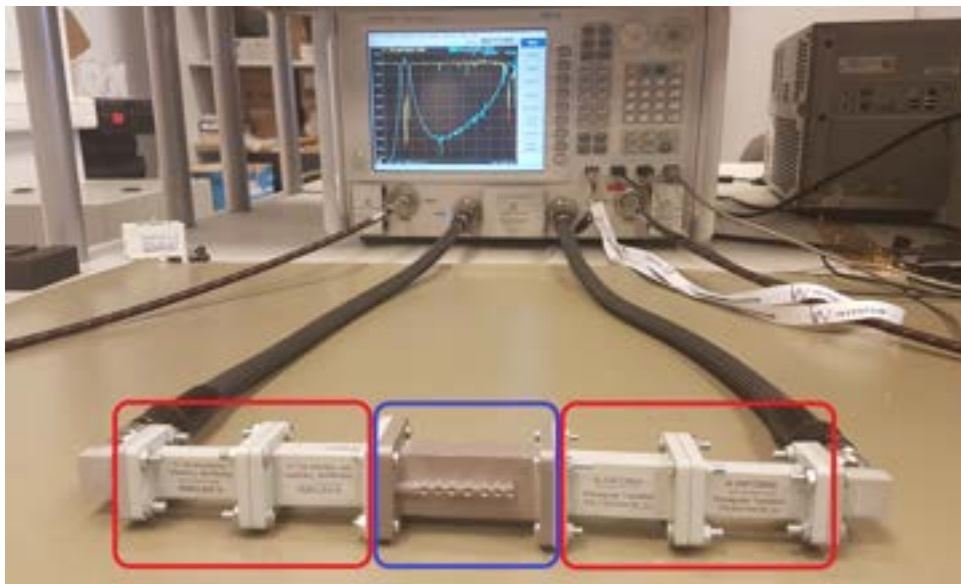


Fig. 4.5: Measurement setup using WR-51-to-WR-75 waveguide adapters (red). Measured prototype: third-order filter (blue).

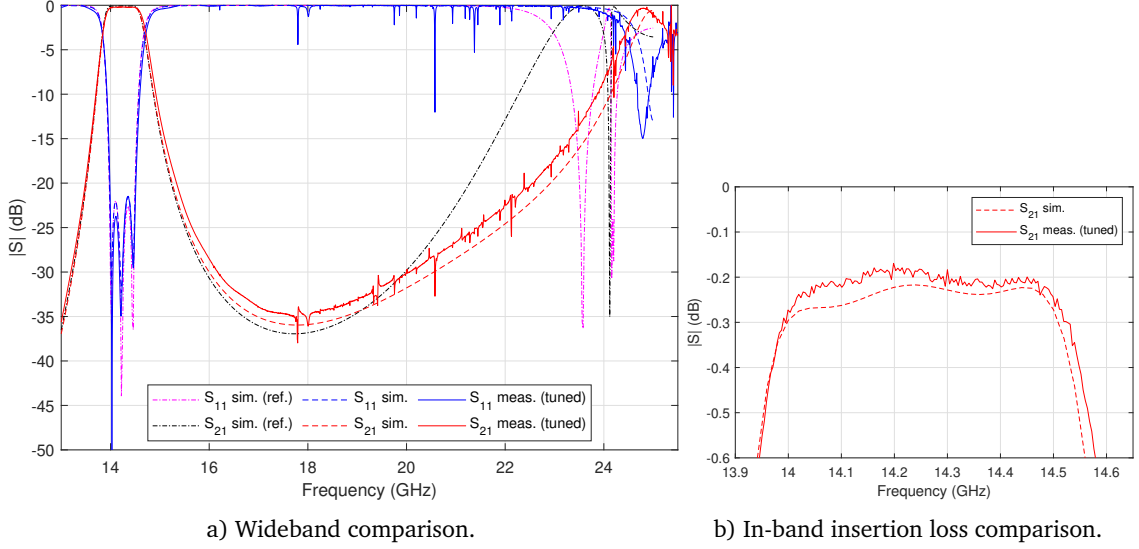


Fig. 4.6: A comparison of simulations and measurement of the third-order BPF: reference H -plane filter design: dash-dotted lines; optimized design with deformed cavities (EM simulation): dashed lines; measurements (tuned): solid lines. The IL simulation in (b) considers a lossy material with $\sigma_{\text{eff}} = 4.2 \cdot 10^6 \text{ S/m}$.

lizing the framework described in Section 3.2 that supports efficient optimization with concurrent shape deformation, including adjustment of control point locations and their displacements. The resulting filter geometry is presented in Fig. 4.4f.

A prototype of the filter was manufactured using the SLM process with a Trumpf TruPrint 1000 printer [100] using an aluminum-silicon-based alloy (AlSi10Mg). To measure the filter's performance across a wide frequency range, WR-51-to-WR-75 waveguide adapters were used to prevent the influence of higher-order modes typically excited in WR-75 waveguide-to-coax transitions. The printed prototype and the whole measurement setup using a vector network analyzer (VNA) and the aforementioned adapters is shown in Fig. 4.5. The comparison of the measurements with full-wave EM simulation is presented in Fig. 4.6. The simulated response of the reference design—the standard iris filter shown in Fig. 4.4d—is also included in the plot. The measured response matches the EM simulation very well: the in-band RL (after manual tuning) is greater than 21 dB, and the IL at f_0 is equal to 0.2 dB. Based on the measured insertion loss of the printed prototype, the effective conductivity of the 3D-printed aluminum alloy was approximated to be $\sigma_{\text{eff}} = 4.2 \cdot 10^6 \text{ S/m}$. This value was then used for the EM simulation in Fig. 4.6b and in the following simulations of the subsequent designs for obtaining a more accurate loss estimation. The measured spurious-free region extends to 24.5 GHz, which translates into an improvement of approximately 1.5 GHz compared to the reference design. Although relatively small, this enhancement confirms the validity of the proposed approach, and leads to a more significant gain in the out-of-band performance obtained with the subsequent filter design, as detailed later in this section.

The second design follows the same idea which was presented above, but for a sixth-order BPF. The design specifications are the same as for the third-order filter, that is: $f_0 = 14.25 \text{ GHz}$, BW

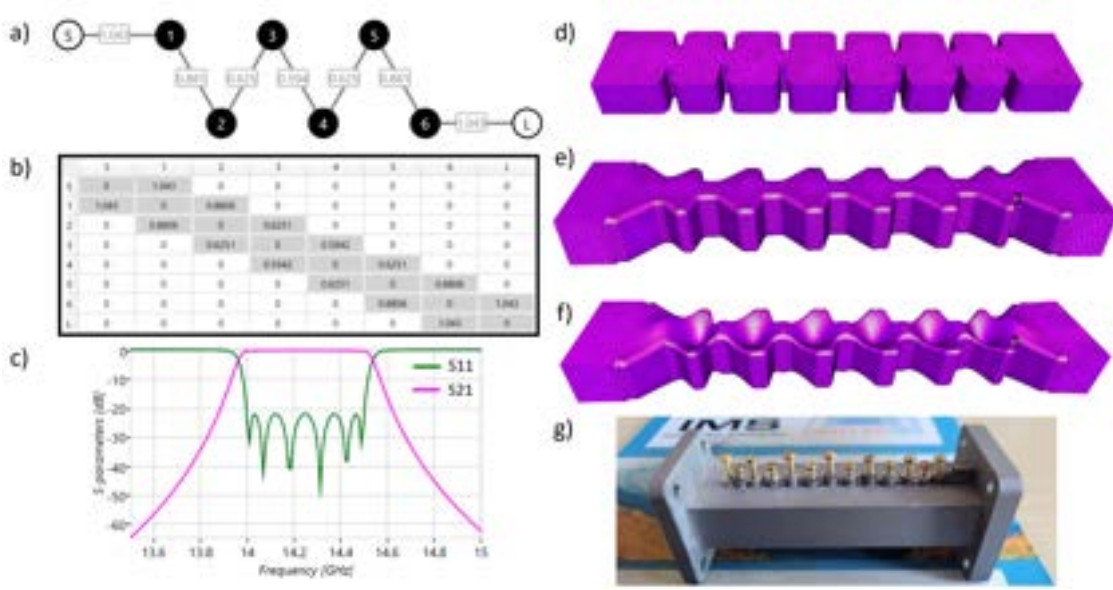


Fig. 4.7: Design of the sixth-order filter: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) EM model of the standard iris filter, e) EM model of the modified AM-compliant filter, f) final design with deformed cavities, g) photograph of the manufactured prototype with tuning screws.

= 500 MHz and $RL \geq 20$ dB. The positive effect of cavity squeezing on moving away higher-order modes in the filter's stopband was fully exploited by applying a deeper deformation of the cavities (3.75 mm). Once again, the design procedure started with the synthesis of a sixth-order in-line BPF, for which the coupling-routing diagram, coupling matrix and corresponding ideal response are presented in Fig. 4.7a-c. The filter was implemented in the WR-75 waveguide, as shown in Fig. 4.7d. Afterwards, the filter was redesigned to be fit for direct fabrication with AM, which was again done by adjusting the shape of the resonant cavities, as demonstrated in Fig. 4.7e. Finally, the modified filter from Fig. 4.7e was optimized by applying shape deformation using the proposed framework, yielding a structure shown in Fig. 4.7f. The prototype of the filter, depicted in Fig. 4.7g, was additively manufactured with SLM from aluminum alloy and was measured using the same setup as in Fig. 4.5, including the WR-51-to-WR-75 adapters to capture the wideband characteristics. The measurements are presented in Fig. 4.8 along with EM simulation and the response of the reference H -plane filter. From the comparison of wideband responses in Fig. 4.8a it is evident that the new design significantly extends the spurious-free region with respect to the reference, clearing the upper stopband up to around 27.2 GHz, which translates to $1.9 f_0$, effectively providing about 4 GHz wider out-of-band rejection than the standard iris filter from Fig. 4.7d. The passband performance of the printed component can be assessed in Fig. 4.8b. The in-band IL, measured at central frequency f_0 , is equal to 1.15 dB. In comparison with a lossy simulation assuming effective metal conductivity σ_{eff} , this value is higher than expected, but it can be attributed to the presence of brass tuning screws.

It should be noted that the two designs just described were the first microwave components developed and manufactured as part of this thesis using AM technology. Since then, new ways to

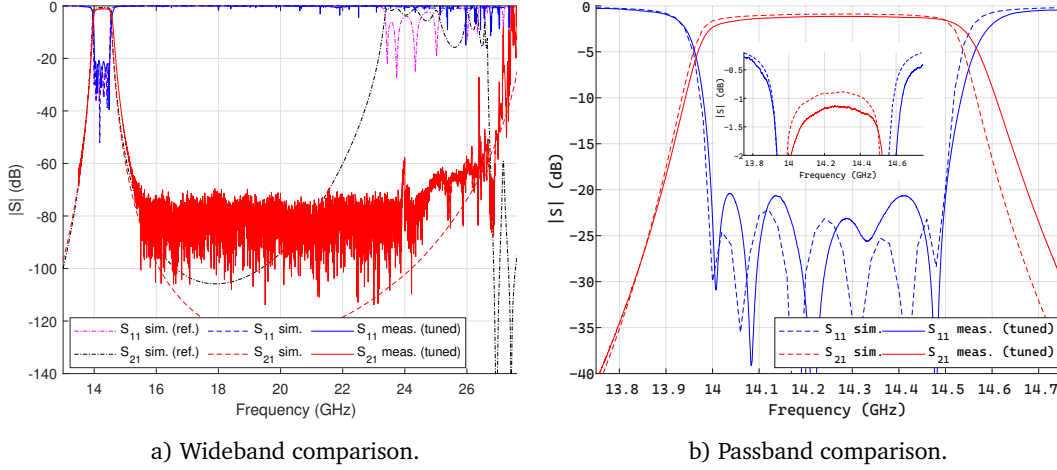


Fig. 4.8: A comparison of simulations and measurement of the sixth-order BPF: reference H -plane filter design: dash-dotted lines; optimized design with deformed cavities (EM simulation): dashed lines; measurements (tuned): solid lines. The IL simulation in (b) considers a lossy material with $\sigma_{\text{eff}} = 4.2 \cdot 10^6 \text{ S/m}$.

reduce losses have been sought, e.g., through the use of high-quality manufacturing processes tailored to microwave engineering applications [101] or post-printing treatments such as surface polishing, addressed in Section 4.5.

4.1.3. 90° twisted sixth-order waveguide filter

The third and last component described in the first publication [MB2] is a twisted waveguide filter. The goal of that example was to further demonstrate the capabilities of EM design enhanced by FFD: Here, shape deformation was utilized to integrate a 90° waveguide twist and a bandpass filter into one compact device. Since the adoption of 3D-printing techniques in microwave engineering, there have been several attempts at developing such integrated components, taking advantage of new manufacturing capabilities [29, 59]. This example shows that with the aid of shape deformation techniques, the computer-aided design process of such structures is not only feasible, but also very effective. The twisted filter was designed in several steps. The first stage involved the design of a sixth-order Ku-band BPF with a smooth profile, as first described in [MB1]. The specifications of the filter are: $f_0 = 13.9 \text{ GHz}$, $\text{BW} = 1800 \text{ MHz}$ ($\text{FBW} = 13\%$), $\text{RL} \geq 20 \text{ dB}$. The synthesized design is presented in Fig. 4.9a-c. The filter was developed by squeezing the side walls of a straight WR-75 waveguide (as in Fig. 4.9d) to shape inductive irises. The result, visible in Fig. 4.9e, is a specially modeled H -plane iris filter suitable for 3D-printing—all its walls are inclined so that direct AM along the direction of propagation is possible without extra supports or overhanging parts. In the second design stage, the filter was twisted by applying a deforming operation to the structure from Fig. 4.9e. The deformation was defined by 24 control points: twelve at the input and twelve at the output of the waveguide filter model. The points at the input flange were fixed, whereas those at the output were dislocated to introduce a gradual rotation of the waveguide along its propagation direction. The design was then optimized in several

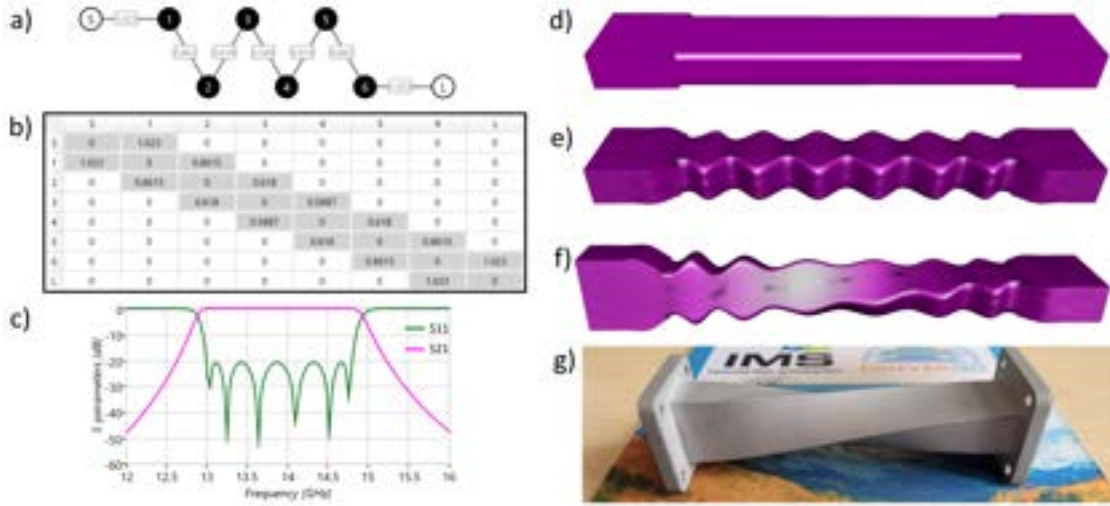


Fig. 4.9: Design of the twisted filter: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) starting point: straight WR-75 waveguide, e) EM model of the straight AM-compliant BPF, f) final twisted filter design, g) photograph of the manufactured prototype.

optimization rounds, by incrementally increasing the rotation angle from 0 to 90 degrees: that is, at first the twist angle was set to 15° and the passband response was recovered by optimization; then, the angle was increased to 30° and so on. Ultimately, the final BPF design was reached with a full 90° rotation between the input and output of the waveguide, as shown in Fig. 4.9f. The prototype, pictured in Fig. 4.9g, was 3D-printed using SLM with AlSi10Mg alloy and measured. No manual tuning was involved in this case. The comparison of the measurement and simulation is provided in Fig. 4.10. Apart from one mismatch point at 13.1 GHz, the passband performance of the filter is very satisfactory: the in-band RL is over 19 dB and the IL between 0.1 and 0.4 dB. The degraded performance in the mismatched region could be recovered by tuning screws. Nev-

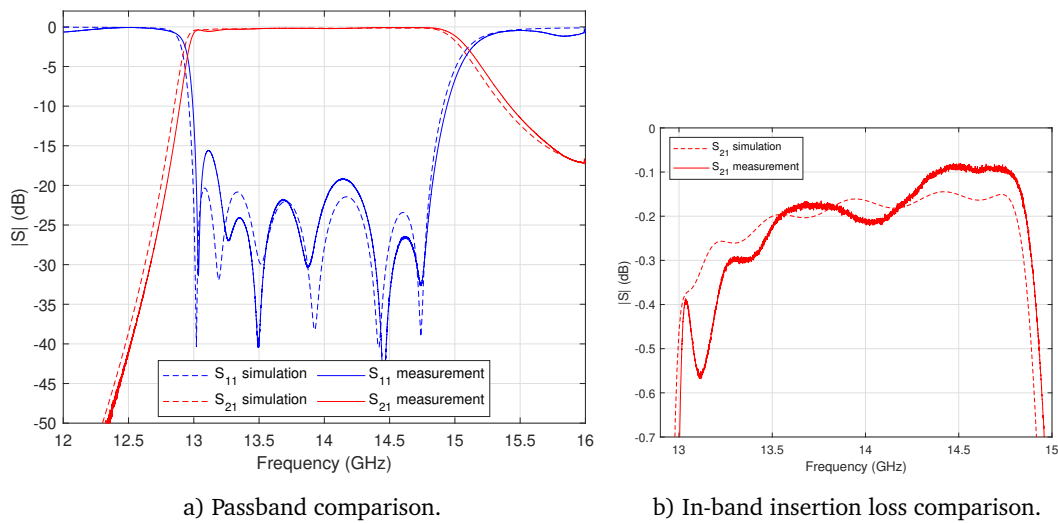


Fig. 4.10: A comparison of simulations and measurement of the twisted BPF: EM simulation: dashed lines; measurements: solid lines. The IL simulation in (b) considers a lossy material with $\sigma_{\text{eff}} = 5 \cdot 10^6 \text{ S/m}$.

ertheless, the fabricated test unit confirms the viability of applying FFD techniques to design this class of waveguide structures.

As demonstrated by the three cases presented in the publication, the proposed optimization framework facilitates the design of microwave components with a wide range of complex geometries and leads to promising results. Therefore, the full potential of this concept has been further explored and described in the remainder of this chapter.

[MB2] The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques

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Author's contribution:

Implementation of the described design optimization framework: writing MATLAB scripts controlling shape deformation operations; investigation of cavity resonators modified using shape deformation techniques, design of the three waveguide filters presented in the paper, preparation of the 3D models for AM, measurements of the prototypes, manual tuning of the two filters described in Sections V.A. and V.B.

The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques

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Abstract—This paper introduces shape deformation as a new approach to the computer-aided design (CAD) of high-frequency components. We show that geometry deformation opens up new design possibilities and offers additional degrees of freedom in the 3D modeling of microwave structures. Such design flexibility is highly desirable if the full potential of Additive Manufacturing (AM) is to be exploited in the fabrication of RF & microwave devices. The use of deformation techniques in the design of high-frequency components allows the attainment of improved electrical parameters, such as high-quality factors for cavity resonators and wide higher-order mode separation. In this work, shape deformation with radial basis functions (RBFs) is integrated with an electromagnetic field simulator based on the 3D finite-element method (FEM), allowing the semiautomated optimization of microwave components, such as cavity resonators and filters. The proposed strategy is used for the design of high Q-factor cavity resonators, cavity bandpass filters with improved spurious mode separation and of a compact twisted waveguide filter. Three designs of waveguide cavity filters with complex geometry are experimentally verified using 3D-printed prototypes fabricated with the selective laser melting (SLM) technology.

Index Terms—cavity resonator, CAD, design optimization, shape deformation, radial basis functions, additive manufacturing, microwave filter.

I. INTRODUCTION

The use of additive manufacturing technology (also known as 3D printing) to fabricate high-frequency components has been increasing in popularity in recent years [1]–[4]. AM allows for the production of complex 3D structures of almost arbitrary shapes, with user-defined surface curvatures. This flexibility in model complexity is a great advantage over established technologies, as the limitations of conventional techniques such as CNC milling no longer apply. In many cases, AM permits the production of a model that would not be achievable using other manufacturing methods. In this context, AM is a key enabling technology that opens up new

possibilities to RF & microwave engineers. However, in order to take advantage of this design flexibility, the limitations of the CAD tools currently used in the industry need to be overcome. These limitations stem from the fact that virtually all CAD software used for high-frequency electronic design relies on the constructive solid geometry (CSG) approach. CSG is the prevailing paradigm for the 3D modeling of electromagnetic structures. It is based on using a set of simple, predefined geometrical primitives—such as boxes, cylinders, and cones—and applying to them Boolean operations, such as union, subtraction, and intersection, or shape modifications with edge filleting or chamfering. Alternatively, arbitrarily shaped surfaces based on parametric, equation-based objects can be used, but this is cumbersome and difficult to use. The design of more complex geometrical models—such as those with unconventional twisted or bent shapes—is very challenging in this approach. For decades, the CSG approach has been regarded as satisfactory, as design was limited by the available manufacturing technologies, like CNC milling. This has now changed with the widespread availability of repeatable and accurate 3D printing.

To date, many novel microwave components have been developed using AM. However, when these designs are examined, it becomes clear that designers are still restricted by the limited capabilities of 3D EM tools based on CSG. A literature review shows that all designs using 3D printing technology for RF & microwave components have followed the CSG approach [5]–[9]. The development of new CAD tools is thus essential if the full potential of AM in microwave technologies is to be unleashed. The present work can be considered a step towards achieving this goal; to enable the design of microwave components with truly arbitrary shapes, our proposed solution is to empower CSG-based 3D solid modeling tools with a free-form shape deformation technique.

Geometry deformation can be treated as an additional step in the design process. It can be performed on objects created in CAD tools with CSG. The roots of shape deformation techniques lie in computer graphics applications, where they are often used for object manipulation. One of the first deformation techniques was free-form deformation (FFD), proposed by Sederberg and Parry in 1986 [10]. Several other methods were developed later, including FFD based on nonuniform rational B-splines (NURBS) [11], [12]. Recently, shape deformation has been introduced into computational simulations of fluid dynamics [13]–[15] and low-frequency electromagnetics [16]. Some attempts to shape optimization using, e.g., level-set and topology gradient methods and Bézier curve parametrization

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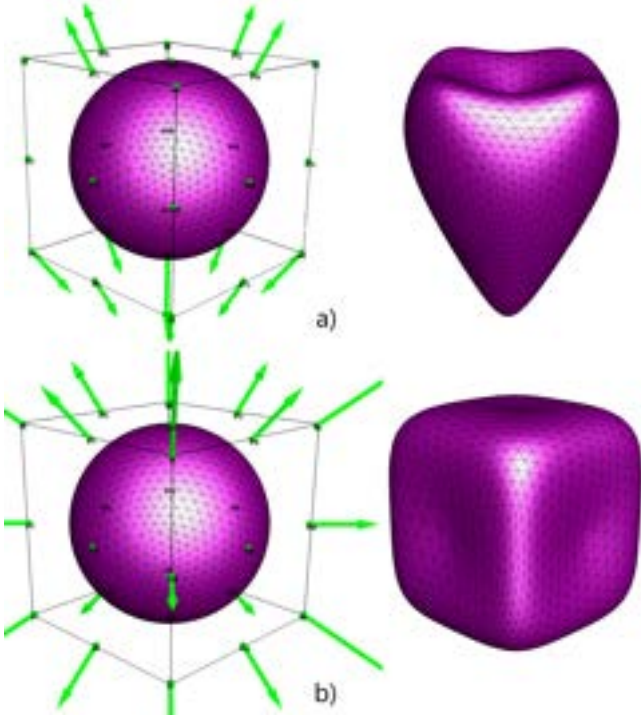


Fig. 1. Two examples of different shape deformations performed on the same spherical object. Left: the initial sphere; right: the result of the deforming operation. Control points and their displacements are represented by green points and arrows.

have been reported for the design of microwave components, [17]–[20] but as far as we know, no attempts have yet been made to use shape deformation techniques to model complex 3D microwave structures.

The basic principle of FFD is to define a set of points around the object that will control the deforming operation. The deformation is then begun by shifting some of these control points to certain new coordinates, and a vector displacement field is constructed using these displacements. The original structure is then mapped according to this smooth displacement function. The deforming operation can be performed on 3D objects represented by a surface or volume mesh. In this work, we consider a tetrahedral mesh used in 3D FEM. Two examples of different geometry deformation operations, shown in Fig. 1a and Fig. 1b, illustrate how the shape deformation method can be applied to 3D objects. In these two examples, the starting structure is a sphere on which we perform deformation. In both cases, a set of 26 control points is defined around the object. Then, the points are either kept fixed or moved to new coordinates, which results in the creation of complex geometries shown on the right. It can be seen that, by starting from a basic shape and simply moving the control points in different directions, various complex geometries can be obtained. The continuity of the object's surface (topology) and its smoothness are also preserved.

In this paper, shape deformation is integrated into the design process of high-frequency electronics. Mesh deformations are coupled with the 3D FEM electromagnetic field simulator InventSim [21], allowing for the semiautomated design of

various microwave components. The deformation operation is executed as an additional step in the EM simulation process, directly after the mesh of the analyzed structure is generated. The aim of this work is to show that this approach can remove current design limitations, and thus improve control over the electrical parameters of structures, such as unloaded quality factor and higher-order mode separation of cavity resonators and filters. The improvement of these parameters is highly desirable, e.g., in microwave filters for space applications [1].

In this paper, the idea presented in [22] has been further developed and extended. Shape deformation with radial basis functions is described in detail in Section II, along with several illustrative examples. A general framework for design optimization using these deformation techniques is presented in Section III. An investigation of cavity resonators exploiting geometry deformations for the Q-factor and spurious-free performance is given in Section IV. An extension of this concept is provided in Section V, featuring the design of deformed waveguide cavity filters. The prototypes of the third-order, sixth-order and 90°-twisted bandpass filters are fabricated using AM, and the measurement results are compared with simulations. Finally, the concluding remarks are given in Section VI.

II. SHAPE DEFORMATION WITH RBFs

Shape deformation is implemented here using interpolation with radial basis functions (RBF), as reported in [13], [23]–[25]. In general, a RBF is a real-valued function φ whose value is only determined by the distance from the origin, so that $\varphi(\mathbf{x}) = \hat{\varphi}(\|\mathbf{x}\|)$. The principle of the reshaping operation is represented in general by a mapping function $\mathcal{D}$:

$$\begin{aligned} \mathcal{D}(\mathbf{x}) : \mathbb{R}^3 &\rightarrow \mathbb{R}^3 \\ \mathbf{x} &\mapsto \mathcal{D}(\mathbf{x}) = \mathbf{x} + \mathbf{d}(\mathbf{x}) \end{aligned} \quad (1)$$

In this notation, $\mathbf{d}(\mathbf{x})$ can be interpreted as the deformation field applied to the 3D model. In the case of RBF-based deformation, this takes the following form:

$$\mathbf{d}(\mathbf{x}) = \sum_{i=1}^N \gamma_i \varphi(\|\mathbf{x} - \mathbf{x}_i\|) + p(\mathbf{x}) \quad (2)$$

where N is the number of control points defining the deformation, γ_i are the weights corresponding to the radial basis functions φ defined at the positions of the control points $\mathbf{x}_i$ (in 3D $\mathbf{x}_i \in \mathbb{R}^3$, $\mathbf{x}_i = [x_i, y_i, z_i]$), and $p(\mathbf{x})$ is a low-order polynomial (optional).

To perform the deformation, the mapping operation $\mathcal{D}$ needs to be specified. The first step is to define a set of N control points and their displacements, which indicate the direction and value of the offset. Next, a certain type of RBF with its parameters (e.g., radius, exponent) is selected: the choice of basis function affects the manner in which the mesh points are interpolated to match the new positions of the control points. There is a variety of RBFs commonly used for multivariate function interpolation, such as:

- Gaussian spline: $\varphi(\|\mathbf{x}\|) = \exp(-(\frac{\|\mathbf{x}\|}{r})^2)$, (3)

- multi-quadratic biharmonic spline:

$$\varphi(\|\mathbf{x}\|) = \sqrt{\|\mathbf{x}\|^2 + r^2}, \quad (4)$$

- polyharmonic spline:

$$\varphi(\|\mathbf{x}\|) = \begin{cases} (\frac{\|\mathbf{x}\|}{r})^k & \text{if } k = 1, 3, 5, \dots \\ (\frac{\|\mathbf{x}\|}{r})^k \ln(\frac{\|\mathbf{x}\|}{r}) & \text{if } k = 2, 4, 6, \dots \end{cases}, \quad (5)$$

where r and k are the radius and power of the RBF, respectively. A linear equation system is then solved to find the unknown weights γ_i given the coordinates of the original and displaced control points [24]. The matrix representation of this equation system is as follows:

$$\Phi\gamma = B \quad (6)$$

where γ is the weight matrix of dimension $N \times 3$, corresponding to N control points defined in a 3D coordinate system. Φ is an $N \times N$ matrix containing RBF values computed for the Euclidean distances between the corresponding control points, $\Phi_{i,j} = \varphi(\|\mathbf{x}_i - \mathbf{x}_j\|)$, and B is an $N \times 3$ right-hand side matrix with the new (displaced) coordinates of the control points. The system is dense, symmetric, positive definite, and its conditioning depends on selected RBF and control points distribution. Once the weights are computed, the deformation field $d(\mathbf{x})$ is constructed and can be applied to smoothly move all the mesh points of a 3D object. It can be seen that the mapping function can be easily defined with a limited number of control points, and then applied to create complex 3D objects. The computational cost of the deformation is also not significant, as the size of the linear problem to be solved is usually small. To illustrate this framework, we will take the simple example of the sphere presented in Fig. 1. In order to alter the geometry of this structure, a number of control points are selected around it. In this example, the control points are defined on a box enclosing the sphere. There are 26 control points in total: eight points at the vertices, six points at the center of each face, and twelve points at the midpoints of each edge. The displacement vectors are then defined for some of these points. Depending on the direction and number of control point offsets, the original shape is deformed to produce a different final geometry. It should be noted that the shape deformation scheme does not limit the number of control points. More complex deformation schemes can easily be employed by adding more control points with movement in arbitrary directions. Additionally, the definition of object deformation through displacement vectors of predefined control points is intuitive, and its result is easy to predict, as the deformation follows the displacements.

The technique can also be applied to bend or twist complex objects, as shown for the example of a waveguide cavity filter in Fig. 2. We can also define a series of deforming operations which can be executed one after another, such as squeezing, twisting, and bending, as shown in Fig. 2e.

III. GENERAL FRAMEWORK FOR DESIGN OPTIMIZATION WITH SHAPE DEFORMATION

Shape deformation can be easily integrated with design optimization, enriching the standard design parametrization

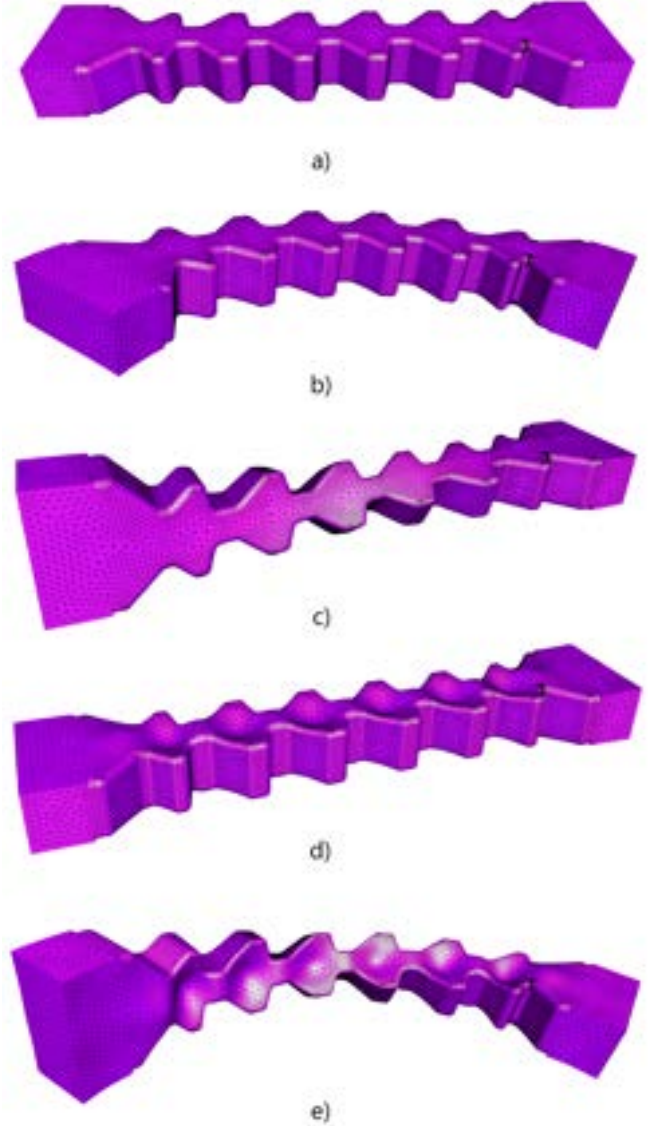


Fig. 2. Example of more complex shape deformations: a) original structure, b) bending, c) twisting 90° , d) squeezing resonant cavities vertically, e) squeezing + twisting + bending.

with a new set of design variables that define the deformation operation—namely the positions and movements of selected control points. The proposed design procedure is shown in the block diagram in Fig. 3. The optimization loop involves managing two sets of design variables: geometrical parameters altering the basic shape of the primal structure that is constructed using CSG, and deformation parameters controlling the deforming operation (position and displacements of the control points). Shape deformation is performed on the volume mesh generated for the primal structure. The deformed structure is then evaluated by an EM simulator to calculate its response in the frequency domain (S-parameters) or the frequencies and Q-factors of the resonating modes with an eigenmode solver. Optimization of the structure continues until

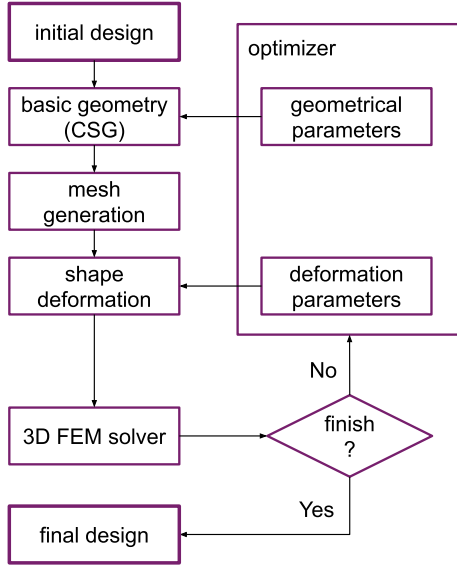


Fig. 3. A flowchart representing the general design optimization workflow.

termination conditions are met (e.g., error value below a preset threshold or maximum number of iterations reached). The final mesh of the deformed structure can be converted to a 3D model in STL file format and further processed in 3D geometry modeling software—for example, to prepare production files for additive manufacturing.

IV. OPTIMIZATION OF CAVITY RESONATORS USING SHAPE DEFORMATIONS

The deformation scheme was used to optimize cavity resonators with enhanced electromagnetic performance. The goal was to investigate whether shape deformation can improve the selected electrical parameters. In particular, we focused on maximizing the quality factor and higher-order mode separation. Let us consider the simple example of a rectangular cavity resonator with rounded edges presented in Fig. 4a. In order to alter the geometry of the structure, a set of control points is selected. In this example, a 3D grid of points is again defined, consisting of a total of 26 control points, as shown in Fig. 4b. The non-zero displacement vectors are then defined for some of these points. Here, two offset vectors are specified at the center of the cavity's top and bottom planes, pointing into the cavity. Given the list of control points and displacement vectors, the deformation field can be evaluated and applied to the volume mesh, resulting in a smoothly deformed geometry, also shown in Fig. 4b. The deformed mesh is then used by a 3D FEM electromagnetic field solver [21] to perform eigenmode analysis in order to calculate the frequencies and quality factors of the modes excited in the structure.

A. Investigation of Q vs. Spurious-Free Performance

In this experiment, the goal was to obtain the highest possible separation (Δf) of the first higher-order mode resonating at f_1 from the fundamental one at f_0 , while also achieving a given quality factor Q and preserving the resonant frequency

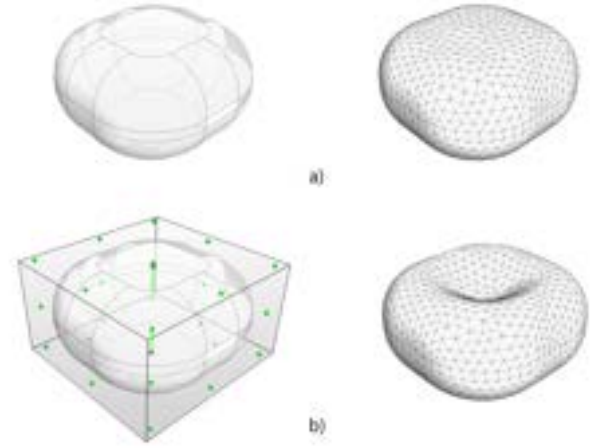


Fig. 4. Example of shape deformation: a) basic rectangular cavity with rounded edges and its mesh, b) definition of a squeezing operation and resulting mesh. Green arrows show the direction of the deformation, and green dots stand for fixed points.

at f_0 . In the simulations, we assumed finite conductivity of the cavity corresponding to that of aluminum ($\sigma = 3.8 \cdot 10^7$ S/m) and a resonant frequency equal to $f_0 = 10$ GHz. The optimization loop involves managing two sets of design variables: geometrical parameters altering the basic shape of the primal structure (cavity height, length, and width), and deformation parameters controlling the deforming operation; in this study, only deformation along the Z-axis (squeezing in the vertical direction) was considered. A quasi-Newton gradient optimization was used in this resonant cavity design workflow.

The optimization was run several times for different values of Q and Δf , which resulted in several resonator geometries, all having the same resonant frequency $f_0 = 10$ GHz and various Q-factors and spurious mode separations. The results are summarized in Fig. 5 in the form of a trade-off chart, showing an achievable Q vs. mode separation $\Delta f = f_1 - f_0$. The result for a rectangular resonator used as reference in [26] with dimensions of $22.86 \times 10.16 \times 19.9$ mm is also

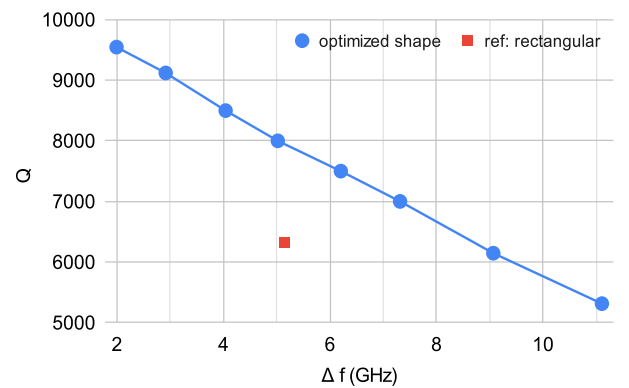


Fig. 5. Trade-off chart: resonator Q-factor vs. higher-order mode separation for $\sigma = 3.8 \cdot 10^7$ S/m.

TABLE I
EM-SIMULATED SPECIFICATIONS OF THE CAVITY RESONATORS

Project	f_0 (GHz)	f_1 (GHz)	f_1/f_0	Q_u
Hemispherical [26] (CSG)	10	19.8	1.98	7100
This work (shape deform.)	10	20.2	2.02	7100

reported. It can be observed that a greater higher-order mode separation can be achieved at the cost of lowering the Q-factor. Depending on the particular application, a resonator shape realizing the desired Q and Δf can be selected.

B. Cavity Resonators with Extended Spurious-Free Region

A second test was carried out to further demonstrate the possibilities of resonant cavity design based on user-defined shape deformation. For this reason, a recently reported high-Q cavity resonator with an extended spurious-free region was examined [26]. This design was a geometrically shaped hemispherical cavity prepared using the CSG approach by performing a number of Boolean operations and edge filleting. For comparison, a new deformed cavity resonator was designed using the shape deformation approach to outperform the simulation results achieved in [26]. An optimization procedure was run using the framework described in Section III to obtain the greatest possible mode separation while maintaining the Q-factor. The basic rectangular cavity with optimized dimensions of $18.77 \times 18.77 \times 12.72$ mm and filleted edges was squeezed vertically from top and bottom. The two central control points were moved towards the center of the cavity by 5 mm each. The final shape is the result of a RBF interpolation with a Gaussian spline of radius $r = 4$ mm, as in (3). A 3D model of the resulting structure is provided in Fig. 6. The comparison of Q and f_1/f_0 for both simulated models is summarized in Table I. The unloaded quality factor Q_u achieved by both designs was 7100 for metal conductivity $\sigma = 5.97 \cdot 10^7$ S/m (copper). The resonant frequency of the hemispherical cavity was 10 GHz, and the spurious-free region went up to 19.8 GHz. In this study, geometric deformation gave a greater spurious mode separation (up to 20.2 GHz), while preserving the Q-factor. The electrical performance achieved is slightly better than in the compared work, but a very basic model deformation was performed, rather than several geometrical operations.

V. DESIGN OF WAVEGUIDE CAVITY FILTERS USING SHAPE DEFORMATIONS

These two cavity-resonator examples have shown that shape deformation can be successfully applied to the CAD of simple microwave resonators, yielding enhanced electromagnetic performance. This approach should thus be used to design more complex components, such as waveguide cavity filters, in order to demonstrate its full potential in practical applications. We will consider three filter designs of Ku-band filters for satellite communication systems. All three designs are interfaced with a WR-75 rectangular waveguide standard. In the first two designs, we want to achieve an improvement in terms of out-of-band rejection and spurious-free performance by applying RBF-based shape deformation to the basic filter geometry. The

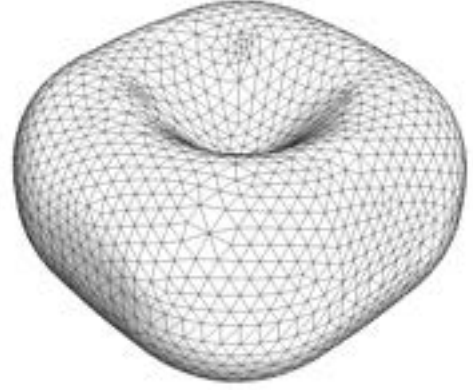


Fig. 6. Shape of the cavity resonator with $f_0 = 10$ GHz, $\Delta f = 10.2$ GHz and $Q = 7100$ (copper, $\sigma = 5.97 \cdot 10^7$ S/m).

goal of the third design is to combine a 90° waveguide twist and a filter in one compact device. The prototypes are to be additively manufactured as single pieces, and this fabrication technology imposes certain constraints on the allowable shape. This has to be taken into account very early on, even before the shape is deformed. It has to be noted that the Q-factor of the fabricated filter cavities will be limited due to the surface roughness and the conductivity that can be achieved in AM technology. In our designs, the Q and the actual metal conductivity obtained in the AM process will be estimated by investigation of the insertion loss of the fabricated structures. Note that, although microwave filters with better spurious-free performance have already been reported for other technologies (such as combline, ceramic, and coplanar waveguide filters [1], [27]–[31]), this work focuses on presenting the improvements achieved with a simple geometry deformation, allowing the final structure to be 3D-printed in one piece without internal supports.

A. Third-Order Bandpass Filter with a Complex Shape

The first structure shown is a third-order cavity bandpass filter (BPF) with inline topology. The reference structure and the starting point for our design is the conventional third-order BPF shown in Fig. 7a. This design is prepared to meet the requirements of Ku-band broadband input filters [32]. The main specifications to be met are a center frequency of $f_0 = 14.25$ GHz, bandwidth equal to 500 MHz (fractional bandwidth FBW = 3.5%), and return loss (RL) in the passband better than 20 dB. Even though shape deformation techniques allow shapes that are optimal in terms of performance to be obtained, several practical constraints have to be taken into account during the design process. As the goal is to enable 3D printing as a single manifold, the filter model must accommodate the constraints of the SLM process. Should the structure be tilted at an angle greater than 45° along the printing axis, extra supporting posts need to be placed inside the model. These supports then need to be removed later in a post-printing step, which could be a troublesome operation. A standard waveguide filter, shown in Fig. 7a, is not compliant

with the SLM process. For this reason, the primal filter model designed in this way is modified to meet the requirements of AM and to avoid the need to insert supporting posts inside the structure during printing; this is why the resonators have a quasi-triangular shape (see Fig. 7b).

Shape deformation was then used to squeeze the top and bottom of the resonator surfaces. In this case, the deformation depth of the cavities is 2.5 mm. As with the previous resonant cavities, shape deformation involves Gaussian spline RBFs (3). At this stage, the design procedure involved geometric optimization, which included the design parameters of the basic structure (length of the resonators, widths of the irises) and simultaneous shape deformation. The zero-pole method was used as the optimization technique [33], [34]. The entire process was performed in the InventSim EM simulator within the newly developed module allowing efficient design optimization with simultaneous shape deformation, including the optimization of control point positions and their movements. The final shape of the designed filter is shown in Fig. 7c.

The simulation results for the final third-order filter are presented in Fig. 8. The responses of the standard H-plane filter design from Fig. 7a and of the structure adjusted for 3D-printing from Fig. 7b are also plotted for comparison. The simulated insertion loss for both the deformed and the conventional H-plane filter equals 0.08 dB at f_0 , therefore the estimated Q for both models is around 4500 (for aluminum, $\sigma = 3.8 \cdot 10^7$ S/m). It can be observed that the cavity deformation resulted in an approximately 1.5 GHz wider spurious-free region than the reference design. Our experiments show that the deeper the squeeze of the cavity, the greater the spurious

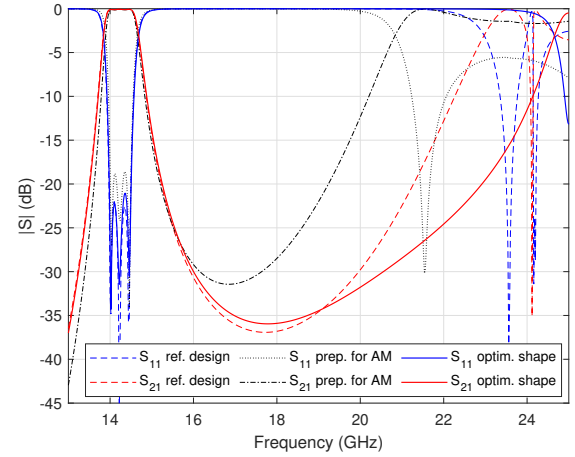


Fig. 8. Simulation comparison of the third-order BPF: standard reference design: dashed blue-red lines; filter prepared for 3D-printing: black dash-dotted lines; optimized design with deformed cavities: solid blue-red lines.

mode separation.

A prototype of the new filter design was manufactured as a single mechanical part using the SLM process [2], [35] with a TruPrint 1000 printer from Trumpf with an aluminum-silicon-based alloy AlSi10Mg, and then measured. In order to observe the wideband performance of the filter, waveguide adapters WR-51-to-WR-75 were used to reduce the influence of higher order modes excited in the WR-75 waveguide-to-coax adapters. The measurement setup using a vector network analyzer (VNA) and adapters is shown in Fig. 9. The simulations and measurements of the printed prototype are compared in Fig. 10. The measurements are in good agreement with the simulations. The insertion loss (IL) at center frequency f_0 is 0.2 dB, the RL is greater than 20 dB and the spurious-free region reaches up to 24.5 GHz. The response of the filter with no tuning screws is also provided. It should be noted that the initial measured response (without tuning) is very satisfactory: the IL at f_0 and spurious-free region match

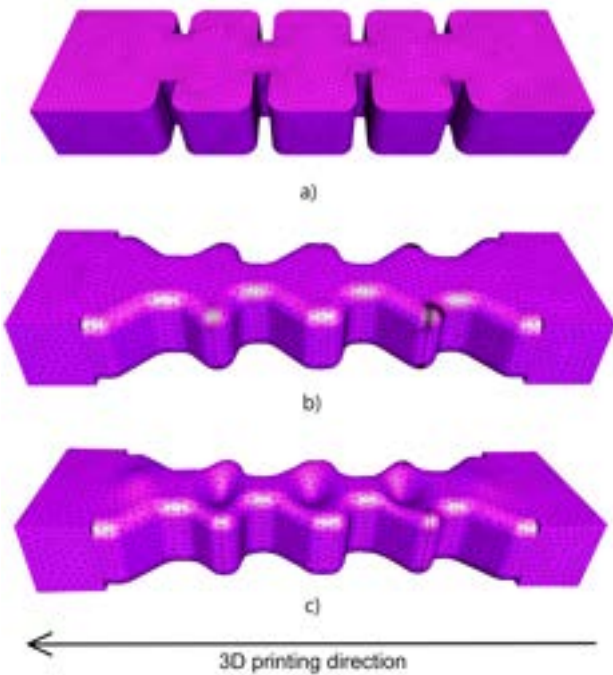


Fig. 7. Geometry of a third-order bandpass filter: a) standard structure, b) structure prepared for AM manufacturing, c) final structure with cavity deformation. The direction of 3D printing is also indicated.

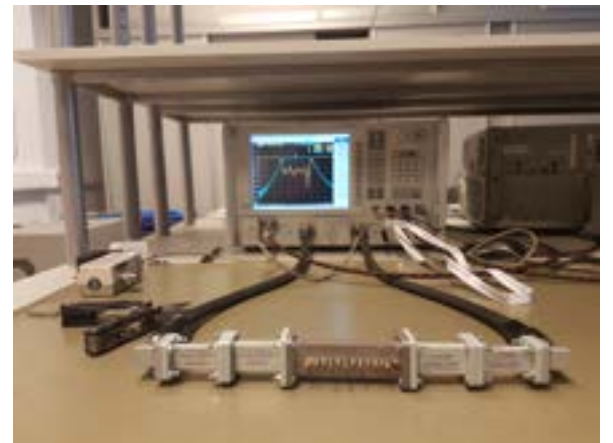


Fig. 9. Measurement setup using WR-51-to-WR-75 waveguide adapters. The sixth-order filter is measured here.

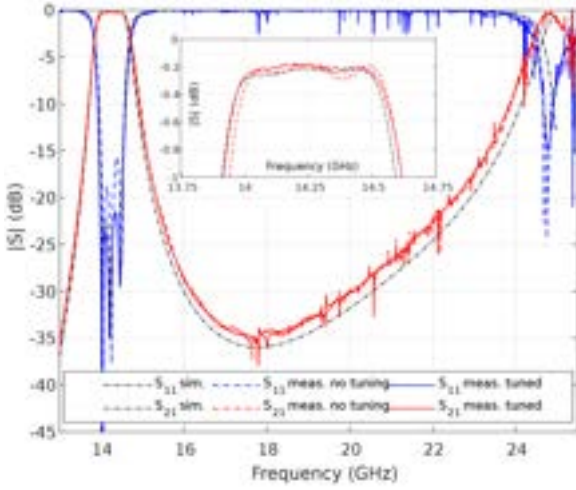


Fig. 10. Wideband comparison of simulations and measurement of the third-order deformed cavity BPF: simulation results: black dash-dotted lines; measurements with no tuning screws: dashed blue-red lines; final measurements (tuned): solid blue-red lines.

the simulations, and the RL in the passband is over 15 dB. Tuning screws were used to improve the in-band RL to the desired level of 20 dB. Based on the IL of the printed filter, we have estimated the effective conductivity of the 3D-printed aluminum powder, which according to these measurements is about $\sigma_{eff} = 4.2 \cdot 10^6$ S/m. This value is comparable to other reported aluminum structures fabricated with SLM—for example, it is slightly better than the effective conductivity achieved in [36]. This estimated conductivity σ_{eff} was later used in EM simulations of subsequent designs. The measurements were also used to determine the approximate Q-factor of the fabricated filter, which is equal to $Q = 1800$.

B. Sixth-Order Bandpass Filter with a Complex Shape

Upon completion of the third-order filter prototype, a sixth-order cavity BPF design was chosen for preparation. The specifications for the new design match those of the third-order filter: a center frequency of $f_0 = 14.25$ GHz, FBW = 3.5% and RL > 20 dB in the passband. Again, the basic filter geometry, shown in Fig. 11a, is modified to comply with the 3D printing process, resulting in the triangle-shaped model presented in Fig. 11b. To achieve greater spurious mode separation in the stopband, the deformation of the cavities is deeper and equals 3.75 mm. EM simulations of the sixth-order BPF are computed for the estimated metal conductivity $\sigma_{eff} = 4.2 \cdot 10^6$ S/m, equivalent to what was previously obtained with the third-order prototype. A comparison of the simulated wideband responses of the reference design and the new model is provided in Fig. 12. Here, the spurious-free region is up to ca. 27.8 GHz, and so equal to $1.95 f_0$ and approximately 4.5 GHz wider than in the traditional cavity BPF design. Note that the new filter model gives very high insertion loss in the stopband, which is another advantage of the deformation-based design. A prototype of the filter from Fig. 11b was fabricated using SLM technology the same way as the previous model and measured using the same setup shown in Fig. 9. The measurement results

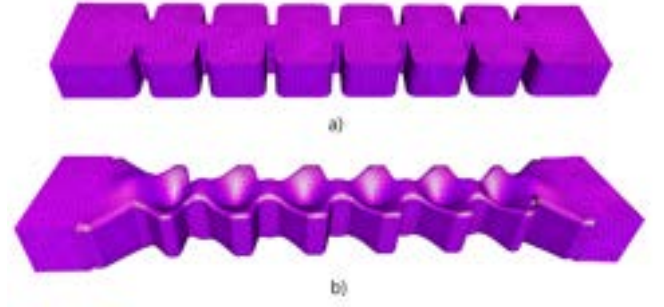


Fig. 11. Geometry of a sixth-order bandpass filter: a) standard structure, b) deformed structure.

are plotted in Fig. 13a (passband response) and Fig. 13b (wideband response). The passband comparison allows us to observe that the filter response is satisfactory and consistent with the simulations. The measurement taken in a wider band becomes unclear above about 26 GHz, due to the limited operating band of the waveguide adapters used. Nevertheless, from this measurement it can be discerned that the out-of-band characteristics match the response provided by EM simulation. The measured IL in the passband is 1.15 dB, which is slightly higher than the expected 0.9 dB received from the simulation with effective metal conductivity σ_{eff} . This is mostly caused by the addition of tuning screws which were not accounted for in the simulations. The estimated Q-factor of the filter equals $Q = 930$. One way of reducing the losses and receiving a higher Q-factor could be the metalization of all surfaces of the filter in a post-printing step. This could be either silver, copper, or gold plating, depending on the fabrication process. 3D-printed, copper, and gold plated waveguide components with an effective conductivity of copper $\sigma_{Cu} = 2.04 \cdot 10^7$ S/m and gold $\sigma_{Au} = 1.24 \cdot 10^7$ S/m are reported in [37]. For such conductor losses, the estimated Q-factor of the presented sixth-order BPF would be equal to $Q_{Cu} = 2550$ or $Q_{Au} = 1970$,

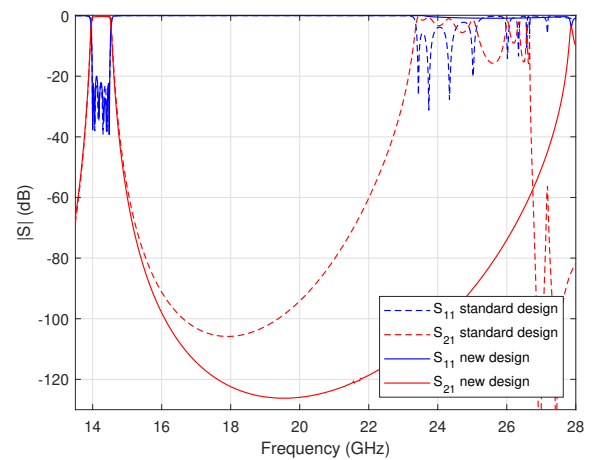
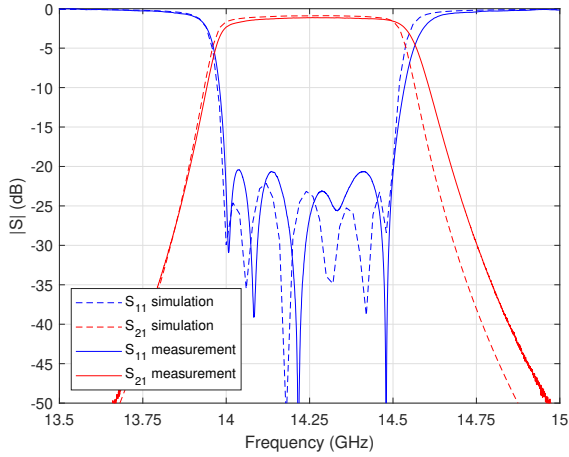
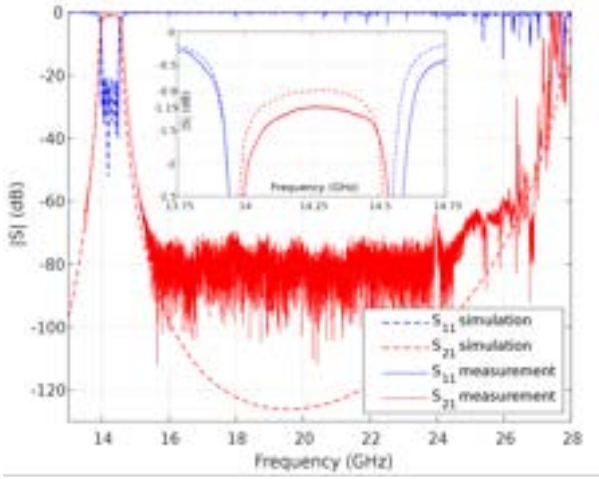


Fig. 12. Simulation comparison of the sixth-order BPF: standard design vs. deformed design.



(a)



(b)

Fig. 13. Comparison of simulations and measurement for the sixth-order deformed cavity BPF. (a) passband response. (b) wideband response.

respectively. Silver plating is presumed to bring similar, if not better, improvements, and should be considered in further research.

C. 90° Twisted Sixth-Order Waveguide Filter

To further demonstrate the possibilities introduced by shape deformation to EM CAD, we consider another design example. This design is a twisted Ku-band waveguide bandpass filter. The goal here is to realize a 90° waveguide twist with a filtering function combined in the same component. The efforts to reduce mass and size by merging waveguide filters, twists or bends have been recently reported in the literature [38]–[41]. Now, with the use of shape deformation, the design of such complex devices is significantly easier and can be realized by applying one or more deforming operations in the design process.

The specifications of the filter are a wide passband of 13 – 14.8 GHz (center frequency $f_0 = 13.9$ GHz, FBW = 13%) and an in-band RL over 20 dB. The design procedure was performed in several steps. The first step was to design

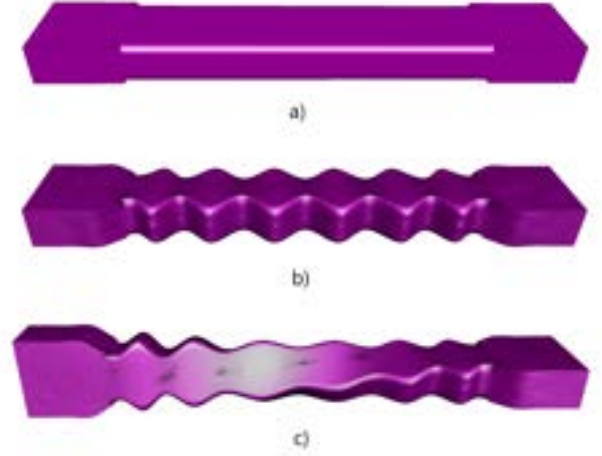


Fig. 14. Geometry of a 90-degree twisted bandpass filter: a) starting structure, b) straight filter structure, c) twisted filter structure.

an inline sixth-order bandpass filter by squeezing the sides of a straight WR-75 waveguide section to form inductive irises. The second step was to introduce another shape deformation, a 90-degree twist, as a second operation performed after squeezing. The twisting operation was performed by defining 24 control points: 12 points at the input section of the waveguide that were fixed, and the other 12 at the output section for which the non-zero displacements were defined. The new displaced coordinates were rotated by a given angle α . Then, the design was optimized again in several steps by gradually increasing α from 0 to 90 degrees. The starting structure and the consecutive design stages are shown in Fig. 14.

The component was fabricated using SLM with the same alloy AlSi10Mg and measured as is without tuning. The effective conductivity of the metal was estimated as $\sigma_{eff} = 5 \cdot 10^6$ S/m. The comparison of measurements and simulations along with the photograph of the component are provided in Fig. 15. Good agreement with the simulated model is achieved: the

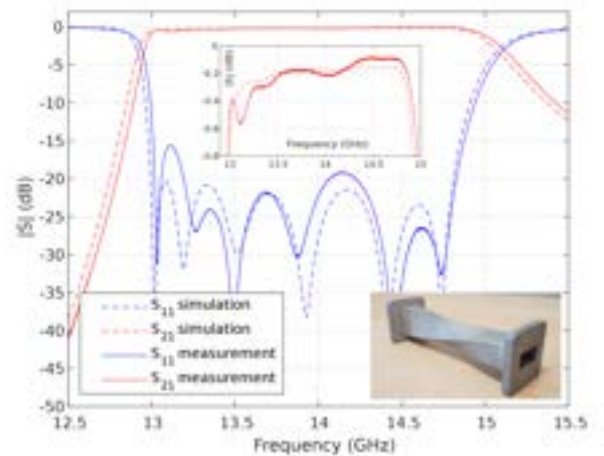


Fig. 15. Comparison of simulation and measurement results for the twisted waveguide filter. The inset shows a photograph of the fabricated component.

in-band RL is over 19 dB except for the mismatch at 13.1 GHz, where it equals 15 dB. The passband is shifted ca. 20 MHz above the specified band, which is only 1% of the bandwidth. These discrepancies could be corrected by tuning screws. The insertion loss is 0.19 dB at f_0 and varies from 0.1 to 0.4 dB over the passband, except for the mismatched region, where it is equal to 0.55 dB. The Q-factor estimated from the measured insertion loss value is equal to 1700. The manufactured twisted filter prototype validates the proposed application of the shape deformation technique in the design of these category of waveguide structures. The proposed optimization framework allows for a fast and effective design of microwave components with various complex geometries.

VI. CONCLUSION

This paper introduced and experimentally verified a shape deformation technique for the CAD of microwave components. This approach opens up new modeling possibilities to RF & microwave engineers by offering more degrees of freedom than the common CSG approach. The proposed deformation of 3D solids is easy to couple with known EM CAD tools and is intuitive thanks to the idea of altering the shape by moving arbitrarily chosen control points. First, the novel cavity resonators and filters with enhanced electromagnetic performance were designed using this deformation strategy and manufactured in a single mechanical part with 3D printing. These prototypes were used to verify the designs and to test the quality of AM with SLM technology for this category of high-frequency components. The measurement results confirmed the effectiveness of the new deformation-based design approach and demonstrated the possibilities offered by 3D printing. The deficiencies of this fabrication process, such as deteriorated metal conductivity, were also recognized. Future research directions can include the further investigation of complex microwave structures (filters, couplers, antennas) employing shape deformation, as well as enhancing the AM process by silver-plating the printed components.

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4.2. A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication Systems

The second paper contributing to this thesis, published in *IEEE Microwave and Wireless Components Letters*, reports on the application of the shape deformation technique in a very practical scenario, namely the design of channel filters for output multiplexers (OMUXs) in satellite communication systems [102]. In this letter, a novel circular waveguide dual-mode (CWDM) resonant cavity design with high Q and significantly enhanced spurious mode separation was proposed. The improvement in out-of-band performance was obtained by modifying the shape of the cavity using FFD. Based on the field analysis of the excited cavity modes, a deformation operation was selected that affects the next higher-order mode (here: TE_{211}), while having little influence on the operating one (TE_{113}). The newly developed cavity was subsequently employed to design a highly selective fourth-order Ku-band channel filter, intended to operate in a wideband OMUX (having up to 2 GHz included bandwidth). The proposed solution, due to its wide spurious-free band, is superior to other state-of-the-art high- Q channel filters, and would allow the realization of a wideband OMUX without the need of any additional lowpass filters. Several prototypes of the presented filter were additively manufactured in metal using the SLM process and measured, experimentally validating the design. Similar to [MB2], this investigation was conducted as part of the ESA-funded activity entitled "*RF filters with Enhanced Electromagnetic Performance for Space Applications Exploiting Shape Optimization with Free-Form Deformation and Additive Manufacturing Techniques*". The study is described in subsequent sections, with a particular focus on the design and mode analysis of a single resonant cavity. Additionally, supplementary material in Section 4.2.2 presents the initial test design of a second-order BPF based on a single modified CWDM cavity, developed during this investigation and not included in the referenced publication [MB3].

4.2.1. Resonant cavity analysis

As mentioned above, the goal of this work was to develop a novel solution aiming for operation in contiguous channel OMUX systems [103]. The main requirement for bandpass filters in such applications is to exhibit narrow bandwidth and high selectivity in order to accommodate multiple channels at adjacent frequencies in a single system. Moreover, such filters have to handle high power and have low transmission loss. Therefore, due to its compliance with the above-mentioned requirements, the most widely adopted and established choice for an OMUX channel BPF, at the time of publication, was the solution based on CWDM cavities [104–106]. Filters based on these cavities typically work with TE_{11n} modes, $n = 1-4$, thanks to their high attainable Q -factors. The dual-mode cavity configuration, i.e., operating with two degenerate modes excited in two orthogonal planes, enables the implementation of more selective filter responses with TZs. However, conventional CWDM filters suffer from poor out-of-band performance, especially when operating with higher-order TE modes such as TE_{113} or TE_{114} , because of their close proximity to

other modes excited in the cavity. This is a serious limitation which significantly constrains the design of OMUX systems with broad operational bandwidths. Therefore, new designs improving the spurious-free range have been sought after, among which the stepped circular waveguide dual-mode (SCWDM) filter [107] proved to be the most effective. The SCWDM cavity yields at least 30% wider spurious-free window than that of conventional CWDM resonator. Nonetheless, for broadband OMUX applications that result is still deemed insufficient, which is why a new solution improving over the state-of-the-art has been investigated.

The investigation begins with an analysis of the conventional circular waveguide (CW) resonant cavity, shown in Fig. 4.11a. The analysis of modes excited in a cylindrical cavity is a well-established topic in microwave engineering, thus the resonant frequencies of each mode can be obtained analytically from the following equations [108]:

$$f_{nml}^{TE} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{p'_{nm}}{R}\right)^2 + \left(\frac{l\pi}{H}\right)^2} \quad (4.1)$$

for the TE_{nml} modes and

$$f_{nml}^{TM} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{p_{nm}}{R}\right)^2 + \left(\frac{l\pi}{H}\right)^2} \quad (4.2)$$

for the TM_{nml} modes, where: R denotes the radius of the waveguide, H denotes the height (or length) of the cavity, c is the speed of an EM wave in free space, ϵ_r and μ_r are the relative permittivity and permeability of the material, respectively, and finally p_{nm} and p'_{nm} stand for the m -th root of the Bessel function J_n or its derivative, respectively. Using these equations, one might calculate the resonant frequencies for a desired CW cavity with dimensions R, H . The unloaded Q-factors, Q_u , can also be derived from [108]. As mentioned above, because of the requirements of Ku-band OMUX application, the designed CW cavity should support two orthogonal TE_{11n} modes, both of which will be employed in the filter design, hence the term circular waveguide dual-mode (CWDM). For the first Ku-band channel, the conventional CWDM cavity should operate with two TE_{113} modes resonating at $f_0 = 10.8$ GHz. The appearance of non-desired TE and TM modes at close proximity to the operating ones depends on the relation between the cavity dimensions, R and H . Using equations (4.1) and (4.2), the optimal dimensions targeting a wide upper stopband free of spurious modes can be selected: radius $R = 14.3$ mm, and length $H = 50.6$ mm. Such a cavity sustains the operation of two degenerate TE_{113} modes at $f_0 = 10.8$ GHz, and the first higher-order mode is TE_{212} , resonating at 11.8 GHz. Below f_0 , the nearest spurious mode is TE_{211} , excited at 10.6 GHz. The Q-factor of the operating mode, calculated for silver ($\sigma_{Ag} = 6.3 \cdot 10^7$ S/m), equals $Q_u = 20700$. Thus, such a CWDM cavity could yield a spurious-free window of approximately 1 GHz in the upper stopband; Naturally, these resonance frequencies may change when the resonator becomes externally loaded (i.e., coupled at input and output), which can effectively limit the spurious-free performance even further. The potential for improvement in this regard was identified in [107], where an alternative stepped CW structure was proposed. The SCWDM cavity, shown in Fig. 4.11b, allows for obtaining a clean stopband between 10.5 and 12

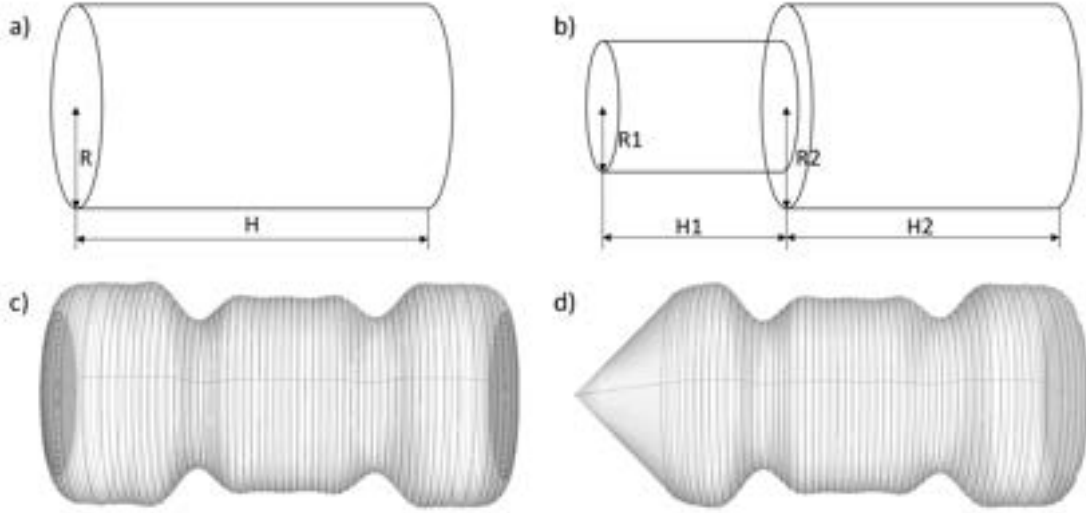


Fig. 4.11: Resonant cavities based on circular waveguide: a) standard CWDM cavity, b) stepped CWDM cavity (SCWDM) [107], c) deformed CWDM cavity, d) cavity modified for 3D printing.

GHz, with the two TE_{113} modes resonating at $f_0 = 10.73$ GHz. Similarly to the CWDM cavity, the spurious-free window in this solution is limited by the TE_{211} mode below f_0 and TE_{212} above. The Q_u of the SCWDM resonator, obtained from an eigenmode analysis in Ansys HFSS assuming silver coating, is equal to 19100. For a Ku-band OMUX, out-of-band performance of the SCWDM cavity supports a 17-channel configuration with 1 GHz of usable bandwidth. However, it remains inadequate for broader OMUX bandwidths—specifically, for accommodating the entire operational range from 10.7 to 12.7 GHz without resorting to additional lowpass filters.

Therefore, various modifications of the basic cylindrical cavity were studied to enhance the spurious-free performance while maintaining a similar Q . For all performed tests, the basic cavity model was modified by filleting its sharp edges and applying shape deformation based on RBF interpolation. The dimensions of the cylinder, R and H , were also altered to influence the modes excited in the structure and, as a result, to investigate the effect of various cavity deformations on different mode types. Ultimately, the H/R ratio selected for a closer study was equal to 4 ($R = 13.21$ mm, $H = 52.84$ mm). Such a choice of dimensions results in a different arrangement of modes excited in the cavity: When positioning the TE_{113} resonance at $f_0 = 10.8$ GHz, the first lower spurious mode is TM_{012} at $f_L = 10.375$ GHz, whereas the first higher-order mode is TE_{211} , appearing at $f_H = 11.39$ GHz. The EM field distribution of each mode was closely examined and based on that investigation, a geometry deformation was found—transversal squeezing at positions $\frac{1}{3}H$ and $\frac{2}{3}H$ —that would move away the upper spurious TE_{211} mode while not influencing the operating TE_{113} mode. This effect can be observed in the mode chart in Fig. 4.12, which illustrates the relation between the deformation depth, applied at positions $\frac{1}{3}H$ and $\frac{2}{3}H$, and the resonance frequencies of each of the modes in question. Such an operation was then used to optimize the cavity shape, solving for an extended separation between modes TE_{113} and TE_{211} while maintaining high Q . The optimization procedure involved altering the geometric variables R, H as well as deformation parameters (control point positions and compression depth) and was

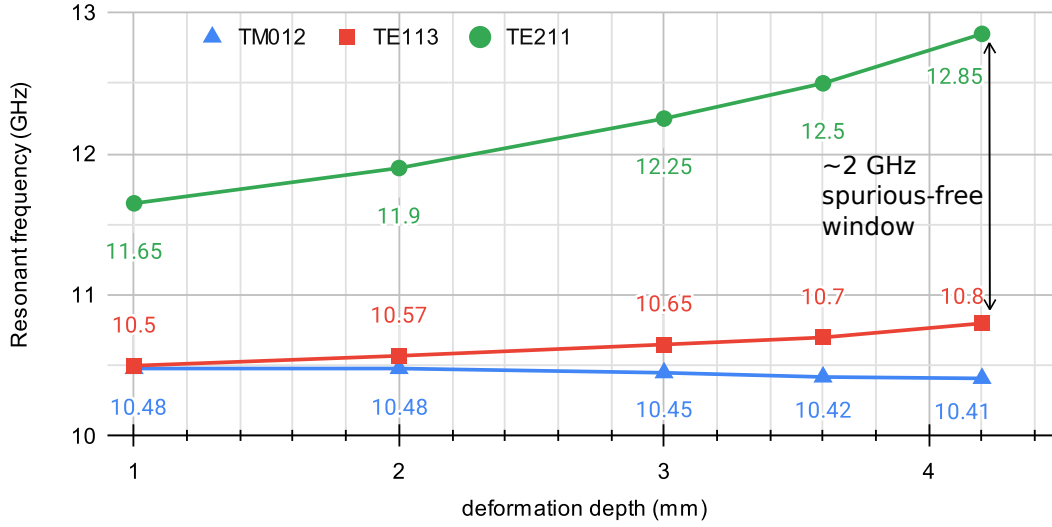


Fig. 4.12: Mode chart of the deformed CWDM cavity resonator illustrating the effect of shape deformation on the modes closest to TE_{113} , for a fixed H/R ratio equal to 4.

performed utilizing an eigensolver to calculate the resonances, f_0 , f_L , f_H , and their corresponding Q-factors. As a result, a new modified cavity was found, shown in Fig. 4.11c, for which the TE_{113} resonance appears at $f_0 = 10.85$ GHz, but the first unwanted mode, TE_{211} , was successfully moved up to $f_H = 12.96$ GHz, creating a wide spurious-free window of around 2 GHz. The first spurious mode below the desired operating band, TM_{012} , remained near f_0 , at $f_L = 10.58$ GHz. The optimized cavity is a smooth-profiled cylinder of dimensions $R = 13.92$ mm, $H = 55.68$ mm, undergoing shape deformation (transversal squeezing) by $d_1 = 4.13$ mm and $d_2 = 4.57$ mm at precisely $H_1 = (\frac{1}{3}H - 3.05)$ mm and $H_2 = (\frac{2}{3}H + 1.39)$ mm, respectively. Subsequently, the deformed CWDM cavity from Fig. 4.11c needed to be adapted for fabrication by means of single-piece metal 3D printing: it was cut at a 45° angle at one end to avoid overhanging surfaces, resulting in a conically tapered model shown in Fig. 4.11d. With respect to the optimized one, the tapered cavity is slightly longer ($H = 61.85$ mm) due to the tip of the conical taper. A summary of the resonances and Q-factors for all the discussed cylindrical dual-mode cavities is provided in Table 4.1. It can be concluded that, by deforming the cavity shape in the described manner, a significant improvement in out-of-band performance was made possible, with only a minor reduction in the cavity's Q . Further modification for AM compliance, however necessary, resulted in a slight degradation in the obtained spurious-free region and Q . Nevertheless, the final AM-compliant cavity also demonstrates superior performance relative to the state-of-the-art SCWDM resonator, it was therefore employed in bandpass filter design.

4.2.2. Initial verification: second-order filter design

After completing a rigorous mode analysis, the proposed deformed CWDM cavity was put to the test in a filter design. The structure selected for testing was a second-order BPF, composed

Tab. 4.1: Analysis of cylindrical cavity resonators

cavity type	standard CWDM	SCWDM [107]	deformed CWDM	AM-compliant CWDM
f_0 (TE ₁₁₃) (GHz)	10.8	10.73	10.85	10.9
f_L (GHz)	10.6	10.5	10.58	10.5
f_H (GHz)	11.8	12.0	12.96	12.77
Q_u (silver)*	20700	19100	18000	17800
spurious-free region (GHz)	1.2	1.5	2.38	2.27
upper stopband	$1.09 f_0$	$1.12 f_0$	$1.19 f_0$	$1.17 f_0$

* The Q_u was calculated analytically for standard CWDM [108] and by means of eigenmode analysis otherwise. The assumed electrical conductivity of silver was $\sigma_{Ag} = 6.3 \cdot 10^7$ S/m.

of a single dual-mode cavity as the one shown in Fig. 4.11d. The filter, shown in Fig. 4.13d, was designed for the first channel of a wideband Ku-band OMUX, i.e., to have a passband located at $f_0 = 10.8$ GHz, bandwidth $BW = 72$ MHz, and $RL \geq 21$ dB. The main purpose of the test filter was to check the obtained stopband performance when the cavity is loaded and penetrated with tuning screws. In this design, the cavity operates with two degenerate TE₁₁₃ modes, which are inter-coupled by two coupling screws, inserted at 45° with respect to input and output planes. Apart from those, there are four additional screws for tuning the resonant frequencies of the two operating modes. The resonator is connected to the input and output WR-75 waveguides through thin irises, and the output is rotated by 90° relative to the input. The ideal filter characteristic, synthesized coupling matrix and coupling-routing diagram are provided in Fig. 4.13a–c. The design was optimized by adjusting the resonances and couplings: either by modifying the irises, or by controlling the screws. Again, the zero-pole method was used for final optimization [90]. Then,

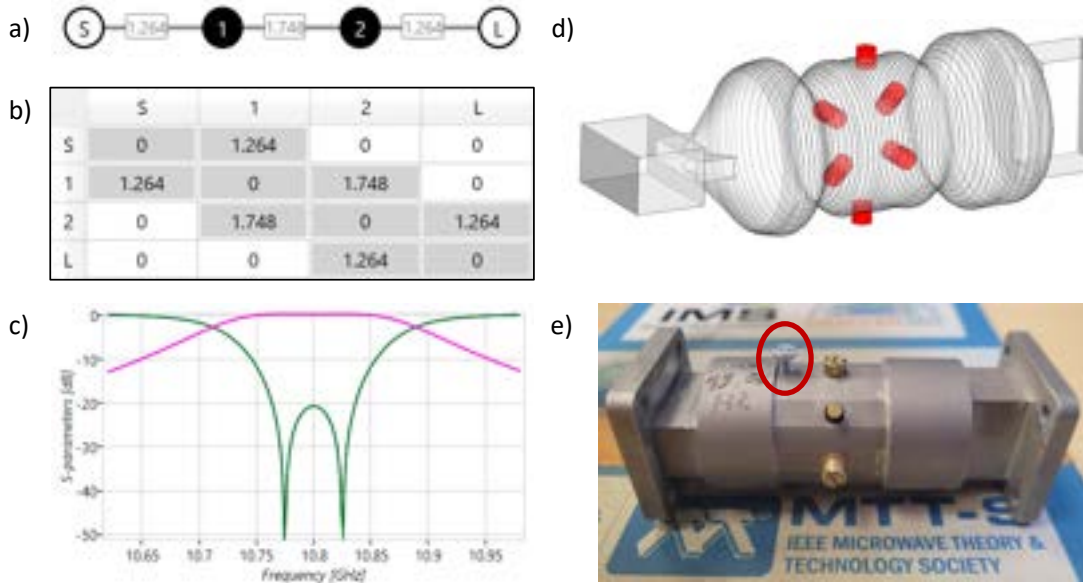


Fig. 4.13: Design of the second-order test filter: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) EM model of the AM-compliant BPF with tuning screws, e) photograph of the manufactured prototype.

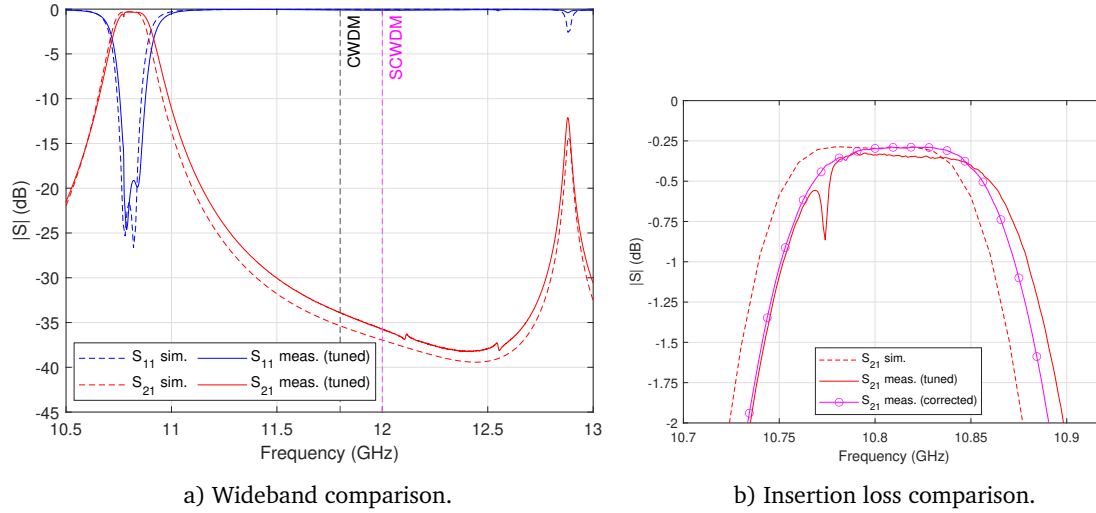


Fig. 4.14: Response of the second-order dual-mode BPF. EM simulation: dashed lines; measurement (tuned): solid lines; measurement after removing the TM_{012} spurious mode: solid-circled line. In (a), the spurious-free region limits for the standard CWDM and SCWDM filters are provided for reference. The EM simulation considers a lossy metal with $\sigma_{\text{eff}} = 4.2 \cdot 10^6$ S/m.

the filter was fabricated monolithically in aluminum by SLM using a TruPrint 1000 printer [100]. The S-parameter response of the final design is compared with the measurement result in Fig. 4.14. It can be concluded that, in terms of spurious-free performance, the BPF grants the desired wide upper stopband enabled by the new cavity. However, when analyzing the measured response, one significant discrepancy was observed: when inspecting the passband response, a point of sharp drop of transmission was noticed, as shown in Fig 4.14b. That spurious resonance was caused by the excitation of the TM_{012} mode, which had unintentionally been moved up in frequency and into the passband by the insertion of tuning/coupling screws. This phenomenon had not been spotted at the design stage, because primarily, ideal symmetry of the screws was assumed. In order to remove that mode from the passband, a pair of additional tuning screws had to be inserted—these are also visible in the photograph in Fig. 4.13e. The location of the corrective screws was chosen in an eigenmode simulation so as to influence only the unwanted, spurious resonance, without affecting the operating TE_{113} mode. The corrected passband response is also plotted in Fig 4.14b. The measured in-band IL is close to 0.3 dB, which corresponds well with the lossy simulation result for a metal with a conductivity of $\sigma_{\text{eff}} = 4.2 \cdot 10^6$ S/m. Thanks to the lessons learned during this trial, this undesirable scenario has been effectively ruled out in the subsequent filter design: asymmetrical insertion of screws was always assumed in the design from this point on, and special attention was paid during design optimization to ensure that the TM_{012} mode is kept away from the passband.

4.2.3. Main design: fourth-order filter

The test filter has proven that the use of the newly designed cavity can indeed bring a significant improvement in clearing out the rejection band. The new cavity was therefore used to

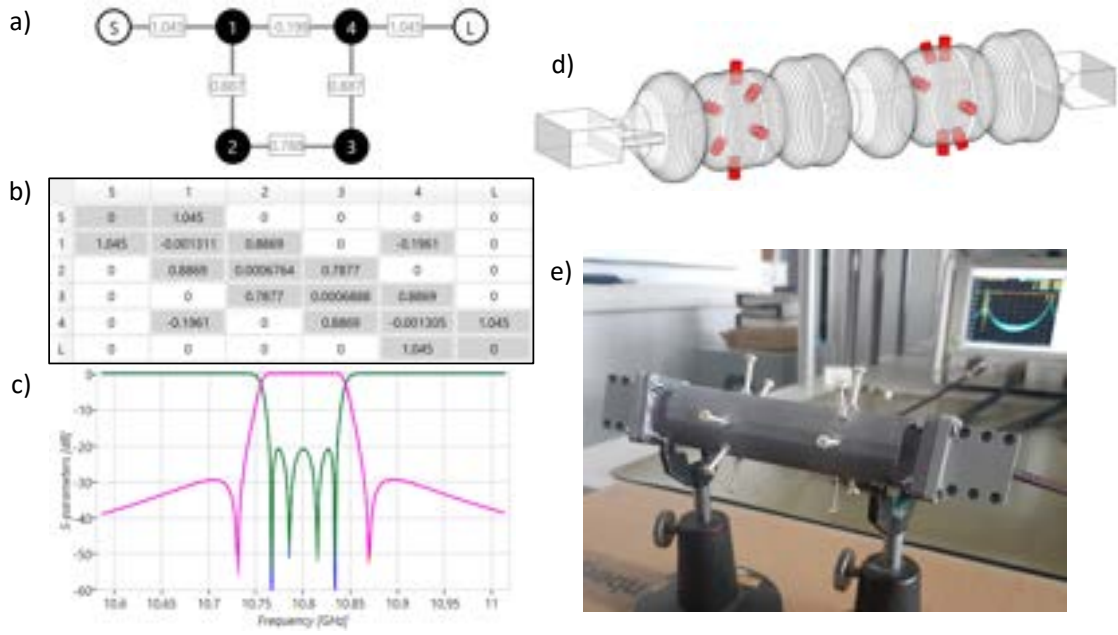


Fig. 4.15: Design of the fourth-order OMUX channel filter: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) EM model of the AM-compliant BPF with tuning screws, e) photograph of the manufactured prototype.

design a fourth-order BPF with two TZs. Just like the previous one, this filter is intended for the lowest channel of a Ku-band OMUX accommodating a wide bandwidth of 10.7–12.7 GHz. The main requirement for this design is therefore to offer a stopband covering that entire OMUX band. Other specifications for that channel filter are the following: $f_0 = 10.8$ GHz, $BW = 72$ MHz, and $RL \geq 21$ dB. The fourth-order filter utilizes two dual-mode cavities, with operating modes arranged in a quadruplet configuration (see coupling-routing diagram in Fig. 4.15a); it therefore produces two TZs. They were placed at 10.73 and 10.87 GHz, respectively—symmetrically about f_0 . The chosen filtering characteristic and coupling matrix are provided in Fig. 4.15. The two deformed CWDM cavities are interconnected via a cross-shaped iris, and interfaced at input and output with WR-75 rectangular waveguide. The internal cavity coupling between the two degenerate TE_{113} modes is once more achieved by means of coupling screws. Together with tuning screws for adjusting the resonant frequencies, the design uses twelve standard metric M3 screws, six per cavity. The filter was optimized using the zero-pole optimization by adjusting the coupling and tuning screws and the widths and heights of the irises (16 variables in total). The deformed DM cavities were pre-designed beforehand and not modified in the final optimization; instead, the resonances were controlled by the tuning screws. In the optimization process, the locations of the filter's reflection and transmission zeros were shifted to match those of the ideal filtering characteristic from Fig. 4.15c. The complete filter model after optimization can be examined in Fig. 4.15d. The dimensions of the cavities are the same as given in Section 4.2.1, i.e., $R = 13.92$ mm, $H = 61.85$ mm, and the total length of the fourth-order BPF is equal to $L = 126.77$ mm (including all the irises, but excluding I/O waveguide sections).

4.2.4. Experimental verification and discussion

The designed OMUX filter was manufactured utilizing the same technological process (SLM) and aluminum alloy (AlSi10Mg) as the second-order test unit. In this case, the TruPrint 2000 printer from Trumpf [100] was used because of its larger build area, which enabled the production of three filter prototypes in a single run. All three units were manually tuned with silver-plated M3 screws and measured, using the measurement setup as shown in Fig. 4.15e. A comprehensive comparison of the measured components with EM simulation is provided in Fig. 4.16. The comparison demonstrates that the measurements of all the examined prototypes are in excellent agreement with the simulation. Most importantly, the upper stopband behavior observed in Fig. 4.16a extends beyond 12.7 GHz ($1.17 f_0$), thus fully covering the wide operating bandwidth of a Ku-band OMUX. Furthermore, the passband analysis from Fig. 4.16b reveals that although the achieved bandwidth is 2 MHz narrower than specified, the fabricated components otherwise met the imposed specifications; the in-band RL is greater than 21 dB and the TZs are located at 10.734 GHz and 10.869 GHz, respectively. The observed insertion loss measured at f_0 varies between 0.96 and 1.03 dB, depending on the component (see Fig. 4.16c). Based on the measured IL, the estimated Q of the filter is equal to 2500. It is evident that such a value is below what is desired for such applications and that it differs significantly from the theoretical unloaded Q for silver (14500), or the Q -factor achieved by the reference SCWDM design made from CNC-machined aluminum [107] (12800). The main reason for this divergence lies in the surface roughness and effective metal conductivity of the 3D-printed aluminum alloy, which was estimated to be equal to $\sigma_{\text{eff}} = 1.72 \cdot 10^6$ S/m. This value is slightly lower than the one achieved with previous designs reported so far ($4.2 \cdot 10^6$ S/m), it is nevertheless comparable to other, similar components reported in the literature, also 3D-printed from aluminum powder [56, 109]. In order to improve the filter's Q and therefore reduce the insertion loss, several actions might be undertaken. Firstly, some post-processing steps such as surface polishing or silver plating could be exercised. However, this is not a trivial task because of the limited access to the inside of the filter, and it might require fabrication and assembly in several parts. Secondly, acknowledging the continuous development of high-precision AM, some newer, more precise 3D-printing services could be employed [21, 101, 110], that take into account the specific needs of microwave electronics. For example, the work in [21] presents a fabrication process based on SLM, involving post-printing treatment of the surfaces, which achieves a conductivity of $2.5 \cdot 10^7$ S/m. For the proposed design, this would translate into a Q -factor of approximately 9000, making it more competitive with the state-of-the-art. Alternatively, assuming the use of the same process based on powder bed fusion [101] that was later employed for other designs presented in the remainder of this work, a conductivity in the order of $1.0 \cdot 10^7$ S/m could be expected, reducing the IL to around 0.4 dB. Consequently, with appropriate post-processing and specialized procedures, the gap between AM and traditional fabrication processes such as CNC can be effectively minimized, allowing the realization of high-end electronic hardware such as the designed channel filter. Moreover, the obtained results

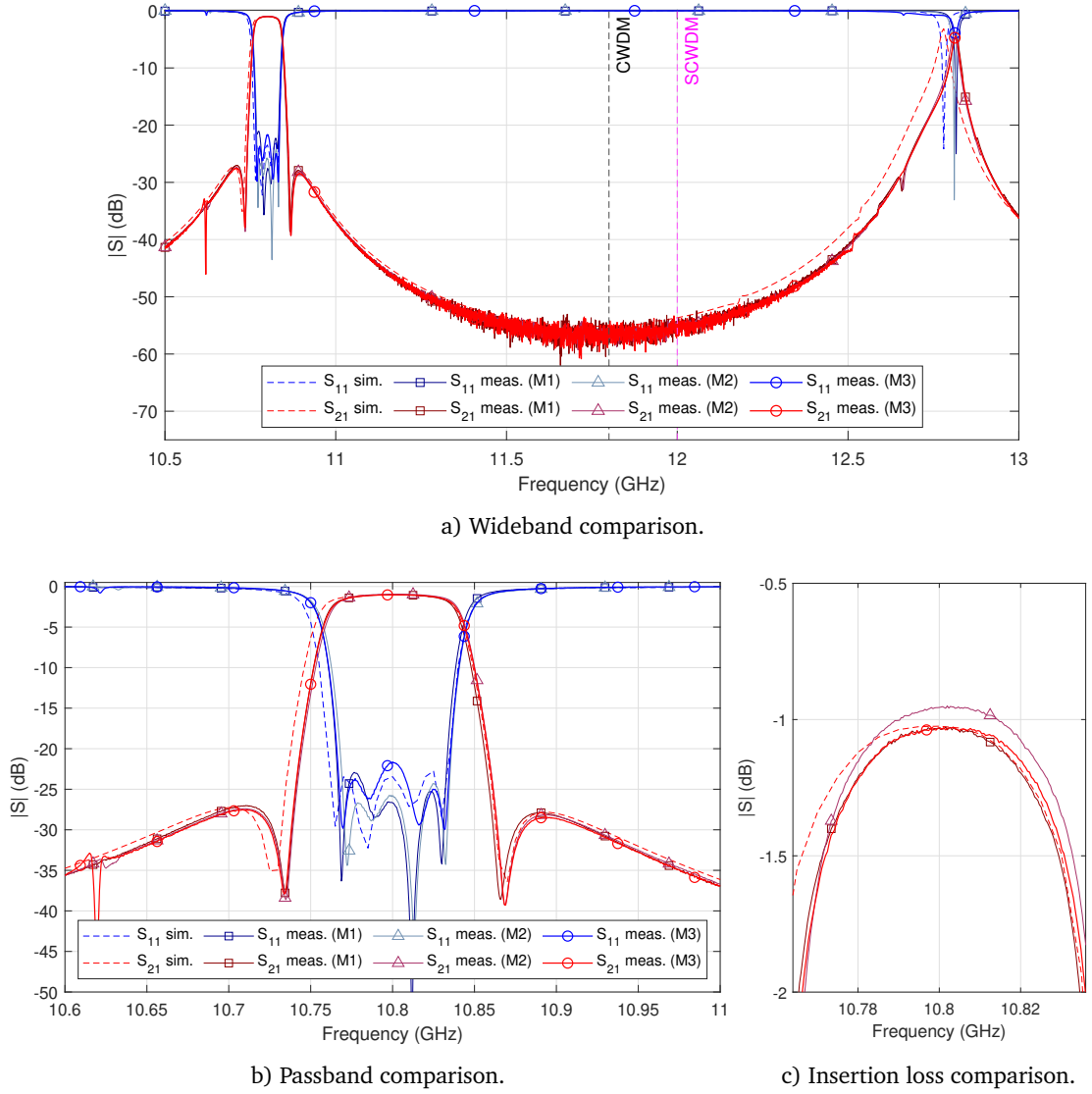


Fig. 4.16: A comparison of simulation and measurement of the fourth-order deformed CWDM BPF. EM simulation: dashed lines; measurement (tuned, three models M1–M3): solid lines. In (a), the spurious-free region limits for the standard CWDM and SCWDM filters are provided for reference. The EM simulation considers a lossy metal with $\sigma_{\text{eff}} = 1.72 \cdot 10^6 \text{ S/m}$.

show that simultaneous 3D printing produced highly consistent filtering performance across the three test units, confirming that AM can be effectively scaled yielding reliable outcomes.

In conclusion, implementation of the proposed design in the lowest channel of a wideband manifold-coupled OMUX would allow its realization without the necessity of using additional low-pass filters. The reported solution, thanks to the application of the new deformed dual-mode cavities, improved the spurious-free performance by at least 35% with respect to SCWDM filter and by nearly 90% when compared to the conventional CWDM design. Notably, the same approach could also be applied to design other channels of the same OMUX, including the highest channel BPF, located near 12.6 GHz—with a spurious-free window covering the passband of the first channel.

[MB3] A Circular Waveguide Dual-Mode Filter with Improved Out-of-Band Performance for Satellite Communication Systems

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Author's contribution:

Analysis and design of the resonant cavities modified using shape deformation techniques, design of the dual-mode filter presented in the paper, preparation of the 3D models for additive manufacturing, measurements and manual tuning of the prototypes.

A Circular Waveguide Dual-Mode Filter with Improved Out-of-Band Performance for Satellite Communication Systems

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Abstract—This letter presents a novel design for a 3D-printed circular waveguide dual-mode (CWDM) filter with a modified cavity shape. The modification leads to a wide spurious-free stopband, which is highly desirable for channel separation in waveguide contiguous output multiplexers (OMUXs) in satellite communication systems. The new resonant cavity design is a result of applying shape deformation to a basic circular cavity in order to move away and suppress parasitic modes. A fourth-order Ku-band channel filter with two transmission zeros is designed, fabricated by additive manufacturing (AM) in one piece and measured. In comparison with the state-of-the-art design of a stepped CWDM filter, an improvement of approximately 35% wider spurious-free range is achieved.

Index Terms—dual-mode waveguide filters, output multiplexer (OMUX), shape deformation, additive manufacturing.

I. INTRODUCTION

In modern satellite communication systems, there is an ongoing need to increase the capacity of the provided services. To meet this demand, a larger number of transponders need to be allocated, which raises the requirements for satellite payload systems. This mostly concerns the design of contiguous channel output multiplexers (OMUXs) [1], [2], which are among the most specialized microwave passive components comprising the satellite payload. The key requirement for OMUX systems is to accommodate multiple channels and exhibit a high power-handling capability and low transmission losses. Therefore, channel filters for OMUX applications should be narrowband, highly selective, as well as maintain a high Q-factor and high power handling. For this reason, the most preferred choice for an OMUX channel filter is the circular waveguide dual-mode (CWDM) cavity filter [3], [4], which has been subject to many modifications and improvements over the years [5]–[7]. This filter type typically

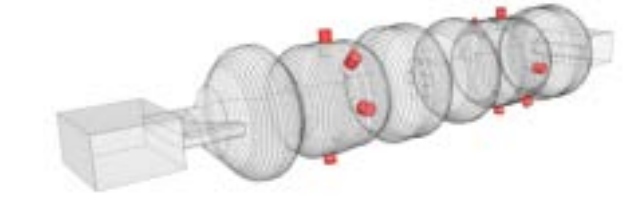


Fig. 1. Geometry of the proposed deformed circular waveguide dual-mode filter.

operates with TE_{11n} modes, $n = 1-4$, which offer high Q-factors, and the dual-mode cavity arrangement allows the realization of advanced filter response with transmission zeros (TZs). However, the conventional CWDM filter has a poor out-of-band performance, which is a major limiting factor in the design of OMUXs with many channels and wide working bandwidths.

In order to accommodate more channels in a single multiplexer manifold and eliminate additional lowpass filters from the OMUX, one of the most critical challenges is to design channel filters which do not interfere with each other and thus have a stopband covering the entire operating bandwidth. The most efficient design to date improving the spurious mode separation is the stepped CWDM filter [8], which can provide an approximately 35% wider spurious-free window than the conventional CWDM design. For Ku-band OMUX realizations, this solution covers the spurious-free region from 10.5 up to 12 GHz, enabling the realization of a 17-channel OMUX with a 1 GHz included bandwidth (10.7–11.72 GHz). However, this is still a relatively narrow spurious-free band and it will be shown that further improvements can be made in this area.

In this letter, a new Ku-band channel filter incorporating a deformed circular waveguide dual-mode cavity operating with quasi- TE_{113} mode is presented. The proposed solution, presented in Figure 1, offers a wide spurious-free band ranging from 10.6 to 12.7 GHz, which would allow the realization of a wideband OMUX (up to 2 GHz included bandwidth). A prototype fourth-order filter with two TZs was fabricated using selective laser melting (SLM) 3D printing technology and measured to verify this design.

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II. RESONANT CAVITY ANALYSIS

The basis for this work is a conventional circular waveguide cavity operating with two orthogonal TE_{113} modes. This classic CWDM cavity was closely examined in terms of Q -factor and spurious mode separation. Then, various cavity shape modifications were investigated to achieve an improvement in the spurious-free region over the state-of-the-art stepped CWDM cavity [8], while maintaining a similar unloaded Q . The edges of the basic cylindrical cavity were filleted to form a smooth "pill" shape, which was then modified using shape deformation with radial basis functions [9]. The basic principle of this technique is to define a set of points around the object that will control the deforming operation. The deformation is performed by shifting some of these control points to certain new coordinates, and the original structure is then smoothly reshaped according to this displacement. The starting point in this investigation was to squeeze the "pill" cavity shape at two longitudinal positions: at one third and at two thirds of the cavity length, maintaining axial symmetry. Such deformation suppresses the spurious mode—in this case, the TE_{211} mode—while having little effect on the operating TE_{113} mode. This can be observed in the mode chart provided in Figure 2.

Next, the shape deformation technique was coupled with an eigenmode solver from the 3D finite-element method (FEM) electromagnetic (EM) field simulator InventSim [10], with aim of finding an optimal cavity shape for Q and higher-order mode separation. The design procedure involved managing the geometrical parameters of the basic structure (cylinder length and radius) and the shape deformation parameters (deformation location and depth). Conventional, stepped, and deformed CWDM cavities are compared in Table I. It can be observed that the deformed cavity provides an over 50% wider spurious-free band than the stepped CWDM cavity, with only a minor decrease in unloaded Q . The cavity was then modified to be compliant with the restrictions of 3D printing—that is, to ensure the structure is not tilted at an angle greater than 45° along the printing axis (the cavity axis). Both modified cavities are shown in Figure 3. The 3D-printable cavity is also

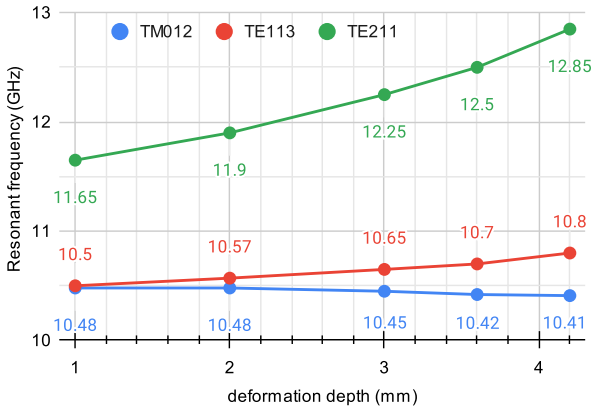


Fig. 2. The mode chart of a deformed CWDM cavity illustrating the effect of cavity deformation on the operating and adjacent modes.

TABLE I
ANALYSIS OF CYLINDRICAL CAVITY RESONATORS

cavity type	conventional CWDM	stepped CWDM [8]	deformed CWDM	modified def. CWDM*
$f_{TE_{113}}$ (GHz)	10.85	10.73	10.85	10.9
f_{lower} (GHz)	10.6	10.5	10.58	10.5
f_{higher} (GHz)	11.8	12.0	12.96	12.77
Q_u (silver)	20700	19100	18000	17800
spurious-free region (GHz)	1.2	1.5	2.38	2.27
upper stopband	$1.09 f_0$	$1.12 f_0$	$1.19 f_0$	$1.17 f_0$

* the deformed CWDM cavity was modified (cut 45°) to allow it to be 3D-printed in a single piece.

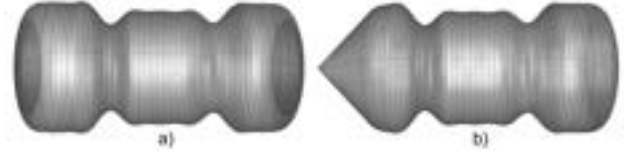


Fig. 3. Geometry of the proposed deformed CWDM cavity: a) optimal shape, b) prepared for 3D printing.

included for comparison in Table I. Adjustment of the prepared cavity resulted in a slight deterioration of its Q -factor and out-of-band performance, but nonetheless the final structure yields a significant improvement in terms of spurious mode suppression over the state-of-the-art solutions.

III. FILTER DESIGN AND FABRICATION

A. Filter Design

The new dual-mode cavities were then used to design a fourth-order Ku-band channel filter with two TZs. This filter has been designed for the lowest-frequency channel of a wideband OMUX with an operating bandwidth of 10.7–12.7 GHz. The design specifications are: center frequency $f_0 = 10.8$ GHz, bandwidth $BW = 72$ MHz and a return loss $RL > 21$ dB. The two TZs are centered at 10.73 GHz and 10.87 GHz. The synthesized coupling matrix of the filter prototype is given in Table II. The new filter design is composed of the two deformed CWDM cavities, coupled via a cross iris and interfaced at input and output with a WR-75 waveguide standard, as shown in Figure 1. The design is tuned by a total of twelve standard M3 tuning screws, with six screws for each dual-mode cavity. The design was optimized by altering the tuning screws and the widths and heights of the coupling irises. The zero-pole method was used as the optimization technique [11], [12]; this is available as a module in the InventSim EM field simulator.

B. Fabrication and Measurement

A prototype of the new filter was manufactured in one piece using the SLM process [13], [14] using a TruPrint 2000 printer

TABLE II
COUPLING MATRIX COEFFICIENTS

MS1 = 1.045	M14 = -0.196	M34 = 0.887
M12 = 0.887	M23 = 0.788	M4L = 1.045

from Trumpf with an aluminum-silicon-based alloy AlSi10Mg. It should be noted that the conductivity and surface roughness obtained in the process significantly limit the Q -factor of the filter. The prototype was tuned using silver-plated M3 screws. The measurement setup, which employs a vector network analyzer (VNA), is visible in Figure 4a.

A comparison of the simulation and measurement results is given in Figure 4. The measured passband response, shown in Figure 4a, is in excellent agreement with the simulation. The bandwidth is 2 MHz narrower than expected, but the filter otherwise matches the specifications. The in-band insertion loss at f_0 is $IL = 1.03$ dB and the return loss is greater than 21 dB. Based on the measured IL , the extracted Q of the filter is 2500, which corresponds to an effective conductivity of $\sigma_{eff} = 1.72 \cdot 10^6$ S/m. This value is comparable to other reported aluminum structures fabricated with SLM [9], [15], [16]. This estimated conductivity σ_{eff} was taken into account in the EM simulations shown in this comparison. The measured wideband response is provided in Figure 4b. The

TABLE III
COMPARISON WITH STATE-OF-THE-ART OMUX CHANNEL FILTERS

Ref.	this work	CWDM	[8]	[17]	[18]	[19]
f_0 (GHz)	10.8	10.73	10.73	11.52	20.175	11.483
FBW (%)	0.67	0.5	0.5	0.32	0.16	0.47
Order	4	4	4	5	4	4
No. of TZs	2	2	2	2	2	2
Q	2500 / 14500*	14000 (Al)	12800 (Al)	15500 (Ag)	25500 (Ag)	2900 / 12500*
IL	1.0 / 0.2*	0.2	0.25	0.45	0.4	1.0 / 0.8*
spurious-free region (GHz)	10.6–12.7	10.6–11.7	10.5–12	9.3–12.5	19.5–20.9	8.2–12.4
upper stopband	$1.17f_0$	$1.09f_0$	$1.12f_0$	$1.09f_0$	$1.04f_0$	$1.08f_0$
cavity type	deformed CWDM	CWDM	stepped CWDM	dielectric-loaded	super Q DM	spherical DM

* Q or IL of the measured 3D-printed prototype vs. simulation result for silver.

fabricated filter guarantees a very wide upper spurious-free stopband up to 12.7 GHz ($1.17 f_0$).

C. Comparison and Discussion

A comparison of the proposed deformed CWDM filter with other channel filters for OMUX applications [8], [17]–[19] is presented in Table III. It can be concluded, of the state-of-the-art solutions included in this comparison, the one presented in this letter offers the widest upper stopband clear of spurious modes. This is an improvement in terms of its spurious-free range, which is about 35% greater than that of the stepped cavity filter and nearly 90% greater than that of a classic CWDM cavity filter. The limited Q of the manufactured filter results from the high surface roughness of the 3D-printed metal, and could be overcome by silver plating the inside walls of the filter. The simulated Q -factor of the silver-plated component is $Q_{Ag} = 14500$, comparable to other reported designs. It should be pointed out that, while the Q -factor achievable with the SLM process we used for fabricating the filter prototype is much lower than for the silver plated filters machined with CNC, a state-of-the-art SLM using a proprietary process was recently reported [20], and involves postprocessing of the surfaces by depositing copper and then silver plating, resulting in an efficient conductivity of $2.5 \cdot 10^7$ S/m, which is an order of magnitude greater than what we could achieve. Consequently, with appropriate postprocessing, SLM can become competitive with traditional fabrication processes for filter applications, including in terms of insertion loss.

IV. CONCLUSION

In this letter we report on a new Ku-band channel filter employing a deformed circular waveguide dual-mode cavity shape. The new cavity design is achieved by applying shape deformation to a conventional cylindrical cavity in order to achieve a better out-of-band performance. A highly selective narrowband fourth-order filter was designed and manufactured using AM technology. The fabricated filter demonstrates a very wide upper stopband, which would allow realization of a wideband manifold-coupled OMUX without the insertion of any additional lowpass filters.

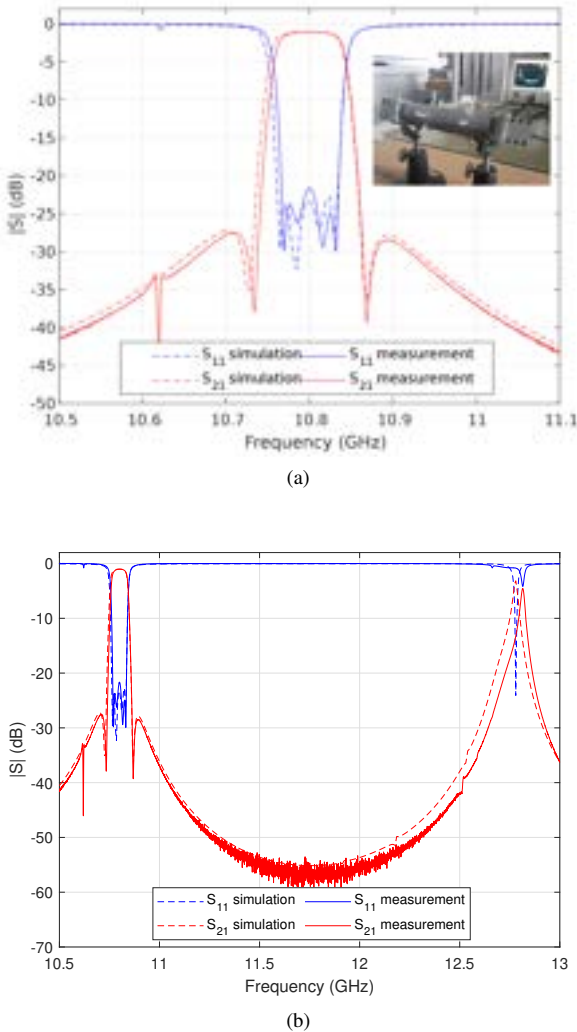


Fig. 4. Comparison of simulations and measurement for the proposed dual-mode filter. (a) passband response. (b) wideband response.

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4.3. Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators

The investigation of novel designs for microwave components benefiting from shape deformation and fabrication by AM was continued within the framework of the work described in publication [MB4], published in *IEEE Access*. This study concerned the development of rectangular waveguide filters with smooth profile, based on specially deformed dual-mode (DM) cavity resonators. The proposed resonant cavity geometry was derived by deforming the conventional rectangular waveguide cavity to achieve dual-mode operation, and adapting it to single-piece fabrication through AM. Internal coupling of the two orthogonal cavity modes (TE_{101} and TE_{110}) was accomplished through geometric alteration (i.e., breaking the symmetry of the structure), removing the necessity for additional coupling elements. Furthermore, the adopted design methodology based on free-form modeling resulted in models having smooth surface profiles, which are advantageous for high-power and low-loss applications.

The study was summarized in the following sections, starting with the design of a single resonant cavity, including a concise analysis of the operating modes and achievable couplings. A more rigorous description is provided in the publication [MB4]. Based on a single DM cavity, a second-order filter was first created to demonstrate its operation. It was then followed by the development of a fourth-order BPF composed of two DM cavities, and finally concluded with the more advanced eighth-order filter design. The fourth- and eighth-order filters were experimentally verified by prototypes, which are both characterized by very low insertion loss figures as for 3D-printed components.

4.3.1. Resonant cavity analysis

The main assumption of this design is to develop a monoblock, truly AM-oriented structure with smooth surface profile and continuous transitions between cavities. Therefore, although comprising a cascade of resonators, the developed filters will feature a single mechanical part, formed of smooth-profiled resonant cavities coupled to each other by means of gradually narrowing waveguide interconnects; thus, the separation between cavities is not as explicit as is usually the case with cascaded resonator filters. This study begins by a close examination of a single resonant cavity, operating with two orthogonal TE modes. The investigation concerns the formation of a dual-mode resonant cavity, the examination of the resonating modes and managing the couplings to allow DM operation and creation of higher-order filtering configurations. The initial resonator model, and a starting point for this investigation, was the smooth-profile waveguide cavity which was used in the 3D-printing-oriented filter design in [MB1]. That model, shown in Fig. 4.17a, operates with the fundamental TE mode (labeled TE_{110} assuming the XYZ coordinate system shown), yielding a single resonance per cavity. The primary feature of this cavity, in comparison with a classic rectangular one, is its smooth, rounded side walls, specially adapted for AM. Such a 3D shape allows monolithic 3D printing without internal supports. In the next step,

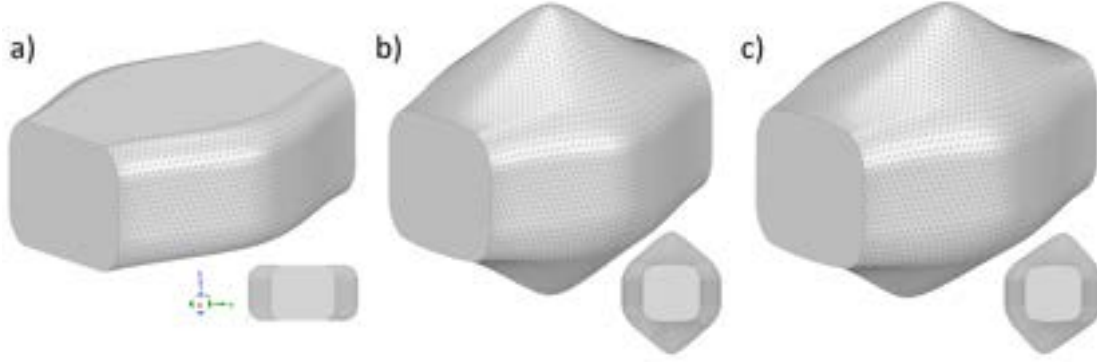


Fig. 4.17: Geometry of the proposed resonant cavities: a) basic, single-mode smooth-profile cavity, b) vertically stretched DM cavity, c) modified DM cavity with introduced disruption of symmetry.

the cavity was vertically stretched by free-form deformation techniques to enable its dual-mode operation. The top and bottom of the cavity walls were extended along the Z-axis to allow the second mode, TE_{101} , to resonate in the structure. The resulting model is shown in Fig. 4.17b. Eigenmode analysis was performed to examine the EM field in the cavity and identify the mode types, their resonance frequencies and unloaded quality factors (Q_u). Based on this analysis, it was determined that the two orthogonal quasi- $TE_{110/101}$ modes resonate at around 12 GHz, whereas the first higher-order mode is the quasi- TM_{011} mode, resonating at 13.2 GHz. The subsequent higher-order modes are of the TE_{2nm} type, resonating at around 16.5 GHz. The calculated Q_u is equal to 9500 for the first TE mode and 9200 for the other, assuming aluminum as the lossy metal ($\sigma_{Al} = 3.8 \cdot 10^7$ S/m). The magnetic field distribution (cross-sectional view at YZ plane) of the two fundamental quasi-TE modes as well as the first spurious mode, TM_{011} , is provided in Fig. 4.18. Although the cavity was modified to enable DM operation, the second resonance, however, will not be visible in the response when excited by waveguide ports at input/output (I/O); the propagating TE mode will couple only to the horizontal TE_{110} mode, and the second mode, TE_{101} , will not be excited. This is because the cavity symmetry is preserved and thus there is no energy transfer between the two degenerate field modes. In order to make use of both resonances and obtain DM operation, the two modes need to be coupled together. Several approaches may be employed

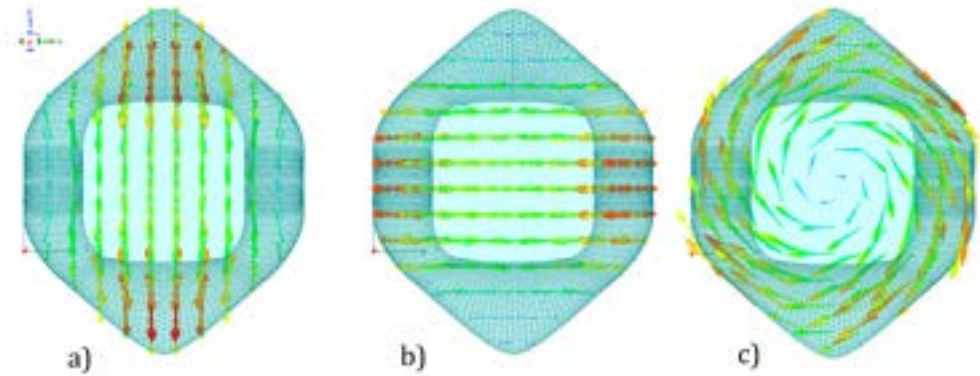
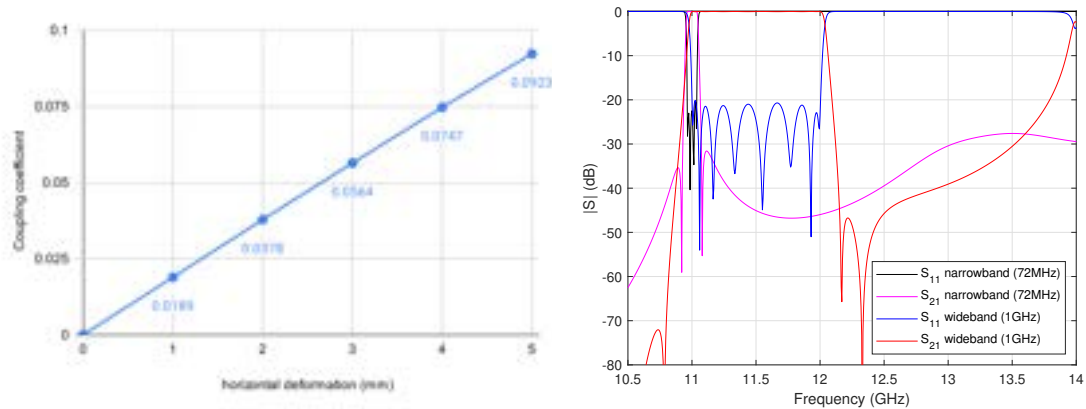


Fig. 4.18: Resonating modes (H field) in the deformed (vertically stretched) rectangular waveguide cavity: a) Mode 1: TE_{101} , b) Mode 2: TE_{110} , c) first spurious: TM_{011} mode.

to introduce such intra-cavity coupling. Introducing coupling screws into the cavity is one such method, as it disrupts symmetry; however, this modification would lower the cavity's Q-factor and add to the design complexity. Instead, what is proposed in this work is to excite the two modes through a horizontal deformation of the cavity along the Y-axis in conjunction with the previously described vertical stretching. Under this modification, the two vertices of the structure are moved outward relative to the central axis, resulting in a cavity model shown in Fig. 4.17c. As a result, the symmetry along the Y-axis is disrupted, allowing energy exchange to occur between the two modes. Fig. 4.19a shows a graph depicting the coupling coefficient between the two TE modes as a function of the horizontal offset of the cavity. As can be seen, this relationship is linear; By controlling the horizontal deformation of the cavity, one can fully control the coupling and obtain a wide range of coupling values. Such coupling span would enable the design of filters having various bandwidths, ranging from very narrow (72 MHz at $f_0 = 11$ GHz, FBW = 0.65%), up to moderately wideband (1 GHz at $f_0 = 11.5$ GHz, FBW = 8.7%), as shown in Fig. 4.19b.

The modified DM cavity (with broken symmetry) was loaded at the input and output port with WR-75 standard waveguide sections, as shown in Fig. 4.20a. The I/O coupling is controlled by compressing the side walls of the waveguide sections, forming inductive irises. In such an arrangement, the I/O waveguide ports are aligned, leading to the coupling-routing diagram (CRD) given in Fig. 4.20b. Both the source and the load are coupled to the same quasi-TE mode sustained by the cavity, while the orthogonal mode is only directly coupled with the first one by means of disturbed cavity symmetry. The resulting frequency response deviates from that of a bandpass filter, as it exhibits two reflection zeros separated by one transmission zero, as shown in Fig. 4.20e (dashed lines). An alternative inline CRD, shown in Fig. 4.20d, can be obtained when the output port is rotated 90° with respect to the input, as in Fig. 4.20c. In this configuration, the incident EM wave entering through the input waveguide excites the first ("horizontal") mode, and the rotated output section is correspondingly coupled to the second ("vertical") mode. As a result, a second-order BPF is obtained, generating a two-pole response with center frequency $f_0 = 11.5$ GHz and



a) Intra-cavity coupling vs. symmetry disruption. b) Achievable passband range (EM simulation).
Fig. 4.19: The achievable coupling coefficient between the two TE modes operating in the DM cavity and the resulting allowable range of filter bandwidths.

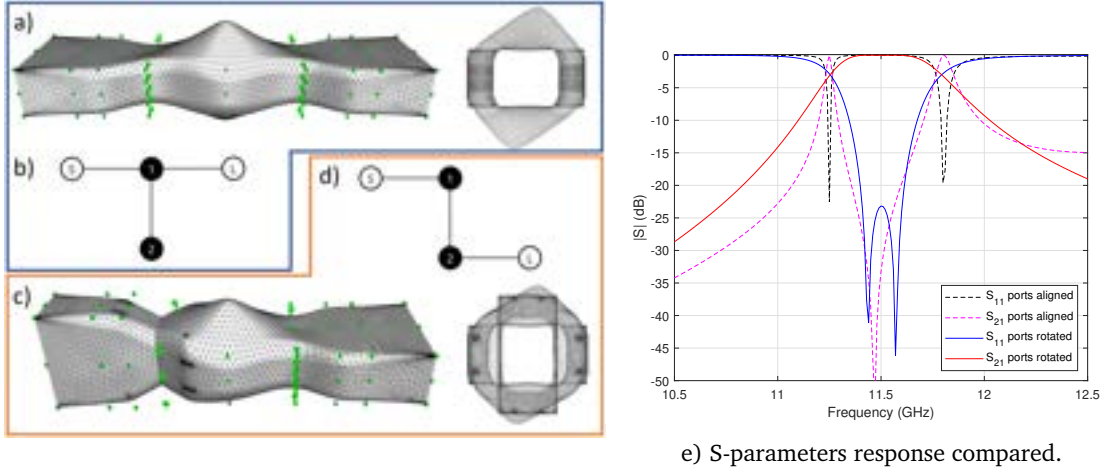


Fig. 4.20: Loaded DM cavity configurations and their respective S-parameter responses (EM simulation).

a 200 MHz bandwidth, as shown in Fig. 4.20e (solid lines). Such a modified DM cavity design can function as a high- Q , second-order BPF with a 90° output rotation or it can be used for developing more advanced filter structures.

4.3.2. Fourth-order filter design

The described single-cavity filtering structure from Fig. 4.20c was expanded to develop a fourth-order BPF composed of two cascaded DM cavities, interconnected by a smoothly varying quasi-square iris. The design of the filter is visually portrayed in Fig. 4.21. Each DM cavity supports the resonance of the two TE modes discussed in the previous section, and the quasi-square aperture of the central waveguide constriction controls the coupling of the corresponding modes between the two cavities. The orientation of the I/O waveguide connections determines which modes are excited and received at each end. Rotation of the output port by 90° results in the coupling-routing diagram shown in Fig. 4.21a, where modes 1 and 2 resonate in the first DM cavity, and modes 3 and 4 operate in the other. Such a filter topology produces a four-pole response with one TZ, as shown in the ideal characteristic in Fig. 4.21c. The coupling matrix corresponding to that filtering response is provided in Fig. 4.21b. Compared to the arrangement with both ports in alignment, such a configuration offers a 90° rotation of the ports at the expense of an additional TZ below the passband. The filter was designed to meet the following specifications: passband 11.25–11.75 GHz (center frequency $f_0 = 11.5$ GHz, bandwidth BW = 500 MHz), return loss RL > 20 dB, TZ location: $f_{TZ} = 12$ GHz. Because of its monoblock, smooth-profiled structure, direct analytical synthesis of the filter cannot be applied. Instead, the filter design was carried out in two steps. First, the initial physical model was reached by tuning the deformed DM cavities to resonate at the specified frequencies. Then, the cavities were interconnected and coupled to the I/O waveguide sections, and the couplings were set approximately by adjusting the iris size. Subsequently, a comprehensive full-wave design optimization was performed using the framework described in

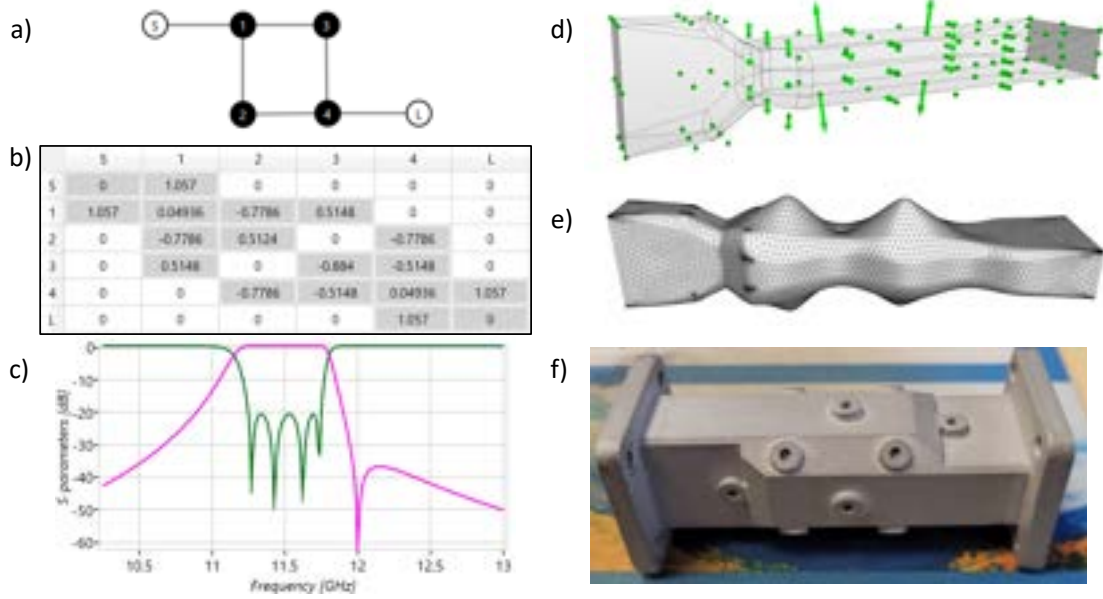


Fig. 4.21: Design of the fourth-order DM filter with a 90° rotated output: a) coupling-routing diagram, b) synthesized coupling matrix, c) ideal filtering characteristic, d) shape deformation definition, e) EM model of the BPF, f) photograph of the manufactured prototype.

Chapter 3.2. The free-form deformation setup used in the design process involved the definition of the control points and displacements as in Fig. 4.21d. The parameters governing the RBF interpolation were maintained constant, with the interpolating function defined as a polyharmonic spline of order $k = 3$, as in eq. (3.8). This selection of basis function ensures that the resulting deformed surface exhibits a smooth and continuous profile. The zero-pole method was used to optimize the filter, i.e., the goal function involved matching the locations of zeros and poles in the filter's response with those in the ideal characteristic [99, 111]. The optimization controlled twelve variables, including the lengths of cavity sections and deformation parameters responsible for shaping the cavities and the irises.

The optimized filter design, shown in Fig. 4.21e, was prepared for fabrication and 3D printed as a single mechanical part with aluminum alloy (AlSi10Mg) using the SLM process. The fabricated prototype, presented in Fig. 4.21f, was measured using the measurement setup with a VNA and WR-75 waveguide-to-coax adapters, and the results are compared with EM simulation in Fig. 4.22. As can be seen in Fig. 4.22a, the passband is shifted roughly 40 MHz above the specified band and the in-band RL level is at least 17 dB, which is a very good result considering that no tuning was applied. The insertion loss, shown in Fig. 4.22b, is equal to 0.11 dB at f_0 , which results in an estimated Q-factor of around 3500. The effective metal conductivity based on the measured IL was estimated as $\sigma_{\text{eff}} = 9 \cdot 10^6$ S/m. This is considered an excellent result for a 3D-printed aluminum component, but most of all, the achieved filter response validates the design idea based on the proposed DM cavities as well as the overall design approach using smooth geometry deformation.

4.3.3. Eighth-order filter design

The next design which was developed under this work was an eighth-order BPF composed of four cascaded dual-mode cavities. The cavities were arranged in a cascade, one-by-one, and grouped in two pairs, each pair creating a quadruplet configuration. Inter-cavity couplings were realized by integrated, smooth waveguide transitions forming quasi-rectangular irises. Such an arrangement of cavities and irises leads to a coupling-routing diagram shown in Fig 4.23a, and results in an eighth-order BPF with four TZs, as in Fig 4.23b (each quadruplet produces two symmetrical TZs). For the chosen filter topology, the irises connecting cavities 1 and 2 (first quadruplet)

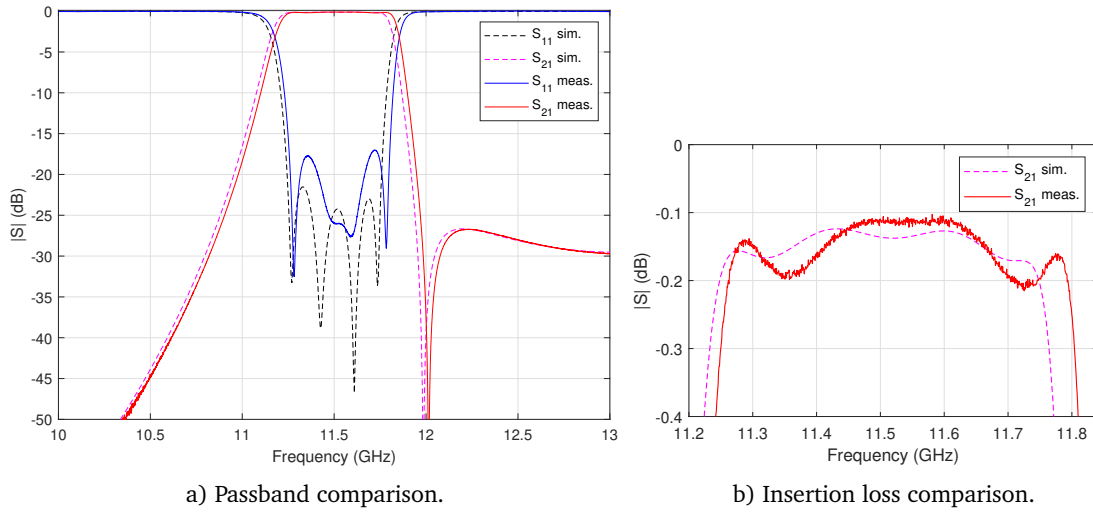


Fig. 4.22: A comparison of simulation and measurement of the fourth-order DM filter with rotated output and one TZ. EM simulation: dashed lines, measurement: solid lines. The EM simulation considers a lossy metal with $\sigma_{\text{eff}} = 9 \cdot 10^6 \text{ S/m}$.

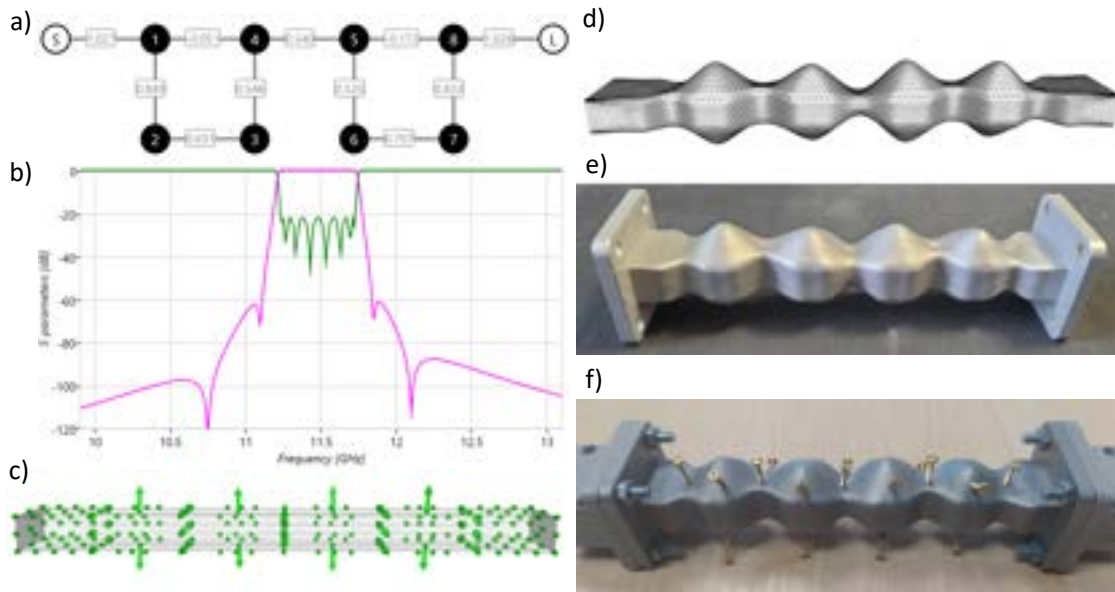


Fig. 4.23: Design of the eighth-order DM filter with four TZs: a) coupling-routing diagram with indicated normalized coupling values, b) ideal filtering characteristic, c) shape deformation definition, d) EM model of the BPF, e) 3D-printed prototype, f) measured prototype with tuning screws.

and between cavities 3 and 4 (second quadruplet), need to couple the corresponding modes of the orthogonal pairs—hence a quasi-square iris shape was used. Conversely, the middle iris, interconnecting the two quadruplets, should allow the coupling of only one pair of TE modes (i.e., mode 4 and 5 in the CRD). For this reason, a narrow quasi-rectangular slit was implemented in the initial filter design, and was later optimized into a quasi-elliptical shape (more details can be found in [MB4]). The design specifications were the same as for the fourth-order filter prototype: center frequency $f_0 = 11.5$ GHz, bandwidth $BW = 500$ MHz, and return loss level $RL > 20$ dB. The two pairs of TZs were placed nearly symmetrically about the passband: $f_{TZ_1} = 11.1$ GHz, $f_{TZ_2} = 11.9$ GHz, $f_{TZ_3} = 10.8$ GHz, $f_{TZ_4} = 12.1$ GHz. The initial model was optimized using the same approach based on the zero-pole method [99]. The RBF used was once again a polyharmonic spline of order $k = 3$. The shape deformation definition with all the control points is shown in Fig 4.23c. There were 22 parameters used in the optimization process: four variables describing the length of each cavity, ten variables modeling the shape of the irises, and eight parameters controlling the deformation of the cavities. The optimized EM model of the filter is presented in Fig 4.23d.

The optimized filter design was prepared for 3D-printing and fabricated in metal using a powder bed fusion process by Northern Waves AB [101]. The material used in the process was aluminum alloy (AlSi10Mg), and the powder layer thickness was $30 \mu\text{m}$. The prototype, presented in Fig 4.23e, was 3D-printed in one piece along the propagation dimension, i.e., from the input flange to the output. Physical polishing by sandblasting was applied in post-processing, which resulted in a reduction of the average surface roughness (R_a) from $8\text{--}10 \mu\text{m}$ as built, down to $4\text{--}6 \mu\text{m}$, according to the manufacturer. The fabricated component was measured and the measurement results are evaluated against the simulation in Fig. 4.24. The measured response exhibits good agreement with the simulation, though it is shifted by approximately 20 MHz below

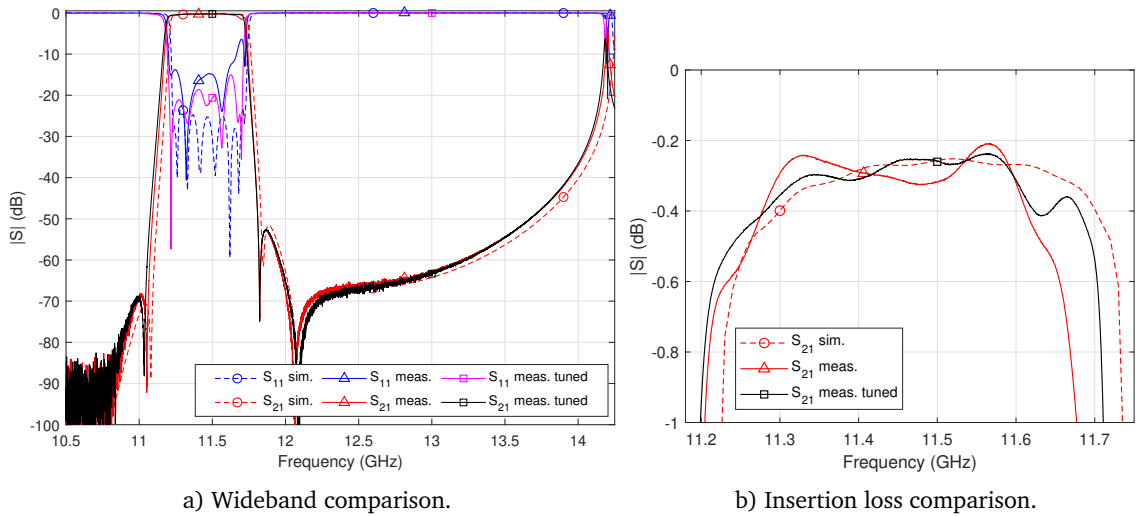


Fig. 4.24: A comparison of simulation and measurement of the eighth-order DM filter with four TZs. EM simulation: dashed lines, measurement: solid lines. The EM simulation considers a lossy metal with $\sigma_{\text{eff}} = 1 \cdot 10^7$ S/m.

the specified band. As can be inferred from the S_{11} plot in Fig. 4.24a, the manufactured prototype is not fully matched, with RL level generally exceeding 15 dB across the passband, except for a mismatch near the upper band edge. However, the S_{21} response, including the positions of the transmission zeros, accurately reflects the simulation results. In order to improve the in-band RL, tuning screws were placed in the cavities and irises (16 in total), as shown in Fig 4.23f, and the prototype was manually tuned. The improved filter response after tuning is also plotted in Fig. 4.24.

A close-up on the in-band insertion loss is provided in Fig. 4.24b, where the measured IL before and after tuning is compared with EM simulation. Finite metal conductivity of the surfaces was assumed in the simulation to match the measurements. The IL at f_0 is equal to 0.25 dB and ranges from 0.25 to 0.55 dB in the passband. These values closely match the simulated effective conductivity of $\sigma_{\text{eff}} = 10 \text{ MS/m}$, which is a relatively high value compared to the conductivity of ideal aluminum (38 MS/m), and the highest result among the 3D-printed metal components presented throughout this work. Based on the measurements, the estimated Q of the fabricated filter equals 4500, which is also higher than the Q -factors of the previously discussed components. The primary reason for the observed improvement in Q is that this design was treated with sandblasting to polish the inner surfaces, which in turn decreased the conduction loss attributed to high surface roughness. The effectiveness of surface polishing on the overall power loss in 3D-printed microwave components will be discussed in more detail in Chapter 4.5.

In summary, the designs presented in [MB4] are yet another demonstration of novel passive microwave circuits that were developed using shape deformation and benefit from a complex geometry and smooth surface profile. The coupling inside the DM cavity is achieved by modifying its shape, thus eliminating the need for additional coupling elements. Such models and geometries can only be fabricated by means of additive manufacturing. The two manufactured components exhibit low insertion loss, relatively high Q and a smooth profile, making them potentially useful in a variety of applications. The lower-order 3D-printed filter produced results so satisfactory that they could be accepted in their current form, without manual tuning. It is also noteworthy that the eighth-order filter prototype weighs only 75 g, which represents a relatively small payload for a waveguide filter operating in the X/Ku-band [112, 113].

**[MB4] Low-Loss 3D-Printed Waveguide Filters Based on
Deformed Dual-Mode Cavity Resonators**

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RESEARCH ARTICLE

Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators

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ABSTRACT This paper introduces a new type of waveguide filter with smooth profile, based on specially designed dual-mode (DM) cavity resonators. The DM cavity design is achieved by applying a shape deformation scheme. The coupling between the two orthogonal cavity modes is implemented by breaking the symmetry of the structure, thus eliminating the need for additional coupling elements. The modes operating in the cavity are carefully analyzed and a scheme for managing the spurious modes is discussed. Two filter prototypes employing the designed DM cavities are developed and described in detail. The first design is a fourth-order bandpass filter (BPF) with a 90° rotated output and a transmission zero (TZ), whereas the second design is an eighth-order filter with four TZs. Both designs are developed, taking into account the limitations of 3D printing technology to enable their single-piece fabrication without internal supports. The structures benefit from additive manufacturing (AM) by having a smooth surface profile and reduced weight, which is often highly desirable for high-power and low-loss applications. Filter prototypes were manufactured using selective laser melting (SLM) from aluminum alloy and tested to validate the designs. Measurement results are consistent with the simulation and prove the validity of the proposed solutions. Both measured BPF prototypes demonstrate low insertion loss, i.e., 0.11 dB and 0.25 dB for the fourth-order and eighth-order designs, respectively. The estimated Q-factors reach 3500 and 4500, which is a very good result for 3D-printed parts.

INDEX TERMS Additive manufacturing, design optimization, dual-mode cavity resonator, dual-mode waveguide filter, shape deformation.

I. INTRODUCTION

Additive manufacturing (AM) has become one of the key emerging technologies for efficient manufacturing of passive microwave components, such as waveguide filters, couplers and antennas [1], [2], [3], [4], [5], [6], [7], [8], [9], [10]. The use of 3D printing methods for making high-frequency electronics has grown rapidly over the last decade, due to their low cost and high flexibility in creating complex 3D models. With the advancement of professional AM processes in the industry, 3D printing can not only be used for quick prototyping, but also for professional

manufacturing of high-frequency components, along with the traditional subtractive methods such as Computer Numerical Control (CNC) milling. The freedom in obtaining complex 3D structures can help overcome the limitations of the conventional manufacturing methods, thus opening the way for new solutions in microwave component design. The new microwave components developed using AM can have non-standard profiles, leading to improvements in their electromagnetic (EM) performance, such as lower loss, superior quality factor Q , wider spurious mode separation [11], [12], [13], [14], and better power handling capability and thermal stability [15]. 3D printing can also help shrink the size of systems by combining several useful functions in one component [16], [17]. However, to use AM processes

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effectively in microwave engineering, specialized tools are required for the design, simulation, and optimization of 3D structures with complex and smooth geometries.

One way of addressing this challenge is to develop new design methods oriented toward fabrication by AM, e.g., analytic method for the synthesis of waveguide filters with a smooth profile [18]. Another possible solution focuses on adapting computer-aided design (CAD) tools, well established in microwave engineering, to provide greater freedom in 3D modeling. Enhancing conventional 3D CAD of high-frequency passive components with the shape deformation technique was presented in [19]. The geometry modification algorithm based on radial basis function (RBF) interpolation was developed and integrated with a 3D finite element method (FEM) in the EM field simulator InventSim [20], which is utilized as a main tool to optimize the presented designs. This technique makes it possible to create truly arbitrary 3D models by means of reshaping predefined basic geometries. The working principle of this method is based on manipulating a set of points defined around the object which control the deforming operation. The deformation is performed by shifting some of the control points to certain new coordinates. The original structure, treated as a 3D mesh of points, is then smoothly reshaped according to this displacement. This method was already applied to the development of several microwave filter designs and demonstrated very good results, e.g., by improving the out-of-band performance of waveguide filters [11], [19]. In this work, the shape deformation technique is applied to design smooth-profile rectangular waveguide dual-mode (DM) cavity filters suitable for direct metal additive manufacturing using the selective laser melting (SLM) process. Initially, deformation is applied to form a cavity geometry, facilitating the coupling of the two orthogonal modes together and achieving dual-mode operation in the cavities. Subsequently, this concept is further developed and employed to design a waveguide filter structure with a smooth profile, prepared for 3D printing in one piece, thereby eliminating the need for parts assembly.

Many 3D-printed filter designs implementing various cavity types have been reported in the literature, including dual- or multi-mode filter structures [15], [21], [22], [23], [24], [25], [26], [27], [28], [29], [30], [31]. Some notable examples are the designs based on spherical cavities [15], [21], ellipsoidal cavities [29], [30], or dimpled ellipsoidal cavities [31]. These designs take advantage of AM because spherical or ellipsoidal cavity shapes are often difficult to fabricate using conventional subtractive methods. However, 3-D printing of spherical cavities may result in overhangs and degraded surfaces due to the lack of internal supports, which was discussed and demonstrated in [6]. Additionally, in all the referenced DM filter designs, a coupling mechanism is needed to excite the two orthogonal cavity modes and allow for energy transfer between them. In [15], [26] and [31], internal coupling between the two modes operating in the cavity is implemented by inserting coupling elements, corner

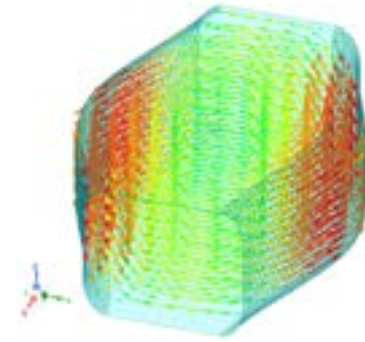


FIGURE 1. Operating fundamental TE₁₁₀ mode (H field) in the basic smooth-profile rectangular waveguide cavity.

cuts or ‘dimples’, thereby breaking the symmetry of the cavity. In [29] and [30], however, the two modes are not coupled together, but rather excited separately by rotating the input and output ports (i.e., direct coupling from the source or load to both cavity modes). This paper addresses these issues while maintaining complex cavity geometry that can only be made by 3D printing. In the proposed method, the second cavity mode is excited by perturbing the cavity shape to get a non-symmetric cross-section, similarly to what was done in [32].

The designs proposed in the paper are based on ideas originally described in [33], which have been further developed and expanded. To this end, Section II presents a detailed study of a single DM resonant cavity. The operating fundamental and higher-order modes resonating in the cavity along with mode coupling mechanisms are identified and analyzed. Section III describes the whole process of filter design and fabrication. For clarity, a simple fourth-order filter, initially presented in [33], is thoroughly discussed and compared with other solutions in Section III-A. Next, a more challenging design, an eighth-order DM filter composed of four cascaded DM cavities, is presented in Section III-B. The prototype of the filter is fabricated directly in metal using SLM, and measurement results are compared with EM simulation. Our experimentally verified design is compared with other state-of-the-art 3D-printed DM filter solutions, followed by a discussion in Section III-C and concluding remarks in Section IV.

II. RESONANT CAVITY ANALYSIS

A. MODE ANALYSIS

The starting point for this study is a waveguide cavity with a smooth profile, which was adapted for 3D printing [34]. The cavity resonates with the fundamental mode, which is the TE₁₁₀ mode, whose magnetic (H) field distribution is shown in Fig. 1. The mode numbering follows the XYZ coordinate system shown in the figures. Such a single-mode resonant cavity has a constant height and smooth, rounded side walls. To obtain DM operation, the cavity was modified by applying the shape deformation technique. The top and bottom walls of the cavity were stretched vertically

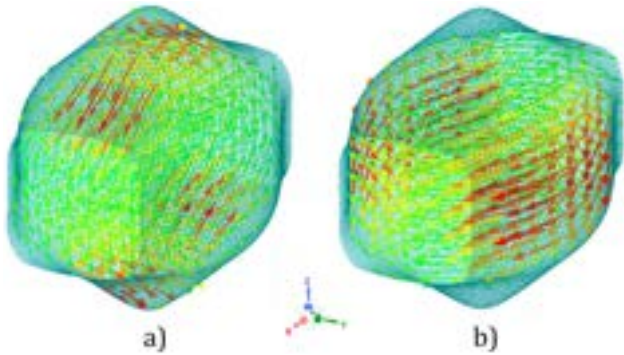


FIGURE 2. Operating fundamental modes (H field) in the deformed (vertically stretched) rectangular waveguide cavity: a) Mode 1: TE₁₀₁, b) Mode 2: TE₁₁₀.

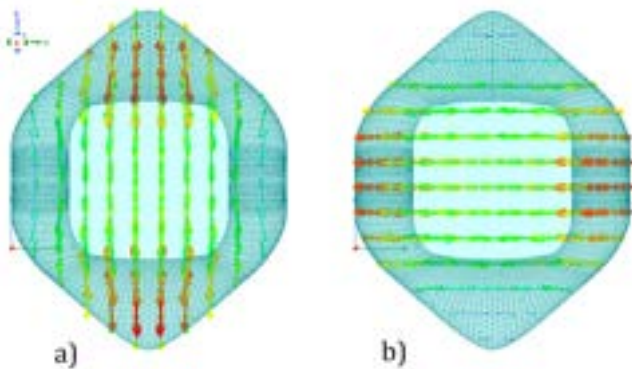


FIGURE 3. Operating fundamental modes (H field) in the deformed (vertically stretched) rectangular waveguide cavity: a) Mode 1: TE₁₀₁, b) Mode 2: TE₁₁₀.

(symmetrically along the Z-axis), allowing the second quasi-TE₁₀₁ mode to resonate in the orthogonal direction, which is illustrated in Fig. 2. A detailed mode analysis was carried out for proper characterization of the resonating modes and constraints imposed by higher-order spurious modes. Eigenmode simulations in Ansys HFSS were performed to determine the resonance frequencies, estimate the unloaded quality factors Q_u , and examine the EM field in the cavity. The results are summarized and compared with a reference spherical cavity in Table 1. The reported Q_u values were calculated for ideal aluminum (electrical conductivity equal to $\sigma = 3.8 \cdot 10^7$ S/m.). The two orthogonal quasi-TE_{110/101} modes resonate at around 12 GHz, yielding a high Q-factor. A cross-sectional view (YZ plane) of the H field distribution of the two fundamental quasi-TE modes operating in the cavity is shown in Fig. 3a and Fig. 3b. The first higher-order mode is the quasi-TM₀₁₁ mode, resonating at about 13.2 GHz, whose H field distribution is provided in Fig. 5. The next higher-order modes are of the TE_{2nm} family, and resonate at around 16.5 GHz.

The influence of vertical stretching on the Q_u of the two operating quasi-TE modes and the quasi-TM mode is shown in Fig. 4. As expected, stretching increases the volume of the DM cavity along one direction, resulting in a slight increase

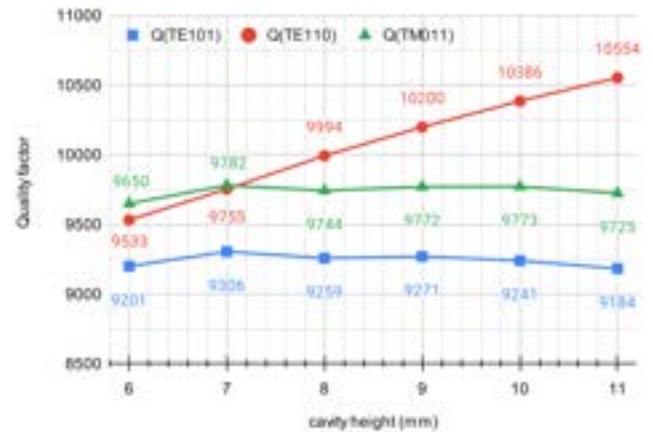


FIGURE 4. Chart showing the influence of cavity height (vertical deformation) on the quality factor of the two operating TE modes and higher-order TM mode. Analysis performed assuming the conductivity of ideal aluminum, $\sigma = 3.8 \cdot 10^7$ S/m.

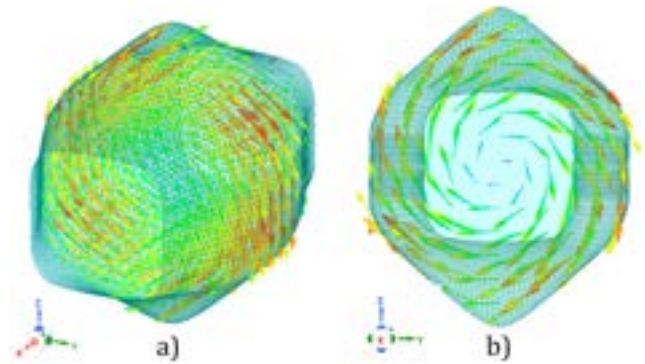


FIGURE 5. TM₀₁₁ mode (H field) in the deformed rectangular waveguide cavity: a) 3D view, b) side view (YZ plane).

in Q_u for one of the operating modes. Similarly, change of the orthogonal cavity dimension (horizontal) would lead to a similar increase in Q_u for the orthogonal mode. Nonetheless, the Q-factors of all three examined modes remain similar and falling in the range of 9000–10500, assuming ideal aluminum conductivity. These values can be considered a good result as they remain satisfactory albeit the Q_u is degraded by 12–15% with respect to the spherical cavity solution, reported in Table 1.

B. SUPPRESSION OF THE TM MODE

As indicated above, the next higher-order mode excited in such a deformed cavity is the TM₀₁₁ mode. The H field distribution of this mode is illustrated in Fig. 5. This mode can affect the spurious-free operation of the dual-mode cavity. Our numerical tests show that, depending on the cavity shape, the TM₀₁₁ mode can resonate close to both useful TE modes, or even become a fundamental one. In filter applications, maintaining the ideal symmetry prevents the TM₀₁₁ mode from coupling to the TE modes and adding an extra reflection zero in the filter response. However, if the symmetry is broken, it may lead to energy loss visible as a “notch” in

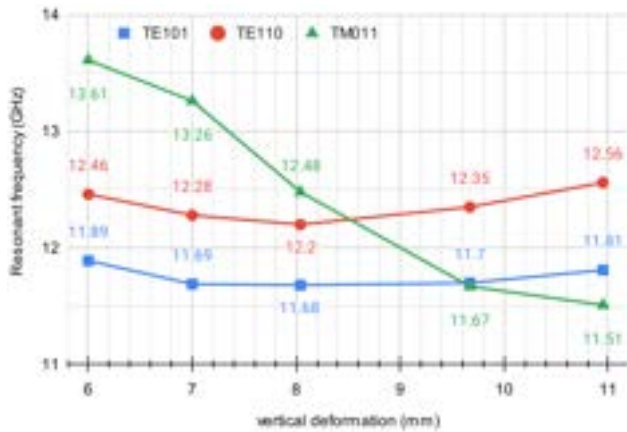


FIGURE 6. Mode chart illustrating the influence of cavity height (vertical deformation) on the two operating TE modes and higher-order TM mode.

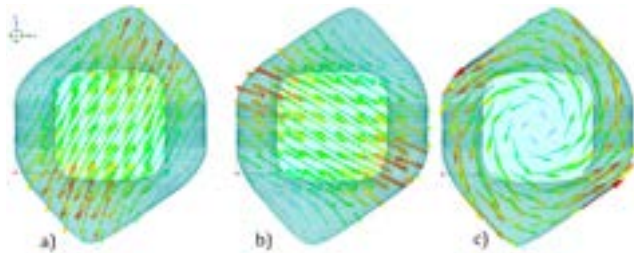


FIGURE 7. Operating modes (H field) in the modified DM cavity: a) Mode 1, b) Mode 2, c) TM mode.

or near the passband. Nonetheless, if the resonance occurs outside of the operational band, it is strongly suppressed and would be practically irrelevant to the filter response. Therefore, during the design stage, efforts should be made to keep the TM mode away from the operating frequency range. The resonance frequency f_{TM} is directly associated with the height of the cavity and can be controlled by vertical stretching: the greater the height, the lower the frequency of the spurious mode f_{TM} . This relationship, illustrated by a mode chart in Fig. 6, must be considered to keep the TM mode away from the operating frequency of the fundamental modes. For this reason, we set a limit on the cavity height and followed it during the design process.

C. COUPLING ANALYSIS

Once the modes operating in a single DM cavity are properly arranged, the proposed resonant structure can be used to create a simple two-pole bandpass filter (BPF). To this end, the cavity from Fig. 2 needs to be connected to input and output waveguide sections. Please note that a simple input/output arrangement will not lead to the creation of a two-pole filter, because the cavity is symmetric. The excitation from the input/output (I/O) waveguide operating with the fundamental TE mode will couple only to the horizontal TE₁₁₀ mode, and the second mode, TE₁₀₁, will not be excited. There are several possible solutions to establish coupling to the other TE mode and achieve effective DM

TABLE 1. Mode and Q_u analysis of DM cavity resonators.

cavity type	DM cavity in Fig. 3	Modified DM cavity in Fig. 7	Spherical cavity
fundamental modes, f_r (GHz)	11.83 (TE ₁₀₁), 12.14 (TE ₁₁₀)	11.69 (TE ₁₀₁), 12.28 (TE ₁₁₀)	11.69 (TM ₁₀₁)
first spurious mode, f_s (GHz)	13.23 (TM ₀₁₁)	13.26 (TM ₀₁₁)	16.48 (TM ₂₀₁)
fundamental modes, Q_u^*	9210 (TE ₁₀₁), 9525 (TE ₁₁₀)	9180 (TE ₁₀₁), 9520 (TE ₁₁₀)	10885 (TM ₁₀₁)

* Q_u given for ideal Aluminum ($\sigma = 3.8 \cdot 10^7$ S/m.).

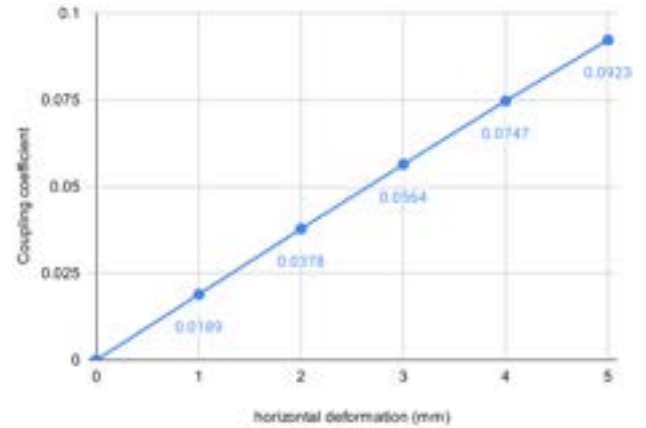


FIGURE 8. Chart demonstrating the variation of the coupling coefficient between the two TE modes operating in the DM cavity vs. horizontal deformation.

operation in the cavity, resulting in a simple two-pole filter. For instance, coupling screws could be introduced in the cavity to break the symmetry, but such an action would reduce the Q-factor of the cavity and increase the complexity of the design. One could also introduce direct coupling from the source or load to both modes, as was done in [29]. However, such a solution does not introduce internal coupling between the two orthogonal modes, which will be necessary for creating the higher-order BPF configurations that we intend to design. Instead, we propose the two modes to be excited by introducing a horizontal deformation of the cavity (in the Y-axis direction) along with the previously specified vertical stretching operation. Specifically, the two vertices of the structure are moved away from the central axis. This way, the symmetry along the Y-axis is broken, leading to energy exchange between the two modes. The modified DM cavity with the proposed horizontal displacement is shown in Fig. 7. By varying the value of the horizontal deformation, we can control the coupling between the two modes operating in the DM cavity, as shown in Fig. 8. The achievable coupling values allow for the design of a wide range of narrow- and wideband filter responses. In our simulation tests, we have successfully designed filters with fractional bandwidths between 0.65% and 8.7%.

We also performed a mode analysis of the modified resonant structure and compared the results with the symmetrical one in Table 1. We found that this modification does not affect the resonances and Q-factors significantly. The deformation

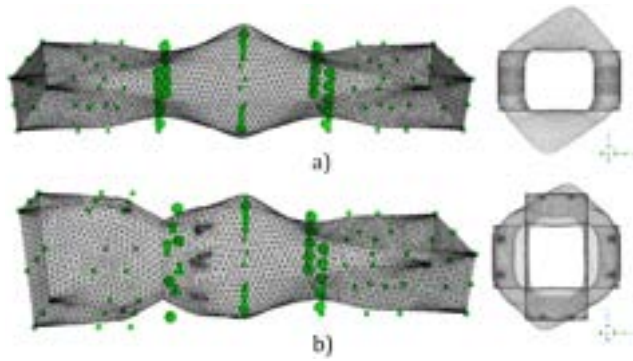


FIGURE 9. Geometry of the resonant cavity with input/output ports: a) modified DM cavity with introduced horizontal offset, b) modified DM cavity with 90-degree rotated output.

slightly rotates the polarization of the two orthogonal modes, as seen when comparing the field distribution in Fig. 7 and Fig. 3. In theory, the mode polarization shift could enable direct coupling from the source or load to the second (orthogonal) TE mode. However, a series of EM simulations were conducted and revealed that such additional coupling is negligible and we can ignore it in the design process. This direct coupling would occur if we rotate the I/O ports relative to the cavity, resulting in the appearance of an additional transmission zero (TZ) close to the passband. This coupling mechanism is well documented and has been successfully applied in [29], but is not used in this work.

The modified version of the DM cavity (with the broken symmetry along the Y-axis) was connected with standard WR-75 waveguide sections, as shown in Fig. 9a. This cavity arrangement aligns the input and output waveguide ports, matching the coupling-routing diagram (CRD) shown in Fig. 10a. In this configuration, both the source and the load are coupled to only one of the two TE modes supported by the cavity, and this mode couples to its orthogonal counterpart via perturbed symmetry. The frequency response of such a structure is not that of a BPF, as it consists of two reflection zeros separated by a TZ. In order to obtain a two-pole BPF response, an inline topology is required. Such a configuration, displayed in Fig. 10b, can be created when the output port is rotated 90° with respect to the input port, as shown in Fig. 9b. This way, the incident EM wave from the input waveguide section excites the first TE mode, and the output section is coupled to the second mode. In all the designs considered in this work, the I/O coupling is controlled by means of inductive irises formed by compressing the side walls of the waveguide sections (as indicated by green arrows in Fig. 9). The modified DM cavity design from Fig. 9b can function as a high- Q , second-order BPF with a smooth profile and a 90° twisted output. Alternatively, the DM cavity can serve as a building block for higher-order filter configurations.

III. FILTER DESIGN

The study of the resonant modes and coupling schemes that are implementable in the proposed cavity presented

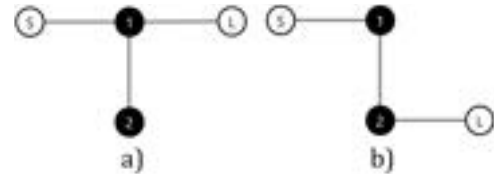


FIGURE 10. Coupling-routing diagram of the modified single-cavity DM filter: a) with the same orientation of input and output waveguide, as shown in Fig. 9a, b) with a 90° rotation of the output, as shown in Fig. 9b.

in the previous section is essential for developing more advanced BPF designs with quasi-elliptic response. In this paper, we examine two such designs in detail. The first one is a fourth-order BPF with a 90° rotated output. This design was presented as a preliminary example in [33]. It offers an advanced filtering function (four-pole response with one TZ) in a compact volume, integrated 90° polarization rotation and high Q -factor, and can be fabricated as a single piece using direct metal AM. The second one is an eighth-order BPF with four TZs, which consists of four cascaded DM cavities interconnected with quasi-square or quasi-elliptical irises. This is a new design aiming for high-power and low-loss applications, and has key features such as compact size, high Q , and high power handling due to its smooth profile. Again, the design is verified experimentally by 3D printing as a single component. All filter prototypes presented in this paper are interfaced with a WR-75 rectangular waveguide and manufactured from aluminum alloy (AlSi10Mg) using the SLM process.

A. FOURTH-ORDER FILTER DESIGN

The first design is a fourth-order BPF consisting of two cascaded DM cavities. The model is built by expanding the single-cavity configuration from Fig. 9b to include a second DM cavity and connecting them via a quasi-square iris. The developed 3D model of the filter and its corresponding coupling-routing diagram are presented in Fig. 11. Each DM cavity supports the resonance of two TE modes, and the quasi-square iris allows coupling to the corresponding modes in the other cavity. Depending on their orientation, the I/O waveguide connections can be coupled to either one of the two modes. The presented asymmetric-box coupling scheme, shown in Fig. 11b, is obtained when the output port is rotated by 90° with respect to the input. This coupling scheme results in a four-pole response with one TZ and offers a 90° rotation of I/O ports at the expense of the additional TZ below passband that would occur if the output port was not rotated.

The design specifications of the filter are: passband 11.25–11.75 GHz (center frequency $f_c = 11.5$ GHz, bandwidth BW = 500 MHz) and in-band return loss RL > 20 dB. The TZ is positioned close to the passband, at 12 GHz. The passband frequency was chosen to fit within the operating band for the WR-75 waveguide standard, which covers the frequency range of 10–15 GHz. The initial design was optimized following the procedure described in [19] and [34].

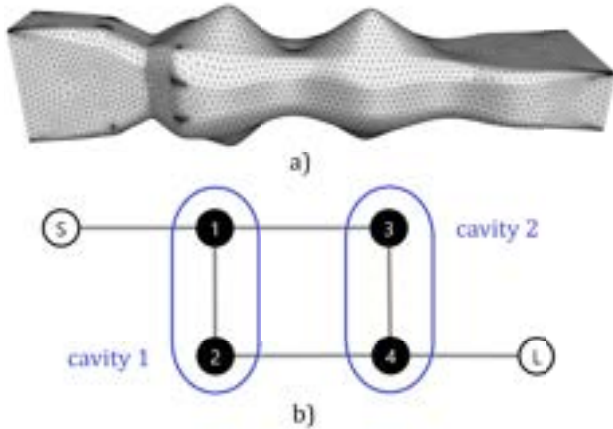


FIGURE 11. Geometry of the fourth-order DM cavity filter: a) 3D model, b) coupling-routing diagram with indicated physical cavities.

Due to the smoothly contoured nature of the filter, direct analytical synthesis is not feasible. Therefore, the initial synthesis adopts a two-step approach. The first step involves the design of dual-mode deformed cavities to resonate at the specified filter passband. Subsequently, the cavities are coupled, and a comprehensive full-wave optimization was performed. The process involved twelve optimization variables, including two geometric parameters describing the lengths of cavity sections, as well as deformation parameters, i.e., four variables standing for the vertical stretching and horizontal deformation of the cavities, and six variables controlling the shape of the irises. The zero-pole method was used for optimization, as it is a highly effective strategy for microwave filter design [35], [36], [37]. The parameters of the shape deformation based on RBFs were kept fixed: the interpolating function was a polyharmonic spline with power $k = 3$. This type of basis function ensures a very smooth profile of the deformed surface is attained. Additionally, the design optimization imposed constraints on the maximum cavity height to keep the TM modes away from the passband, as discussed in Section II-B.

The EM simulation result of the optimized filter design is presented in Fig. 12. The filter response close to the passband perfectly matches the specification. It should be noted that there is another TZ visible at ca. 14 GHz, and a suppressed resonance occurring at ca. 15 GHz. Eigenmode analysis confirmed that these are associated with the resonance of higher order TE modes. In this case, the additional TZ is beneficial to the design and contributes to good upper stopband rejection up to around 14.5 GHz.

To validate the design, a prototype of the fourth-order filter was manufactured using the SLM process. The filter was 3D-printed as a single piece with no internal supports. The part was oriented vertically in the printing platform, i.e., the printing direction followed the filter's length, from one I/O port to the other. The external dimensions of the printed structure are $90.5 \times 38 \times 38$ mm. The fabricated component was measured using the measurement setup with

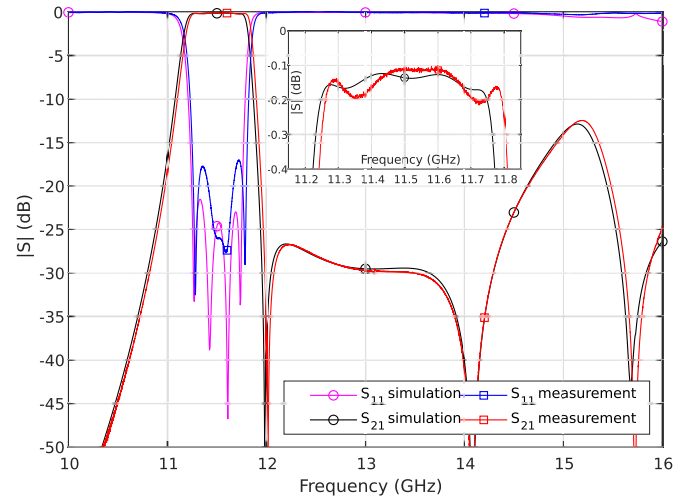


FIGURE 12. Measurement result compared with EM simulation of the fourth-order filter with a 90° output rotation. Inset: close-up view of the in-band insertion loss.

a vector network analyzer (VNA) and WR-75 waveguide-to-coax adapters, as shown in Fig. 13. The comparison of the measurement and EM simulation results provided in Fig. 12 shows excellent agreement with the simulated model. The measured response is shifted approximately 40 MHz above the specified frequency band, but the in-band and out-of-band behavior is accurately reproduced. The return loss level in the passband is greater than 17 dB, which is a very good result since no tuning was applied to achieve this response. The insertion loss (IL) of the 3D-printed filter is shown in the inset in Fig. 12. It equals 0.11 dB at f_c and varies between 0.1 and 0.2 dB across the passband. This corresponds to the estimated Q of the filter of around 3500. Based on the IL value, we estimated the effective metal conductivity to be equal to $\sigma_{\text{eff}} = 9 \cdot 10^6$ S/m, which is a very good result for a 3D-printed aluminum alloy, considering the surface roughness usually obtained in this fabrication process without additional internal surface postprocessing.

A comparison of the developed fourth-order BPF with other state-of-the-art 3D-printed DM filters is presented in Table 2. Several 3D-printed filters of order $N \leq 4$ with various relative bandwidths fabricated using different AM techniques are included in this overview. Among the compared solutions, the presented filter design exhibits the lowest insertion loss and a very high Q value. This is particularly noteworthy for a component printed directly in metal with no post-processing or tuning.

B. EIGHTH-ORDER FILTER DESIGN

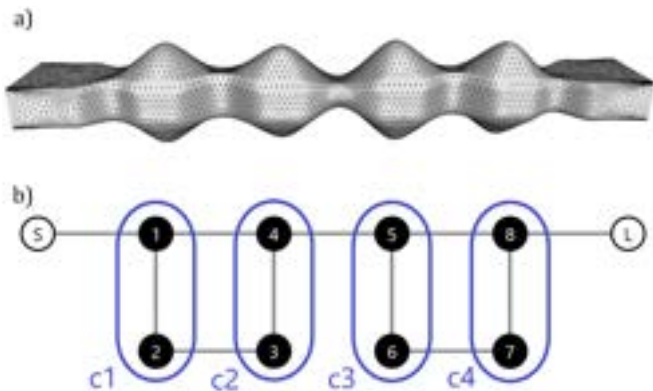
On successful completion of the fourth-order filter prototype, the development of a more challenging design, i.e. an eighth-order BPF with four TZs, was considered in the next step. The structure consists of four DM cavities cascaded one-by-one and coupled together by integrated quasi-rectangular irises, as shown in Fig. 14a. Once again, the filter is developed for 3D printing as a single block with a smooth surface

TABLE 2. Comparison with state-of-the-art 3D-printed dual-mode filters of order $N \leq 4$.

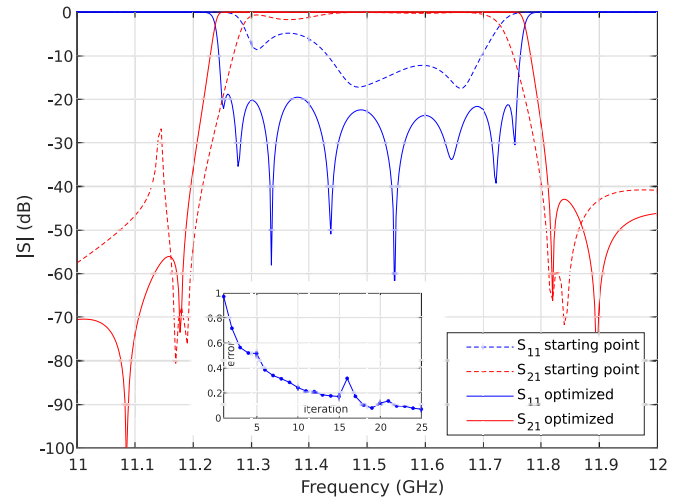
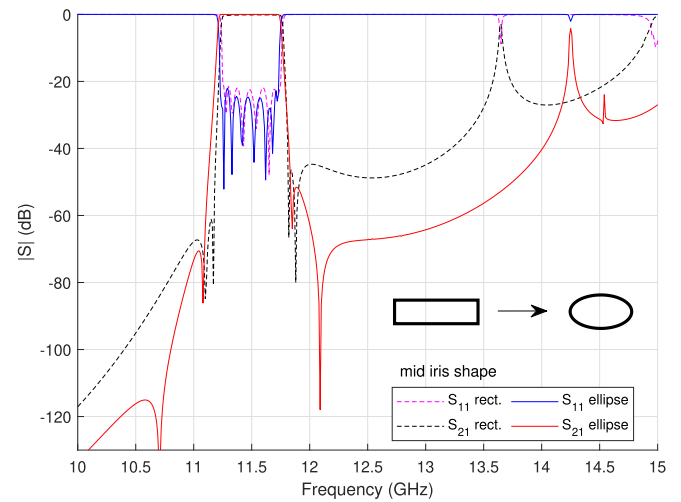
Ref.	Filter order	f_c (GHz)	FBW (%)	IL at f_c (dB)	Q	3D printing technology, material
[38]	4	10.0	3.0	0.25	2500*	SLA, copper-plated
[29]	2	10.0	2.5	0.47	500*	SLA, copper-plated
[27]	4	5.987	3.37	0.6	840	LCM, alumina
[15]	4	11.565	~ 0.47	1.0	5800	SLM, Invar, silver-plated
[15]	4	11.483	0.47	1.5	2900	SLM, Invar, silver-plated after tuning
[11]	4	10.8	0.67	1.0	2500	SLM, aluminum
T.W.	4	11.5	4.35	0.11	3500	SLM, aluminum

Ref.: reference, f_c : center frequency, FBW: fractional bandwidth, IL: insertion loss, SLA: stereolithography apparatus, LCM: lithography-based ceramic manufacturing, SLM: selective laser melting, T.W.: this work.

* Q estimated based on available filter specifications and measured IL.

**FIGURE 13.** Measurement setup using WR-75 waveguide-to-coax adapters.**FIGURE 14.** Geometry of the eighth-order DM cavity filter: a) EM model created using shape deformation, b) coupling-routing diagram.

profile, so there are no sharp corners or edges found anywhere in this design. The specifications for the new filter match

**FIGURE 15.** Optimization of the eighth-order BPF design: starting point (dashed line), optimized response (solid line). Inset: optimization error in each iteration.**FIGURE 16.** Comparison of EM simulation results of the eighth-order filter design with different mid iris shape: rectangular (dashed line) and quasi-elliptical (solid line).

those of the fourth-order design: a center frequency $f_c = 11.5$ GHz, fractional bandwidth (FBW) = 4.35%, and return loss level $RL > 20$ dB. The arrangement of cavities and irises is represented by a coupling-routing diagram in Fig. 14b, where the physical cavities are indicated by blue oval areas. To implement the chosen filter topology, the irises between cavities No. 1 and 2 and between cavities No. 3 and 4, need to support two couplings of the orthogonal mode pairs. Therefore, in their case, a quasi-square iris shape is used, as was done previously in the fourth-order filter design. In contrast, the middle iris, connecting cavities No. 2 and 3, should allow the coupling of only one pair of TE modes. For this reason, a narrow quasi-rectangular slit was implemented in the initial filter design.

The initial model was optimized using the same approach based on the zero-pole method [19], [37]. In this optimization procedure, the goal function compares the locations of zeros

and poles extracted from the filter's transfer and reflection functions with reference values of the prototype filter. There are 22 parameters used in the process: four variables describing the length of each cavity, ten variables modeling the shape of the irises (two per iris, representing its height and width), and eight parameters controlling the deformation of the cavities. The comparison of the S-parameter response of the initial and optimized design, as well as the obtained error value, can be observed in Fig. 15. The optimized result was reached in 25 iterations, and the computations performed on an Intel Core i9-10900X workstation with 64 GB RAM were completed in less than three hours.

It can be observed that the simulated response of the optimized design, shown in Fig. 16, provides a very poor upper stopband region (plotted with a dashed line). The first spurious resonance, which can be seen at the frequency of ca. 13.6 GHz, is associated with the geometry of the mid iris, and can be controlled by its shape and width. Therefore, the first design was modified so that the out-of-band rejection level could be improved. An improvement was obtained by optimizing the shape of the mid iris into a rounded, elliptical shape, which is shown in Fig. 16 (solid line). The spurious resonance was shifted up to ca. 14.25 GHz, and the rejection level was improved to around 60 dB under 13.25 GHz and around 40 dB up to 14 GHz. In addition, the center frequency in the revised design was shifted down by 20 MHz to $f_c = 11.48$ GHz to anticipate the manufacturing impact on the dimensions of the filter. In this design, the positions of the TZs can also be modified: they can be placed closer to or farther from the passband with little effect on the selected bandwidth. The solution chosen in this study was to place one pair of TZs close to the passband—at 11.1 and 11.85 GHz—and the other pair farther away—at 10.75 GHz and 12.1 GHz—to improve the upper stopband rejection level. Although the TZs are positioned almost symmetrically on either side of the bandwidth, the attenuation of the lower stopband is noticeably higher. This is due to the proximity of the waveguide cut-off frequency, which affects the rejection of the filter below the passband. The designed structure was also analyzed for multipactor effects in Ansys HFSS. The analysis reported no occurrence of multipaction up to a power threshold of 10 kW, which confirmed the validity of using such a solution for high-power applications.

The eighth-order filter design was prepared for fabrication and 3D-printed by Northern Waves AB. The fabricated component is presented in Fig. 17a. The filter was manufactured in one piece using SLM, with powder layer thickness of 30 μm . The structure was printed along its greatest dimension, from one flange to the other. The external dimensions of the printed component are $145.5 \times 38 \times 38$ mm. Physical polishing by sandblasting was applied in post-processing. According to the manufacturer, the process should reduce the surface roughness (Ra) from 8–10 μm as built, down to 4–6 μm . The prototype was measured using the same setup from Fig. 13 and the results were compared with EM simulation. The wideband response of the filter

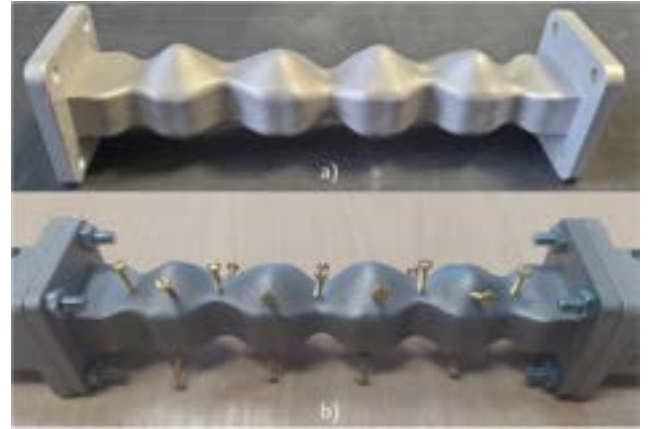


FIGURE 17. Photographs of the eighth-order DM cavity filter: a) 3D-printed filter prototype, b) measured prototype with tuning screws.

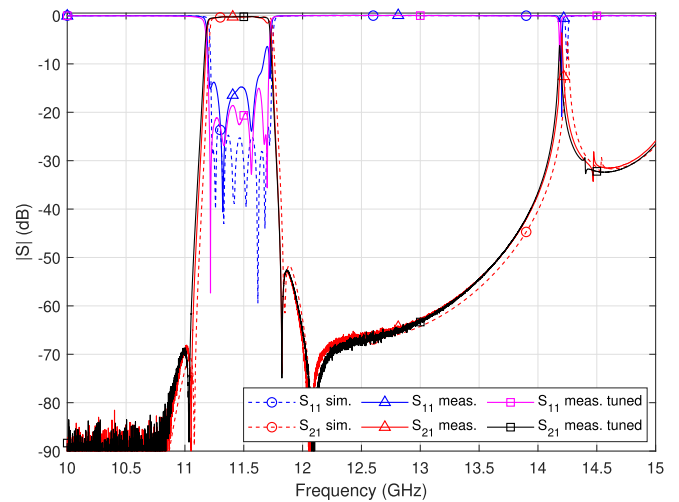


FIGURE 18. Comparison of measurement results (solid line) and EM simulation (dashed line) of the fabricated eighth-order filter with four TZs—wideband response.

is shown in Fig. 18, whereas a close-up of the passband is presented in Fig. 19. There is good agreement with the simulation, although the measured response is shifted ca. 20 MHz below the specified band, and the prototype filter is not fully matched: the return loss is generally better than 15 dB over the passband, apart from the mismatch at the higher edge of the band. The locations of the TZs and the out-of-band response precisely represent the anticipated response. The TZ at the lowest frequency is not visible due to the limited dynamic range of the VNA. Because of the initially unsatisfactory return loss level, tuning screws were incorporated into the cavities and irises, as shown in Fig. 17b, and manual tuning was performed to compensate for manufacturing imperfections, thereby improving the overall in-band RL. A total of 16 tuning screws were used: a pair for each cavity and inter-cavity coupling iris, and one per each I/O iris. The tuned filter response is also shown in Fig. 18 and Fig. 19. The passband insertion loss can be observed in Fig. 20, where the measured response before and after

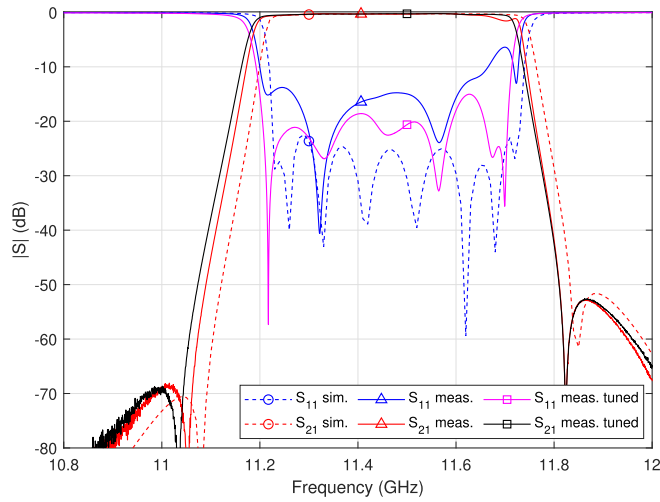


FIGURE 19. Comparison of EM simulation and measurement results of the fabricated eighth-order filter with four TZs—narrowband response.

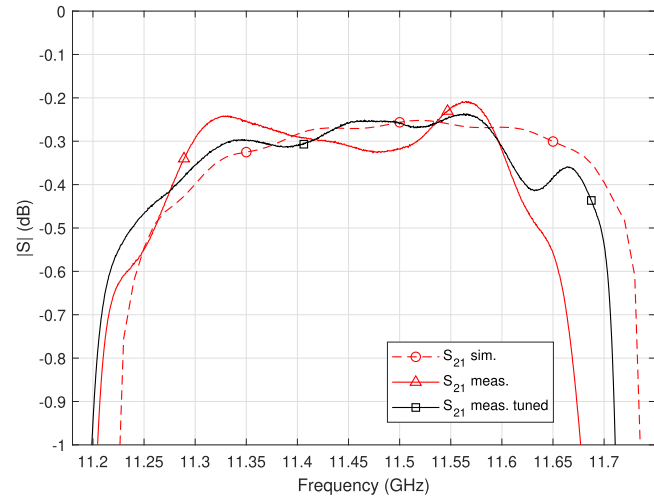


FIGURE 20. Comparison of EM simulation and measurement results of the fabricated eighth-order filter with four TZs—insertion loss.

tuning is compared with the simulation, including conductor losses of the 3D-printed metal surfaces. The IL at f_c is equal to 0.25 dB and spans between 0.25 and 0.55 dB over the passband. This values closely match the simulated effective conductivity of $\sigma_{\text{eff}} = 1 \cdot 10^7$ S/m. The Q of the fabricated filter is estimated as 4500, which is a very satisfactory result for direct metal AM with aluminum alloy. It is also worth noting that the total mass of the component is only 75 g, which is considered a potentially small payload for an eighth-order waveguide filter in the X or Ku-band [38], [39], [40].

A comparison of the proposed eighth-order filter with other 3D-printed filter solutions is presented in Table 3. Our design is listed along with three other BPFs of order $N > 4$ based on DM resonators, with comparable operating frequency and similar or greater FBW. If the Q was not available in the publication, its value was estimated based on the reported IL. Of all the compared solutions, the filter proposed in this work matches the lowest absolute insertion loss, and delivers

TABLE 3. Comparison With State-of-the-art 3D-printed Dual-Mode Filters of Order $N > 4$.

Ref.	Filter order	f_c (GHz)	FBW (%)	IL at f_c (dB)	Q	3D printing technology, material
[41]	6	10.9	7.3	0.7	2000	FDM, copper-coated
[40]	8	8.23	5.47	0.32–0.68	2800*	SLA, copper-plated
[31]	8	8.23	5.59	0.25	5700	SLM, aluminum-copper alloy
T.W.	8	11.46	4.35	0.25	4500	SLM, aluminum

Ref.: reference, f_c : center frequency, FBW: fractional bandwidth, IL: insertion loss, FDM: fused deposition modeling, SLA: stereolithography apparatus, SLM: selective laser melting, T.W.: this work.

* Q estimated based on available filter specifications and measured IL.

a high Q -factor. It has to be noted that a better Q is achieved in [31] by using an aluminum-copper alloy for 3D-printing. Nonetheless, the design presented here represents a viable solution for low-loss, high-power applications.

C. DISCUSSION

The components presented in this article were additively manufactured using the standard SLM process with 30–50 μm resolution, which yielded very satisfactory results for the fourth-order filter model. However, when examining the eighth-order filter prototype, it is clear that the initial measured response (without tuning) did not meet the specifications for in-band return loss. The second design was treated with sandblasting to reduce surface roughness, which in turn decreased the insertion loss and made it possible to obtain a higher Q -factor. However, in order to improve the manufacturing accuracy of the presented prototype and, in consequence, the filter response, more precise 3D-printing methods can be used, such as high-precision micro laser sintering (MLS) with 7 μm layer thickness [42], although this is a considerably slower and more expensive process. Regarding the obtained effective metal conductivity, the estimated value of the 3D-printed aluminum alloy, $\sigma_{\text{eff}} = 1 \cdot 10^7$ S/m, is relatively high compared to the conductivity of ideal aluminum, which is $\sigma_{\text{Alu}} = 3.8 \cdot 10^7$ S/m. However, there are other AM processes reported in the literature involving silver or gold plating of the surfaces, such as [43] and [44], that declare an even more efficient conductivity of $2.5 \cdot 10^7$ S/m. Nonetheless, the fabricated components reported in this paper represent high quality of direct metal 3D printing and validate the design approach based on deformed DM cavities.

IV. CONCLUSION

This article presented a new type of waveguide filter based on deformed dual-mode cavities, designed specifically for 3D printing. The proposed designs were developed using shape deformation and thus benefit from a complex geometry and smooth surface profile, which can only be fabricated by additive manufacturing processes. Fourth-order and eighth-order filters with transmission zeros were

presented and carefully analyzed in this paper. Both filter designs were experimentally verified through prototypes manufactured using the selective laser melting process. The performed measurements validate the design idea and show excellent results compared to the state-of-the-art solutions. The two reported structures exhibit low insertion loss, relatively high Q and a smooth profile, making them potentially useful for low-loss, high-power applications. Although the obtained effective metal conductivity was considered a very good result, further development of additive manufacturing of microwave components along with post-processing techniques, such as polishing [45], or plating [44], should bring even more satisfactory results in the near future.

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4.4. A 3-D-Printing-Oriented Coaxial-Line Filter with Wide Out-of-Band Rejection

The next article included in this thesis, published in *IEEE Microwave and Wireless Technology Letters*, focused on the application of shape deformation and additive manufacturing to design and fabricate novel, air-filled coaxial line components. This research was carried out in close cooperation with the Microwave Components Group of Public University of Navarre (UPNA), Spain. The aim of this work was to develop novel microwave passive hardware achieving superior RF performance, intended for practical, real-life applications such as satellite communication (SATCOM) systems, drawing from the combined advantages of coaxial-line technology, AM and, in particular, the capabilities of free-form modeling. For this reason, the scenario chosen for investigation was a SATCOM output filter for downlink data transmission employed in Earth Observation systems such as Sentinel-1 [114], that requires a very wide stopband rejection and operates with moderately high power signals—because a compact solution based on coaxial line could bring significant improvement in this area. Coaxial lines offer low transmission loss and fairly good power-handling capability brought in compact dimensions, and as such can be considered an attractive alternative to voluminous waveguides, where moderate to high power is considered. Further potential lies in AM-oriented design, which can benefit from non-standard profiles and complex, intricate models, that are now feasible to produce. These directions were explored throughout this study, leading to an innovative filter design that is completely adapted to 3D printing.

The groundwork for this work was laid by the research team from UPNA, who had first examined the possibility of 3D-printing an air-filled coaxial line structure as a single mechanical part made directly in metal [115]. That design incorporates a stepped-impedance lowpass filter structure, integrated with a short bandpass filter section based on quarter-wavelength short-circuited stubs. The LPF provides the main filtering functionality, canceling out the unwanted signals in the upper stopband, whereas the purpose of the BPF is to select the passband and, most of all, to grant the mechanical stability required for monoblock fabrication by AM. Although that solution was well described in [115] and is not the focus of this thesis, the idea to develop a filter as a combination of two filtering structures with different functions was followed in the novel smooth-profile design developed later on. Therefore, the description of the new coaxial-line filter, presented hereafter, has been divided into separate sections (BPF and LPF), each containing a characterization of its specific role as well as detailed design considerations of that circuit.

Unlike its stepped-impedance predecessor, the LPF design presented in [MB5] exhibits a smooth surface profile, optimum for AM, incorporating varying diameters of both the inner and outer coaxial conductors. This configuration enables the realization of the exceptionally wide rejection band (up to $4 \cdot f_0$) necessary for the targeted space application. The smoothly-varying coaxial line profile is first obtained through an inverse scattering synthesis method developed by UPNA scholars [116] and further optimized using shape deformation techniques, as described in Chapter 3.

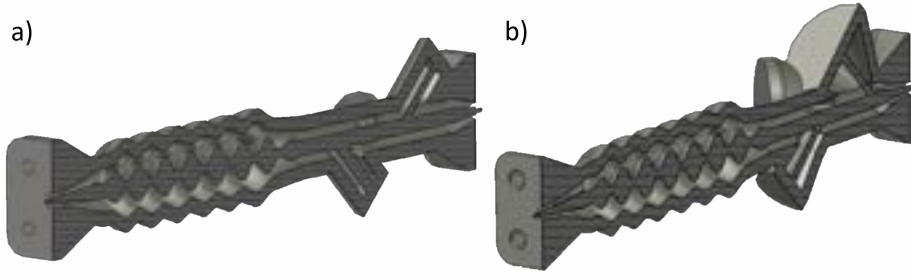


Fig. 4.25: Geometry of the proposed coaxial-line filters (cross-sectional view): a) design with linear stubs, b) design with radial stubs.

4.4.1. Bandpass filter design

The design of the smooth-profile coaxial-line filter consists of two parts. The first one is a short-circuited parallel stub BPF, which facilitates the fabrication of the whole filter in a single piece, while also selecting the passband for the complete device. This part was designed separately, following the design equations given in [117–119], and was later connected with the smooth-profile lowpass section, resulting in the final filter structure shown in Fig. 4.25 (two variants are displayed). The main purpose of the bandpass part is to provide the necessary mechanical stability of the device to allow it being 3D-printed in one piece, i.e., to support the inner coaxial conductor. This condition requires physical contacts (short circuit) between the inner and outer conductors, which can be fulfilled by an electrical circuit composed of parallel shorted stubs, each one a quarter wavelength ($\lambda/4$) in length, separated by sections of transmission line, also $\lambda/4$ long. Apart from providing mechanical stabilization, the shorted stubs also improve the near-band rejection by introducing transmission zeros into the filter response above the passband. The equivalent circuit

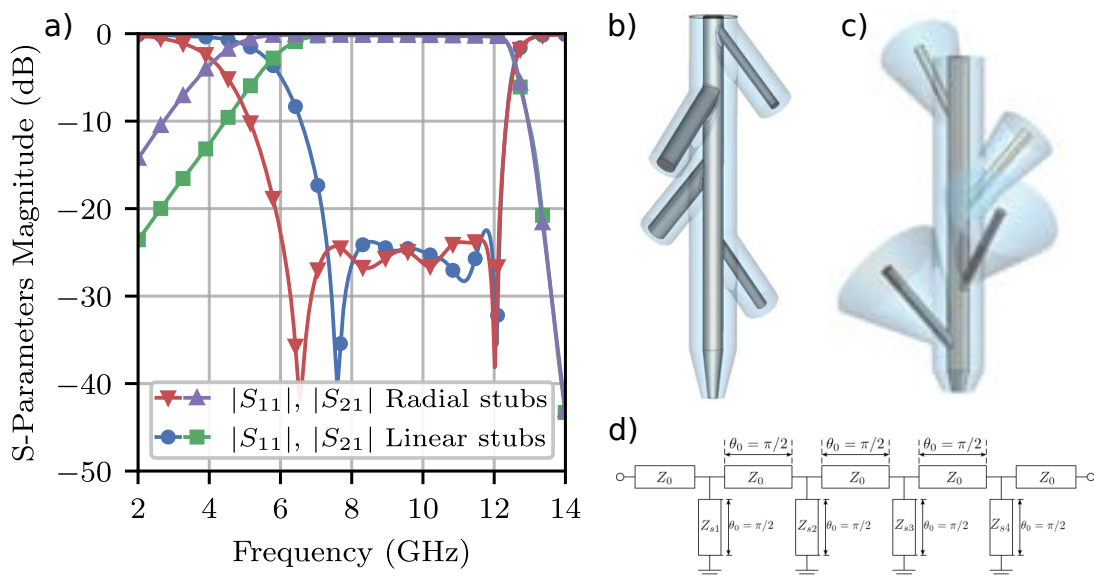


Fig. 4.26: Design of the parallel stub BPF: a) comparison of EM simulation for both BPF designs: linear stubs and radial stubs, b) linear-stub BPF, c) radial-stub BPF, d) equivalent circuit.

for such a BPF structure is given in Fig. 4.26d. The design of this filter section can be carried out following the procedure described in [118], where the design equations to dimension the physical stub lengths and characteristic impedances assume a constant impedance between the input and output port. The equations for the stub impedances ($Z_{s1}, Z_{s2}, Z_{s3}, Z_{s4}$ in Fig. 4.26d) depend on the frequency specifications. Within the framework of this study, two BPF variants were designed and tested: an X-band filter based on linear stubs (Fig. 4.26b) and an ultra wideband filter with radial, conical stubs (Fig. 4.26c). For the first variant, the filter characteristics are: order $N = 4$, Chebyshev response, $RL = 25$ dB, central frequency $f_0 = 10$ GHz, and a bandwidth covering the whole X-band (8–12 GHz, 40% FBW). Thanks to the versatile capabilities of AM, a new conical stub shape could be introduced to profoundly enhance the bandwidth of the filter (FBW reaching 80%). Because of the electromagnetic effects present in the actual coaxial line, in contrast with the ideal transmission line model, an optimization of both designs was carried out in order to obtain the desired frequency response in a full-wave EM simulation. The comparison of passband responses offered by both BPF models can be inspected in Fig. 4.26a. Note that in both cases, the achieved filter bandwidth is slightly larger than the initial requirement, leading to an upper cutoff frequency of about 12.5 GHz—i.e., not overlapping with the LPF cutoff, $f_c = 12$ GHz. This is designed intentionally, to suppress some unwanted resonances which may appear when connecting the lowpass and bandpass sections together. It should also be noted that the chosen BPF design based on $\lambda/4$ stubs allows only for the realization of relatively wideband structures. For a narrower bandwidth, a different filter design should be considered.

4.4.2. Lowpass filter design

The second and central part of this work concerns the development of a smooth-profile lowpass filtering structure responsible for providing a wide rejection band for the whole prototype. The design procedure began with a synthesis of the initial LPF model that implements the selected Chebyshev transfer function. The chosen filtering characteristic assumes a lowpass cutoff frequency $f_c = 12$ GHz, return loss of 30 dB and order $N = 15$. The frequency of maximum rejection was fixed to $f_r = 27$ GHz, resulting in a lowpass prototype filtering function that achieves a wide rejection band, i.e., whose first replica appears beyond 40 GHz. This last characteristic is important in order to ensure a very wide stopband behavior required for the intended SATCOM application.

The lowpass filter synthesis was executed using an inverse scattering method named Integral Layer Peeling (ILP), developed by the researchers in UPNA [116]. First, the target filtering response needs to be obtained from a commensurate-line unit element (UE) prototype that implements the selected Chebyshev transfer function. The UE characteristic impedances can be calculated by applying the techniques detailed in [120]. After the commensurate-line UE prototype satisfying the required frequency response has been obtained, its impulse response is computed and properly interpolated to get the target response for the synthesis method. The S-Parameter response corresponding to the UE model is depicted in Fig. 4.27a. Then, the ILP method can be

applied to synthesize the required coupling coefficient $K(z)$ for the smooth-profiled LPF, as detailed in [116]. The resulting $K(z)$, corresponding to the target response, is depicted in Fig. 4.27b. Note that the obtained $K(z)$ is a continuous and smoothly varying function of z . The characteristic impedance, as a function of the propagation axis z , is related to the coupling coefficient $K(z)$ through the equation [121]:

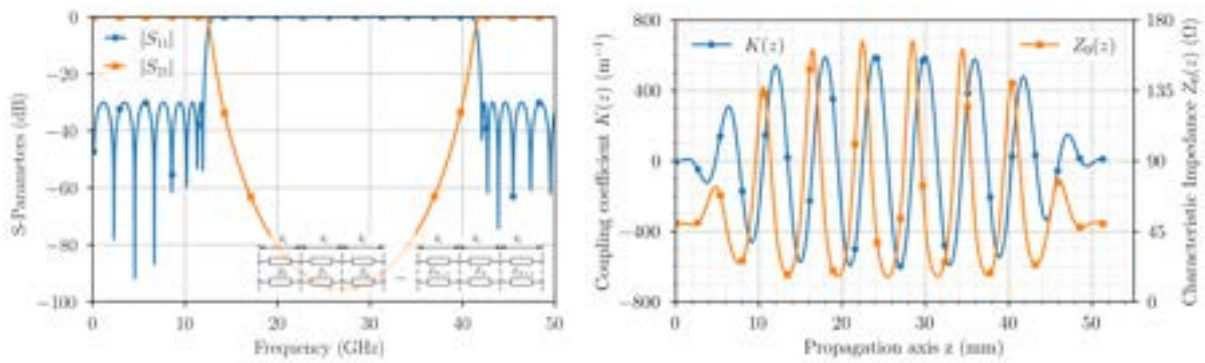
$$Z_0(z) = Z_0(0) \cdot e^{-2 \int_0^z K(p) \cdot dp}, \quad (4.3)$$

where $Z_0(0)$ is the characteristic impedance at the input port and p is a dummy variable of integration. The obtained characteristic impedance profile is also included in Fig. 4.27b. Finally, the coaxial inner (d) and outer (D) diameters can be calculated from the synthesized $Z_0(z)$ profile using the analytic equation for the fundamental TEM mode of the coaxial transmission line given in [122]:

$$Z_0 = \sqrt{\frac{\mu}{\varepsilon}} \frac{\ln(D/d)}{2\pi}, \quad (4.4)$$

where $\varepsilon = \varepsilon_0 \varepsilon_r$ is the dielectric permittivity of the material filling the coaxial line. The application of smoothly varying inner and outer diameters, as opposed to the classic stepped-impedance design, allows one to increase the minimum physical gap between the conductors (for high-power operation) and to obtain more extreme $K(z)$ coefficients for wider rejected bands.

Even though the synthesized design realizes the desired lowpass Chebyshev response when single-mode operation is assumed, a significant deviation in the filter response occurs when a full-wave EM simulation is performed, as shown in Fig. 4.28 (dashed lines). This is caused by spurious coupling to higher-order modes in the structure, which are not considered in the inverse scattering synthesis. For this reason, the synthesized coaxial-line model needs to be further processed. To this end, an efficient design optimization utilizing the free-form deformation technique, which has been proposed in this thesis, was employed to reach the targeted frequency response, with an emphasis on improving the out-of-band performance. The optimization process involved managing the lengths of each monotonously varying impedance section as well as the control points modifying the conductor diameters in each section (see: inset in Fig. 4.28). In this design, both the



a) Target response obtained from the commensurate-line UE prototype, with equivalent circuit (inset). b) Synthesized coupling coefficient $K(z)$, and characteristic impedance $Z_0(z)$, as a function of the propagation axis z .

Fig. 4.27: Smooth-profile LPF synthesis using the inverse scattering method ILP [116].

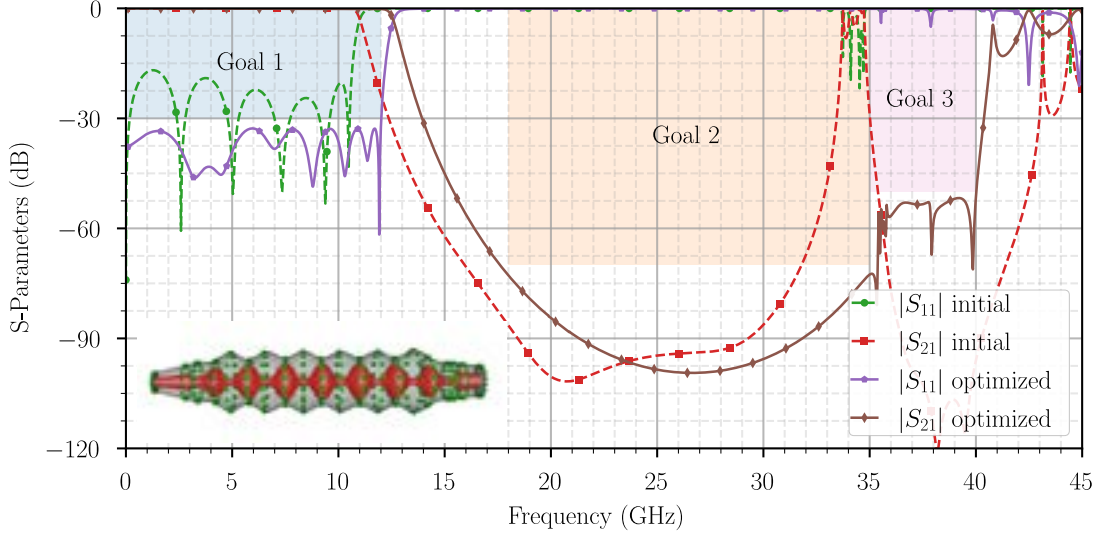


Fig. 4.28: S-parameter response (full-wave EM simulation) of the smooth-profile coaxial LPF. Synthesized design (starting point): dashed line, optimized design: solid line. Inset: Optimized LPF model with control points.

inner and outer conductor surfaces were modified, which allowed for obtaining greater impedance contrasts while keeping a fixed minimum inner conductor diameter of approximately 1 mm, which was required for mechanical stability (monoblock 3D printing). Several optimization goals were employed to perform the task, as shown in Fig. 4.28. For the LPF passband, Goal 1 imposes a RL level of at least 30 dB up to $f_c = 12$ GHz. To accomplish the out-of-band requirements, Goals 2 and 3 constrain the transmission level between 18 and 40 GHz. Additional optimization constraints were imposed on the inner and outer conductors at the input and output interfaces, to match the dimensions of the BPF section at one end, and the 2.4 mm coaxial connector at the other port. As shown in Fig. 4.28 (solid lines), the optimized LPF design fulfilled all the imposed optimization goals: a recovered passband with $RL > 30$ dB, and a stopband extending up to 40 GHz.

4.4.3. Combined filter design, fabrication & measurement

In Fig. 4.29, the S-Parameter responses of the bandpass and lowpass filters, simulated in isolation, are compared to the final response of the assembled filter. It can be observed that the final passband of the combined filter is imposed by the parallel-stub structure, as the S_{21} in the frequency range up to 12 GHz resembles that of the BPF structure. Also, the resulting in-band RL level is similar to that of the BPF, because of its original RL being lower than that of the LPF. On the other hand, the stopband is mostly dictated by the rejection of the smooth-profile lowpass structure, which is enough to ensure the rejection masks described in the paper (rejection better than -70 dB between 18 and 35 GHz, and better than -50 dB between 35 and 40 GHz). It can be observed how, thanks to the out-of-band performance of the LPF, the poorer spurious behavior of the parallel-stub filter is effectively canceled by the cascaded smooth-profiled section. As already mentioned, here it can be noticed that the upper cutoff frequency of the BPF is not exactly equal to

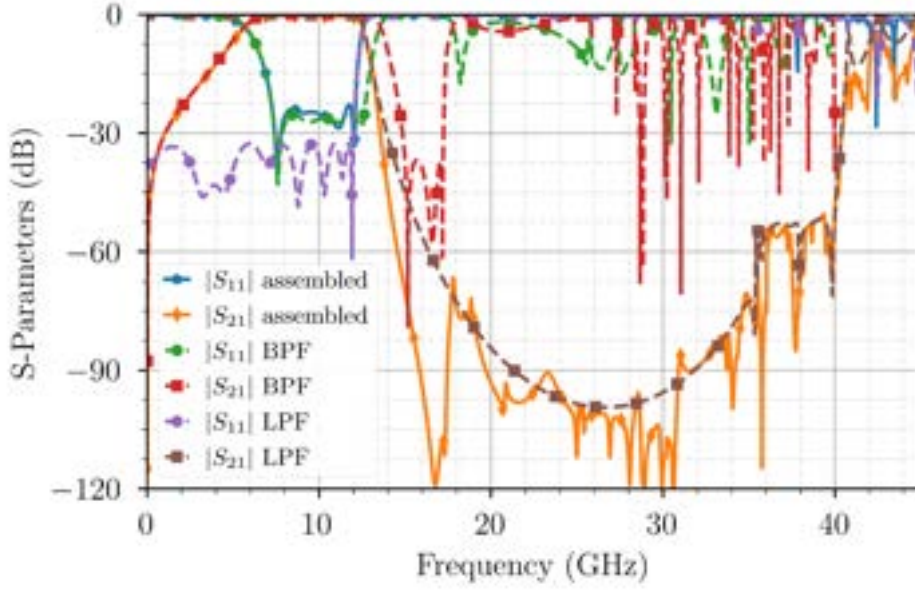
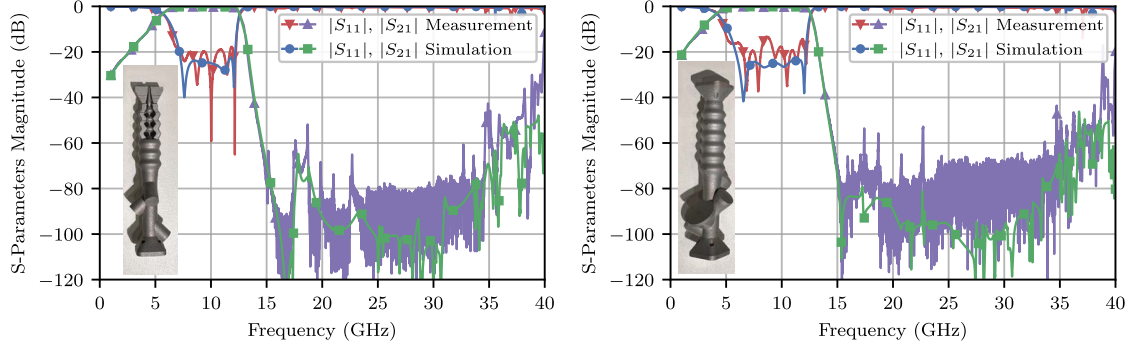


Fig. 4.29: S-parameter response (full-wave EM simulation) of the smooth-profile coaxial filter: BPF (linear stubs), LPF, and combined filter.

that of the LPF (12 GHz); This is intentionally designed as such, as it helps minimize the resonances appearing due to the connection of the two filtering sections.

Two single-piece prototypes composed of the described bandpass and lowpass filtering sections were 3D-printed in metal using SLM by Northern Waves AB [101]. The difference between the two models lies in the design of the BPF part, as described in section 4.4.1. The metal powder used in the fabrication process was an aluminum-silicon-based alloy (AlSi10Mg), with an expected effective conductivity of $\sigma = 1.67 \cdot 10^7$ S/m and an average surface roughness of $3 \mu\text{m}$, according to the manufacturer. Each layer of powder has a typical thickness between $30\text{--}50 \mu\text{m}$. The expected dimensional tolerances at each layer are within $\pm 50 \mu\text{m}$. No post-processing was applied internally, only the surfaces corresponding to the flanges were machined to ensure good contact with the coaxial connectors. A 2.4 mm coaxial connector was employed at each port to enable connection with the VNA for measurement. The results of the measurements are compared with EM simulation of the complete filters in Fig. 4.30a (linear-stub BPF) and Fig. 4.30b (radial-stub BPF). Both measured prototypes exhibit a response which is in very good agreement with the simulations. The in-band RL level is slightly worse compared to the simulation result, especially for the radial stub filter. However, considering a passband RL of approximately 15 dB, the bandwidth achieved by the radial stub design spans from 5.2 to 12.3 GHz, resulting in a FBW reaching 88%. The measured insertion loss at f_0 is ≈ 0.37 dB for the linear stub filter and ≈ 0.52 dB for the radial one. Most importantly, for both manufactured prototypes a rejection level better than 60 dB is achieved between 15 and 34.55 GHz, and better than 40 dB for the higher stopband frequencies between 34.55 GHz and 38.5 GHz. In order to analyze the effect of possible manufacturing deviations and their impact on the final filter performance, a series of simulations (i.e., tolerance analysis) were conducted considering uniform variations along the structure of the inner and outer



a) Response comparison for the prototype with linear stubs. Inset: photograph of the prototype, cut open to show the interior. b) Response comparison for the prototype with radial stubs. Inset: photograph of the manufactured prototype.

Fig. 4.30: S-parameter response of the measured coaxial-line filter prototypes compared with EM simulation.

conductors. To perform this study, variations of $\pm 50 \mu\text{m}$ were taken into account, as stated by the manufacturer. For the inner conductor, the dimensional variation was applied to all the BPF stubs, the length of the inner conductor in the propagation direction, and to the inner conductor of the LPF as well. Similarly, the deviations were applied to all the outer conductors in the stubs, and along the propagation direction for the BPF and the LPF. The results corresponding to this analysis are presented in Fig. 4.31, where the baseline simulated response is plotted in orange and blue colors, and the results corresponding to the deviated geometries are represented with gray and black colors for the S_{21} and S_{11} parameters, respectively. On top of these simulation results, the measured filter response (for the variant with linear stubs) is overlaid in red and green dashed lines for the S_{21} and S_{11} , respectively. Inferring from the presented sensitivity analysis, it can be concluded that the observed small deviations between measurement and baseline simulation lie within the expected bounds, considering the $\pm 50 \mu\text{m}$ tolerances specified by the manufacturer.

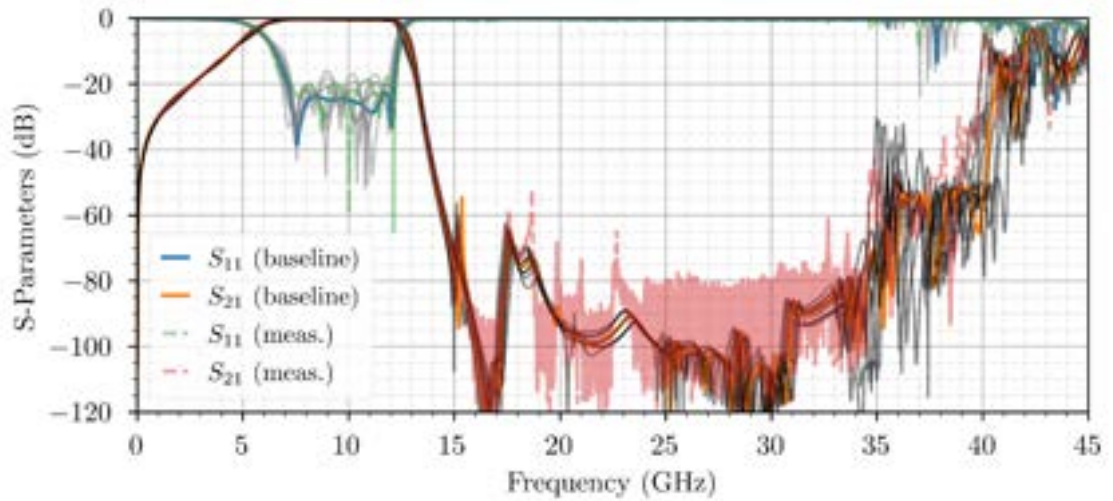


Fig. 4.31: Results of the sensitivity analysis performed for the smooth-profile coaxial-line filter design (the one with linear stubs), including the measurements taken from the manufactured prototype.

Ultimately, the results presented in Fig. 4.30 reveal that both fabricated components satisfy the stopband requirements demanded by current SATCOM systems. The combination of inverse scattering synthesis with shape deformation of the smoothly-varying LPF conductors has led to a significant improvement in the stopband performance of the new designs.

**[MB5] A 3-D-Printing-Oriented Coaxial-Line Filter
With Wide Out-of-Band Rejection**

M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Álvarez-Botero, A. Lamecki,
M. A. G. Laso, and P. Martin-Iglesias

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Author's contribution:

Optimization of the smooth-profile lowpass filter design using shape deformation techniques, participation in the design of the bandpass filter with short-circuited stubs and the combined filter design presented in the paper, preparation of the 3D models for additive manufacturing, preparation and corrections of the manuscript.

A 3D-Printing-Oriented Coaxial-Line Filter with Wide Out-of-Band Rejection

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Abstract—In this letter, a novel design for a 3D-printed, self-supported coaxial-line X-band filter is presented. The filter is intended for Earth Observation (EO) data downlink systems, where it must effectively reject signals in a wide frequency range. The filter design incorporates a 15th-order low-pass filter structure with smooth profile, integrated with a short bandpass section with four $\lambda/4$ short-circuited stubs. The optimization of the low-pass section is attained by means of shape deformation, including the inner and outer coaxial conductors, and leads to a wide rejection band up to around 40 GHz, to suppress the third harmonic and other undesired out-of-band frequencies. A prototype was fabricated in one piece in aluminum alloy using selective laser melting (SLM) and measured, exhibiting excellent agreement with simulations. In terms of out-of-band performance, the proposed coaxial-line filter is superior to other related state-of-the-art solutions.

Index Terms—additive manufacturing, coaxial-line filter, inverse scattering, out-of-band rejection, shape deformation, smooth-profile filter.

I. INTRODUCTION

Coaxial-line technology is very attractive for RF systems due to its balance of compactness, low loss, and power-handling capability, making it a good intermediate solution between bulky waveguides and lossy and power-limited planar technologies. These qualities make it suitable for highly integrated RF modules, such as those in satellite communications. As RF system requirements drive increased geometric complexity to enhance performance, conventional manufacturing techniques like milling often become unsuitable.

Additive manufacturing (AM) offers a potential solution to these limitations. Over the past decade, AM has emerged as a promising method for producing passive microwave components such as filters and antennas [1]–[6]. The ability to create complex structures offered by AM has enabled new designs for microwave components that are not feasible with traditional methods, as demonstrated in the literature [7]–[10].

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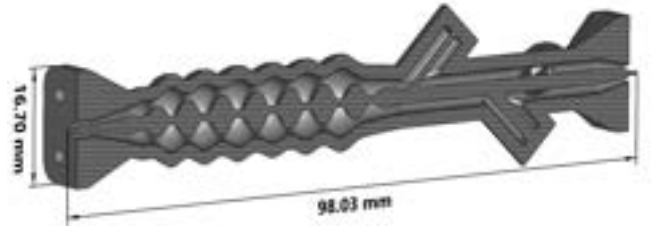


Fig. 1. Geometry of the proposed coaxial-line filter (cross-sectional view).

In this letter, we present a fully AM-oriented coaxial-line bandpass filter (BPF) with very wide rejection band, whose geometry is shown in Fig. 1. This coaxial filter is designed for the X-band Payload Data Handling & Transmission (PDHT) module, commonly employed in Earth Observation (EO) data downlink systems such as Sentinel-1 [11]. The primary purpose of this output filter is to avoid the radiation of unwanted signals. A particularly critical application arises when an X-band PDHT module is integrated with an instrument (e.g., a radiometer or radar) that utilizes Ka-band receivers (e.g., SKIM, a former EE09 candidate, operating at approximately 35.75 GHz [12]). In such scenarios, the X-band PDHT module must effectively reject signals within the receiver band of the collocated instrument. In order to accomplish the stringent rejection requirements, a low-pass Chebyshev transfer function with cut-off frequency $f_c = 12$ GHz and an upper stopband beyond its third harmonic is implemented. The filter features a smooth profile, optimum for AM, synthesized through inverse scattering techniques [13] and further optimized using shape deformation [14] to improve wideband rejection. As opposed to a classical design with stepped-impedance coaxial-lines [15], the proposed design employs varying diameters for both inner and outer conductors, fully exploiting the geometrical capabilities of AM, allowing it to achieve the very wide rejection band required for the intended space application.

II. FILTER DESIGN AND OPTIMIZATION

The design of the wide rejection band filter is composed of two sections. The first one is a bandpass filter with $\lambda/4$ short-circuited stubs that provides the necessary mechanical support for a dielectric-less all-metal selective laser melting (SLM) manufacturing and the required passband for the device. The second section is a new type of smooth-profiled low-pass filter (LPF), which provides the required rejection band for our PDHT application. Indeed, one of the main novelties of this work resides in the design and optimization of the new LPF to achieve a wide rejection band by using inverse scattering and shape deformation techniques, meeting the needs of an EO satellite. In contrast to the previous design from [15], the

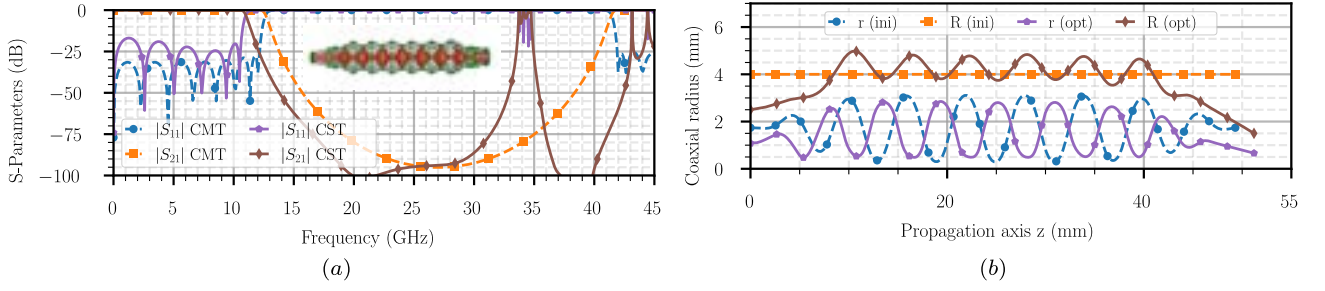


Fig. 2. (a) Comparison between the S-parameter responses of the synthesized LPF with CMT (single-mode) and the CST full-wave analysis (multi-mode). Inset: Optimization setup of the low-pass filter. Control points and displacements are shown by green points and arrows. (b) Computed inner (r) and outer (R) coaxial radii profiles, corresponding to the initially synthesized filter with ILP (dashed lines) and the optimized geometry with InventSim (solid lines).

aim is to enlarge the rejected band as much as possible, while maintaining a well-shaped passband located at X-band.

In order to accomplish the requested frequency specifications, a LPF with return loss $RL = 30$ dB, cut-off frequency $f_c = 12$ GHz, frequency of maximum rejection $f_r = 27$ GHz, and order $N = 15$ is designed. The smooth-profiled LPF design begins by obtaining the target response for the inverse scattering synthesis method from a commensurate-line unit-element (UE) prototype that implements the selected Chebyshev transfer function [13]. The UE characteristic impedances can be calculated by applying the techniques detailed in [13], [16]. Since the coaxial-line is air-filled, a unitary relative permittivity is considered. After the commensurate-line UE prototype satisfying the required frequency response has been obtained, its impulse response is computed and properly interpolated to get the target response for the Integral Layer Peeling (ILP) method which synthesizes the required coupling coefficient $K(z)$ for the smooth-profiled LPF, as detailed in [13]. Finally, the characteristic impedance profile $Z_0(z)$ is obtained from $K(z)$ as described in [17]. The synthesized $Z_0(z)$ profile is related to the coaxial inner (r) and outer (R) radii through the analytic equation for the fundamental TEM mode of the coaxial transmission line as [18]

$$Z_0 = \sqrt{\frac{\mu}{\varepsilon}} \frac{\ln(R/r)}{2\pi}, \quad (1)$$

where $\varepsilon = \varepsilon_0 \varepsilon_r$ is the dielectric permittivity of the material filling the coaxial line ($\varepsilon_r = 1$ in our case). At this point, the manufacturing constraints need to be considered for the optimal selection of the r and R values. Since the minimum thickness that can be printed using SLM is 1 mm in our case, the minimum inner conductor radius value must be fixed to $r_{min} = 0.5$ mm. In order to satisfy this fabrication requirement, a constant $R = 4$ mm along the propagation axis z is chosen. Therefore, using (1) to calculate the inner conductor radii of the LPF, the geometry of the coaxial-line filter corresponding to the synthesized $Z_0(z)$ is depicted in Fig. 2b (dashed lines, labeled $r(\text{ini})$ and $R(\text{ini})$).

In Fig. 2a, the frequency response corresponding to the synthesized profile is analyzed by means of single-mode coupled-mode theory (CMT) [19], verifying that the desired low-pass Chebyshev response is achieved when single-mode operation is assumed (dashed lines). On the contrary, a fullwave simulation with CST Microwave Studio shows that a significant deviation occurs in the filter response (solid lines), which is

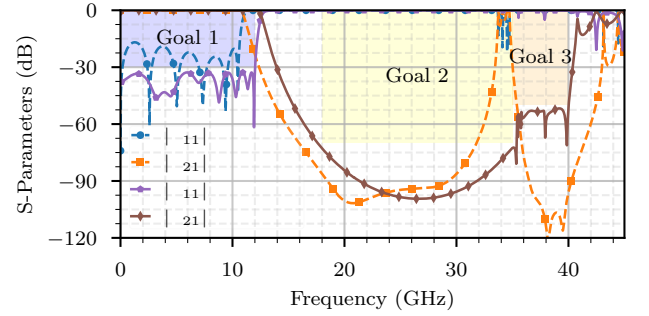


Fig. 3. S-parameter response (EM simulation) of the smooth-profile coaxial LPF: starting point (dashed line), optimized design (solid line).

basically due to the effect of higher-order modes inside the structure. Therefore, we employ a shape deformation technique [14] implemented in InventSim electromagnetic field solver [20], [21] to recover the original low-pass frequency response and, at the same time, improve the out-of-band performance.

The shape deformation technique is based on a definition of a set of points controlling the deformation of the 3D model. The operation is performed by moving these control points to certain new coordinates, and the original structure is then smoothly reshaped according to this displacement. InventSim software enables efficient optimization of the control points' positions, with a goal function targeting the required return loss level in the passband and wideband rejection.

In this design, both the inner and outer conductor surfaces were modified in the optimization process. The non-uniform profile of the outer conductor enables achieving greater impedance contrasts while keeping a fixed minimum inner conductor diameter of approximately 1 mm, for mechanical stability. The design optimization involved managing the geometrical variables (the lengths of each monotonously varying impedance section) as well as the deformation parameters, i.e., the control points determining the radii of the conductors in each section, as shown in Fig. 2a (inset).

Several optimization goals are employed to achieve the filter requirements, as indicated in Fig. 3. For the LPF passband, Goal 1 imposes a RL level of at least 30 dB up to $f_c = 12$ GHz. To accomplish the out-of-band requirements, Goal 2 constrains the transmission level below -70 dB between 18 and 35 GHz, while Goal 3 provides suppression of at least -50 dB between 35 and 40 GHz. Additional optimization constraints were imposed on the radii of the inner and outer conductors at

the input and output (I/O) interfaces, to match the geometry of the BPF section at the input and the 2.4mm coaxial connector dimensions at the output port. Axial symmetry of the structure was assumed during design optimization.

The optimized profile, plotted in Fig. 2b (solid lines), fulfilled all the imposed optimization goals. The comparison of the S-parameter response of the smooth-profiled LPF before and after design optimization is shown in Fig. 3. As it can be seen, by applying the shape deformation method, exploiting also the variation of the outer conductor, we managed to achieve the desired filter frequency response, surpassing the negative effects of the higher-order modes.

The second part of the proposed coaxial-line filter design is the bandpass section. The main purpose of this part is to provide the necessary mechanical support of the device to allow it being additively manufactured in one piece. This requirement necessitates mechanical contacts between the inner and outer conductors. Apart from providing mechanical stabilization, the BPF can also be used to improve the near-band rejection by introducing transmission zeros (TZs) into the filter response. Both the mechanical and EM requirements can be fulfilled by the realization of a design based on parallel shorted $\lambda/4$ stubs, whose design equations and equivalent circuit can be found in [18]. This concept has already been successfully demonstrated in a 3D-printed coaxial-line filter design [15]. Note that the BPF (4th-order Chebyshev) is designed with the desired passband frequency range (X-band: 8-12 GHz), whereas the LPF cutoff frequency matches the upper passband edge, $f_c = 12$ GHz. However, when the two filter sections are interconnected, some unwanted resonances may appear in the near-band rejection region. In order to avoid this spurious effect, the upper BPF cutoff frequency is slightly shifted beyond that of the LPF, in the final optimization.

III. PROTOTYPE MANUFACTURING AND MEASUREMENTS

A monolithic prototype with the bandpass and low-pass filter sections previously described is now 3D-printed in metal by Northern Waves AB [22] using SLM. Each layer of powder has a typical thickness between 30 and 50 μm . The expected dimensional tolerances at each layer are $\pm 50 \mu\text{m}$. The powder employed is a suitable aluminum-silicon-based alloy (AlSi10Mg), with an expected effective conductivity of $\sigma = 1.67 \cdot 10^7 \text{ S/m}$ and a RMS surface roughness of 3 μm . No post-processing is applied internally, only the surfaces corresponding to the flanges are machined to ensure good contact with the I/O connectors. For the connection with a Vector Network Analyzer (VNA) Agilent PNA E8364B, a 2.4 mm coaxial connector is employed at each port, in order to ensure correct measurement of the S-Parameters up to 50 GHz. The prototype is connected to the VNA as shown in the inset in Fig. 4 and measured. The measurement results are compared with the CST simulation of the complete filter (cascaded LPF and BPF sections) in Fig. 4. The measured bandwidth and cut-off frequency show excellent agreement with the full-wave simulation of the filter. The $|S_{11}|$ equalization is slightly worse compared to the simulation, although with RL always better than 19 dB. The measured insertion

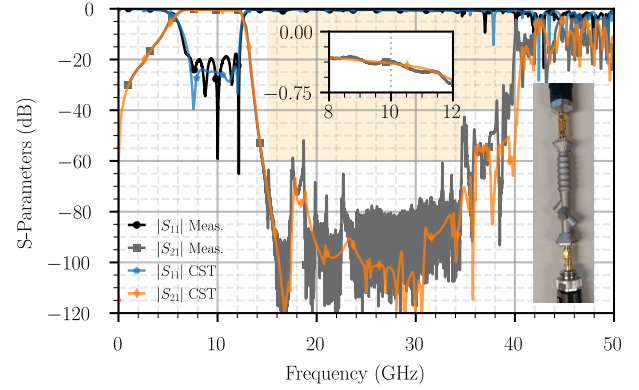


Fig. 4. S-parameter response of the measured prototype, as compared to the CST simulation with lossy materials. Inset (top): IL in the passband. Inset (side): photograph of the manufactured prototype.

TABLE I
COMPARISON WITH RELATED STATE-OF-THE-ART SOLUTIONS

Ref.	f_0 (GHz)	FBW (%)	IL (dB)	Rejection bandwidth (GHz)	Min. A_s (dB)	SF ₄₀	Manufacturing technology	Size (length)
[9]	3	3.3	1.45	3.20 - 9.15 ($3f_0$)	50	2.96	SLA, two halves	$1.28\lambda_0$
[23]	2	1.8	1.34	2.1 - 3.3 ($1.7f_0$)	50	4	CNC+3D-printed	$1.44\lambda_0$
[15]	10	40	0.72	15 - 30 ($3f_0$)	50	2.08	SLM, one piece	$2.99\lambda_0$
T.W.	10	40	0.37	15 - 39.5 ($3.95f_0$)	60/40	1.97	SLM, one piece	$3.27\lambda_0$

Ref.: reference, f_0 : center frequency, λ_0 : wavelength at f_0 , FBW: fractional bandwidth, IL: insertion loss, A_s : stopband attenuation, SF₄₀: shape factor, defined as $BW_{40\text{dB}}/BW_{3\text{dB}}$, i.e., 40dB bandwidth to 3dB bandwidth ratio, SLA: stereolithography apparatus, SLM: selective laser melting, T.W.: this work.

loss (IL) at the passband central frequency is ≈ 0.37 dB. These small measurement deviations align perfectly with the expected tolerance analysis results, considering the tolerances specified by the manufacturer. Furthermore, the rejected band also satisfies the masks required for our PDHT application. A rejection level better than 60 dB is achieved between 15 and 34.55 GHz. Moreover, for the higher stopband frequencies between 34.55 GHz and 38.5 GHz, a rejection level over 40 dB is obtained, perfectly covering the expected performance for an EO satellite. A comparison of the proposed filter with other 3D-printed coaxial-line filters [9], [15], [23] is presented in Table I. It can be concluded that the proposed filter design is superior to related solutions, both in terms of insertion loss and out-of-band performance.

IV. CONCLUSION

In this letter, we report on a new air-filled coaxial-line filter featuring a smooth profile, oriented to 3-D printing in metal. The new design is obtained by applying an inverse scattering synthesis method followed by shape deformation of the smoothly-varying inner and outer conductors of the low-pass structure, in order to achieve a wide rejected-band performance and make it compatible with the requirements of a PDHT module. A very compact coaxial-line X-band filter was designed and manufactured using SLM. The fabricated component meets the stopband demanded by current EO satellite systems and exhibits the widest stopband rejection among the related state-of-the-art solutions.

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4.5. Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components

Throughout all the studies performed in the scope of this thesis which have been described so far, various constraints had to be faced related with direct metal 3D printing as a method for manufacturing high-frequency electronic components. These include, most of all, two limiting factors: (a) printing constraints which need to be followed, and (b) signal power loss related to surface roughness and deteriorated metal conductivity. This section contains a concise author's commentary on the aforementioned factors, with a special emphasis on the lessons learned over the course of these studies. Firstly, a discussion on the specific printing constraints and related design limitations is included in Section 4.5.1. It is followed in Section 4.5.2 by a summary of obtained effective metal conductivity values among the 3D-printed parts produced within all the described activities. Finally, a way for reducing the conductor loss by improving surface roughness was addressed in [MB6], where a low-cost post-processing method of surface polishing was put forward as a potential solution to overcome these issues. This work was done in cooperation with the researchers from AGH University of Krakow, Poland (leading partner) and the University of Pavia, Italy. The author's contribution focused on the design of smooth-profiled filters adapted for AM, which were used in the study as test units. The results related to that investigation are summarized in Section 4.5.3.

4.5.1. Printing constraints

Even though shape deformation techniques allow the modeling of shapes of arbitrary complexity, several practical limitations have to be taken into account during the design process. The constraints of direct metal 3D printing process forbid the surfaces of the structure to be tilted at an angle greater than a certain threshold, usually $\approx 40 - 45^\circ$ along the printing axis. Should such a situation occur, extra supporting posts need to be placed inside the model. These supports then need to be removed in a post-printing step, which could be troublesome or even impossible for some designs which are to be manufactured in one piece. In one of the prototyping attempts of an eighth-order filter design which was described in Section 4.3, a 3D-printed model was first investigated that did not follow these limitations—i.e., it contained overhanging walls, titled at an angle steeper than 45 degrees along the printing direction. Images taken under the microscope, given in Fig. 4.32, show the inside surfaces of the printed model, which was cut open for inspection. The walls which were overhanging are marked with red arrows. It is clearly visible that the roughness of these surfaces is far greater than that of the surrounding areas, which results in a deteriorated Q-factor and greater conductor losses. Moreover, a visible change in the surface level, i.e. a "step" between adjacent faces, can be discerned in these pictures (marked by blue arrows). This discontinuity results in a change of physical dimensions of the structure and is the most probable cause of the filter's mismatch and shift in operating frequency, which can be observed in Fig. 4.33. This ex-

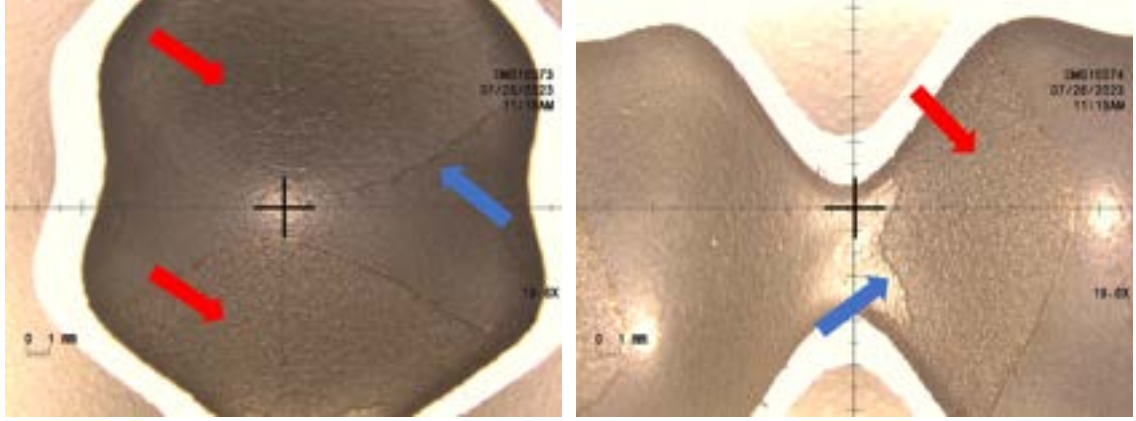


Fig. 4.32: Microscope photographs of the inside surfaces of the 3D-printed filter with overhanging walls (marked with red arrows). The abrupt steps in surface profile are marked with blue arrows.

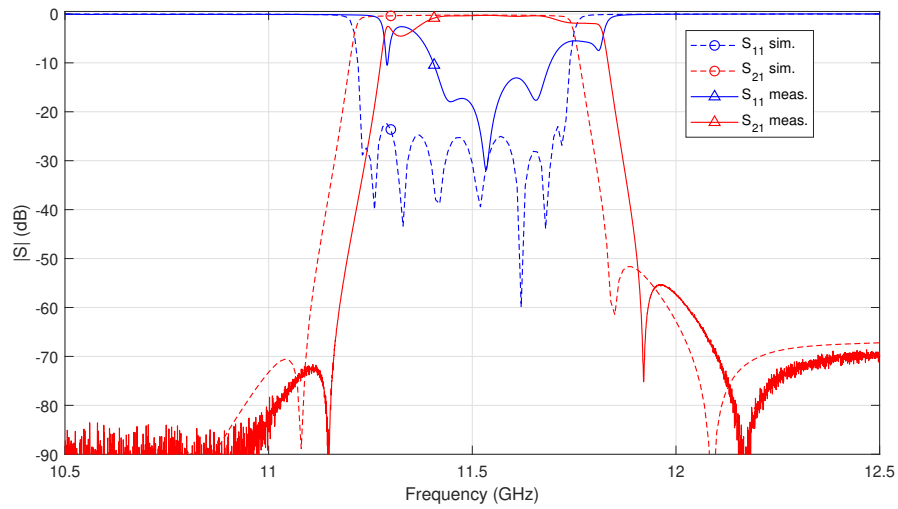


Fig. 4.33: Comparison of S-parameter results for the eighth-order dual-mode filter experimental prototype. Dashed lines: EM simulation, solid lines: measurement.

periment challenged this limitation and demonstrated that achieving high accuracy of 3D-printed microwave components requires adherence to the constraints of AM. Specifically, compliance with AM-oriented design principles, or, alternatively, the incorporation of additional structural supports within the design is necessary to ensure good quality of printing.

4.5.2. Estimation of effective metal conductivity

Another critical consideration in the design of microwave components fabricated through direct metal 3D printing is the conductor loss associated with the printed structures. This parameter can be expressed in terms of the effective electrical conductivity of the additively manufactured metallic parts. In all the experiments described in this work, the effective metal conductivity was estimated by matching the insertion loss level from a full-wave EM simulation assuming lossy metal surfaces with the one obtained from the measurement. In the designs which involved the insertion of tuning screws, such elements were also included in the simulated models. This way, the effec-

tive conductivity parameter can be treated as a loss indicator, taking into account all the effects occurring in the measured model, in particular the surface roughness and limited printing resolution. The comparison of all the fabricated components and their estimated conductivity values is provided in Table 4.2. The table summarizes the available information about the 3D-printing process of each presented component: the model or type of the 3D printer, the laser beam size (spot diameter), layer thickness and, where available, the average surface roughness declared by the manufacturer. It should be noted that, in AM based on powder bed fusion, the printing resolution is conventionally characterized in two dimensions: the XY resolution (horizontal), which denotes the minimum feature size achievable in the horizontal plane and is strongly affected by the laser beam diameter; and the Z resolution (vertical), which is governed by the applied layer thickness and the powder deposition height. Therefore, the overall printing resolution is ultimately depending on a combination of factors, most of all laser beam diameter and layer thickness, but also other aspects such as powder particle size or process control.

When analyzing the data in Table 4.2, it can be concluded that the effective conductivity values obtained for the first five listed components were quite comparable (in the range of 4–5 MS/m), despite the differences in the printing process parameters. Although such a repeatability can be treated as a good result, this estimated conductivity is an order of magnitude smaller than the theoretical value for perfect aluminum ($\sigma_{Al} = 38$ MS/m), and is indeed inferior to the values achievable with conventional fabrication processes. In fact, the worst achieved conductivity was reached for the OMUX channel filter from [MB3], fabricated with TRUMPF TruPrint 2000 printer with good printing parameters (beam size 55 μm , layer thickness 20 μm). It is important to note,

Tab. 4.2: Comparison of estimated effective conductivity values for all 3D printed prototypes

Component	Estimated conductivity (MS/m)	3D printer model / manufacturer	Laser beam size (μm)	Layer thickness (μm)	Declared surface roughness, R_a (μm)
6 th -order inline filter from Section 3.2 [MB1]	4.2	EOS M290	100	60	15
6 th -order 90°-twisted filter from Section 4.1 [MB2]	5	EOS M290	100	60	15
3 rd -order deformed filter from Section 4.1 [MB2]	4.2	TRUMPF TruPrint 1000 [100]	55	20	N/A
6 th -order deformed filter from Section 4.1 [MB2]	4.2	TRUMPF TruPrint 1000 [100]	55	20	N/A
2 nd -order CWDM filter from Section 4.2 [MB3]	4.2	TRUMPF TruPrint 1000 [100]	55	20	N/A
4 th -order CWDM filter from Section 4.2 [MB3]	1.72	TRUMPF TruPrint 2000 [100]	55	20	N/A
4 th -order DM filter from Section 4.3 [MB4]	9	EOS	N/A	30	N/A
8 th -order DM filter from Section 4.3 [MB4]	10 ⁽¹⁾	N/A [101]	N/A ⁽²⁾	30–50	4–6 ⁽¹⁾
Coaxial-line filter from Section 4.4 [MB5]	N/A (16.7) ⁽³⁾	N/A [101]	N/A ⁽²⁾	30–50	8–10

N/A: Not available. ⁽¹⁾ The process involved sandblasting (surface polishing). ⁽²⁾ The manufacturer only specifies the dimensional tolerances at each layer (50 μm). ⁽³⁾ Value declared by the manufacturer. Fair estimation from measurement not possible due to the influence of coaxial connectors.

however, that these parameters primarily influence the quality of the printed model and the accuracy of reproducing its physical dimensions, which is clearly visible in the filter's response (small deviation from simulation or frequency shifts observed in Fig. 4.16). However, the effective conductivity and resulting conductor losses are mainly associated with the conductivity of the material (i.e., the selected metal powder) and the roughness of the printed surfaces. This relation is clearly visible when comparing the printed components from [MB1, MB2], with declared average surface roughness $Ra = 15 \mu\text{m}$, with a more recent design, described in [MB4], specifically manufactured with an intention of minimizing the surface roughness down to $Ra = 4\text{--}6 \mu\text{m}$. The use of specialized AM processes, including post-processing such as internal surface polishing, resulted in a visible improvement in estimated conductivity (10 MS/m compared to 4–5 MS/m), even though the material used was the same (aluminum alloy AlSi10Mg). For this reason, the effect of surface polishing on improving the conductivity of 3D-printed components was examined more closely and summarized in the following section.

4.5.3. Conductor loss reduction by surface polishing

The study in [MB6], led by researchers from AGH University of Krakow, investigated the impact of a low-cost mechanical barrel-based surface polishing method on improving the surface quality of 3D-printed metal waveguide components for microwave applications. Waveguide components fabricated using powder bed fusion 3D-printing processes, such as SLM, typically exhibit surface roughness in the range of 10–50 μm . Such roughness contributes to increased conductor losses, which significantly affect the electrical performance of the devices. Therefore, a method for surface quality improvement was proposed, which makes use of glass microbeads (300–400 μm large) as an abrasive medium. The procedure involves inserting the printed component in a rotary tumbler, where it is exposed to mechanical treatment for a given period of time (e.g., 12 or 24 hours). This post-processing approach enables the abrasive particles to effectively access and treat elongated internal channels and cavities with cross-sectional dimensions ranging from sub-millimeter to several millimeters.

The study in [MB6] was carried out on several waveguide sections and bandpass filters fabricated using SLM, operating from X-band to Ka-band, i.e., ranging from 8 to 40 GHz. The results selected for discussion in this section concern three waveguide filters, developed by the author, which were specifically designed for AM using shape deformation. The focus is placed on filtering structures because their performance is highly sensitive to manufacturing imperfections; specifically in this context, high surface roughness has a great influence on the achievable Q and insertion loss. Therefore, the effects of the proposed surface polishing will be clearly visible on the measured insertion loss. Apart from IL, the results are also demonstrated by a loss factor: *total power loss (TPL)*, defined as the difference between total normalized power (equal to 1) and the sum of reflection (S_{11}) and transmission (S_{21}): $TPL = 1 - (|S_{11}|^2 + |S_{21}|^2)$, expressed in decibels as: $TPL_{dB} = 10 \log_{10}(1) - 10 \log_{10}(|S_{11}|^2 + |S_{21}|^2)$. In such a notation, a lossless circuit with

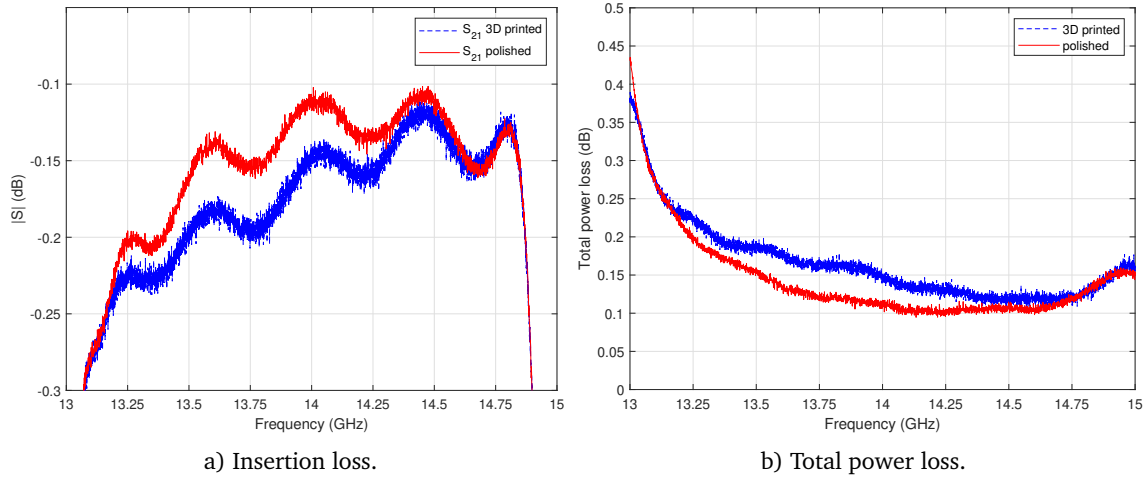


Fig. 4.34: Comparison of measured losses of the sixth-order inline filter in WR-75 waveguide from [MB1] before and after surface polishing. Material used: aluminum alloy (AlSi10Mg).

perfect matching would have a TPL value of 0 dB—i.e., the lower the value, the less lossy the circuit.

The first structure selected for testing was the sixth-order inline filter, primarily presented in [MB1]. The filter was designed in the WR-75 waveguide standard, with center frequency $f_0 = 13.9$ GHz and fractional bandwidth $\text{FBW} = 13\%$. The prototype was 3D-printed in aluminum (AlSi10Mg) by SLM and measured as is, with no post-processing. The test unit was then treated by dry tumbling using the Vevor KD-2000 jewelry grade mini rotary tumbler. The comparison of the measured losses before and after surface polishing is shown in Fig. 4.34. The insertion loss at f_0 is equal to 0.17 dB before polishing and 0.12 dB afterwards. Based on the achieved IL results from Fig. 4.34a, the difference in the estimated Q-factor of the filters is 2500 (polished) compared to 1800 (unpolished). In terms of effective metal conductivity, the untreated prototype exhibits a conductivity of $\sigma_{\text{before}} = 4.2$ MS/m, whereas due to surface polishing it is improved to $\sigma_{\text{after}} = 7$ MS/m.

The second and third test units prepared for examination within this study were also AM-optimized sixth-order BPFs such as the design from [MB1], scaled for higher frequency bands, i.e., 22–24 GHz (WR-42 waveguide) and 36.5–40 GHz (WR-28 waveguide). Assuming that the average surface roughness obtained in a given printing process is the same regardless of the size of the device, it follows that its negative effects on the filters' performance will be more pronounced at higher frequencies. For instance, an average roughness of $R_a = 50$ μm will be larger relative to the wavelength at e.g. 40 GHz ($\lambda = 7.5$ mm), compared to 8 GHz ($\lambda = 37.5$ mm). Moreover, unlike the previous one, these two filters were fabricated from stainless steel, which exhibits substantially lower electrical conductivity (≈ 1.5 MS/m) than aluminum. The use of a more lossy metal implies a change of the skin depth (i.e., the penetration depth of the electric current into a conductor); As an example, skin depth at 30 GHz for copper (60 MS/m) equals ≈ 0.38 μm while for stainless steel (1.5 MS/m) it is ≈ 2.38 μm [123]. For these reasons, the two filters were re-designed to operate

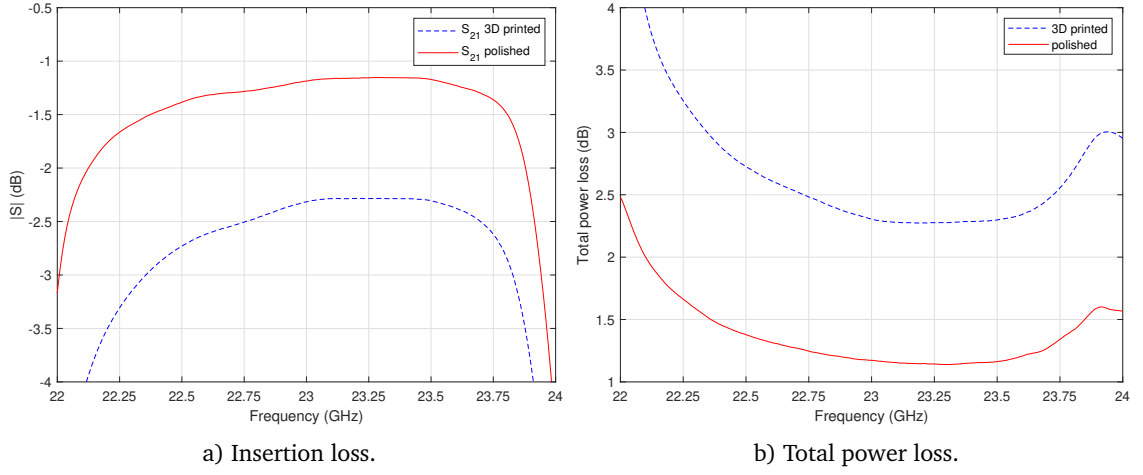


Fig. 4.35: Comparison of measured losses of the sixth-order inline filter in WR-42 waveguide from [MB6] before and after surface polishing. Material used: stainless steel 316L (SS316L).

at higher frequencies, and the prototypes were fabricated in stainless steel 316L, in order to better observe potential improvement attributed to surface polishing. The first of these two components was designed for the WR-42 waveguide, with $f_0 = 23$ GHz and a relative bandwidth of 9%. The filter underwent surface polishing treatment for 48 hours. The compared insertion loss and the resulting total power loss, measured before and after surface polishing, are given in Fig. 4.35a and Fig. 4.35b, respectively. Here, the achieved improvement in IL can be observed most clearly: the IL value at f_0 is 1.19 dB with respect to 2.31 dB pre-polishing. For the third test filter, designed in WR-28 waveguide standard, the following specifications were assumed: $f_0 = 38.25$ GHz, bandwidth $BW = 36.5\text{--}40$ GHz (FBW = 9%). The prototype was measured before and after mechanical polishing, and the compared results are shown in Fig. 4.36. The improvement, although not as strong as for the WR-42 filter, is definitely noticeable: IL at f_0 equals 1.15 dB (after polishing) vs. 1.48 dB (before). One potential reason why the loss reduction in WR-28 BPF is not so effective is that the minimum iris size is only an order of magnitude larger than the beads radius. Therefore,

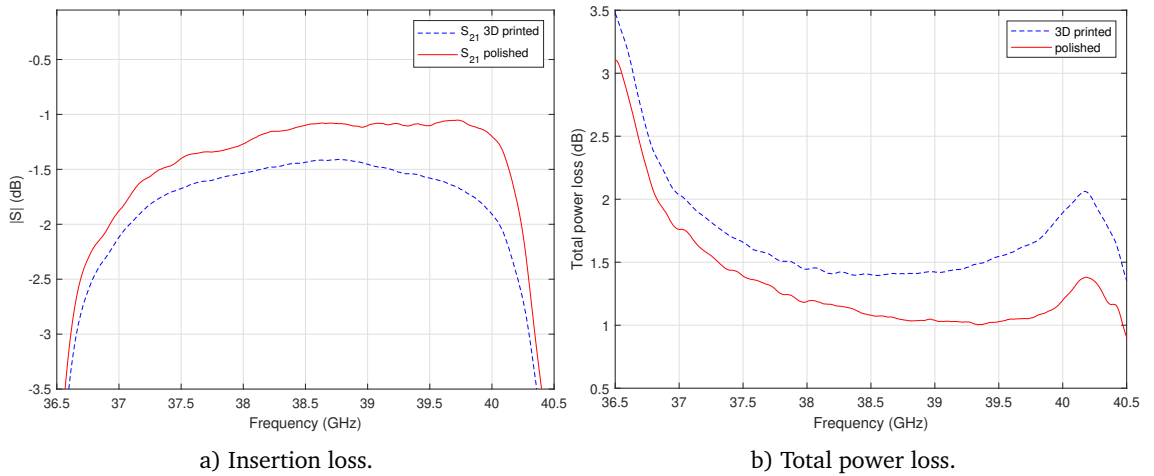


Fig. 4.36: Comparison of measured losses of the sixth-order inline filter in WR-28 waveguide from [MB6] before and after surface polishing. Material used: stainless steel 316L (SS316L).

for higher frequencies, it is assumed that smaller beads (e.g. 150-200 μm in radius) could yield better results, since the abrasive medium must freely penetrate the interior of the structure.

In conclusion, the obtained results validate the use of the proposed low-cost polishing method based on mechanical barrel tumbling for reducing conduction losses of 3D-printed monolithic waveguide components. Apart from the microwave filter examples discussed here, more test structures were manufactured and investigated by the AGH team, such as waveguide sections, twists and bends, all of which were described in the publication [MB6]. All the examined parts demonstrate visible loss reduction after applying the proposed post-processing procedure. In contrast to electropolishing, this approach eliminates the need for chemical reagents or specialized apparatus, while still delivering comparable outcomes in terms of internal channel and cavity penetration and polishing. In terms of surface roughness Ra , the proposed method brings improvement of roughly 15–30%, which is comparable to the industry-standard high-cost and high-performance techniques. The achieved enhancement in surface quality translates into 10 to 40% loss reduction of the tested waveguide filters. The study shows that, with additional post-processing, powder-based AM methods can be used to produce advanced microwave components with surface quality approaching that achievable with more established manufacturing techniques. It is expected that with the continuous development of high-precision 3D printing technology and its increasingly widespread use in areas such as microwave engineering, the observed gap between AM and other fabrication methods will continue to diminish.

**[MB6] Low-Cost Method for Internal Surface Roughness Reduction
of Additively Manufactured All-Metal Waveguide Components**

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Design of the two smooth-profile sixth-order waveguide filters used as test units in the paper,
preparation of the 3D models for AM, contribution in writing and editing the manuscript.

Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components

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Abstract—In this study, a novel low-cost polishing method for internal surface roughness reduction of additively manufactured components, developed for waveguide (WG) circuits operating in the millimeter frequency range is proposed. WG components fabricated using powder bed fusion (PBF) generally feature roughness of ten to fifty microns, which influences the increase of roughness-related conductor power losses having a major effect on the electrical performance of additively manufactured all-metal WGs. To improve and decrease the surface roughness of circuits fabricated using PBF, glass microbeads as an abrasive medium are proposed to be used in combination with a rotary tumbler. This technique allows the abrasive medium to efficiently penetrate internal long channels and cavities, having cross section dimensions in the range of sub- to a few millimeters. An experimental study was carried out on an example of WG sections and bandpass filters fabricated using PBF through selective laser melting (SLM), operating within the 8.2 to 40 GHz range. Polishing impact on both mechanical and electrical properties was studied showing surface roughness reduction by 18% and sixth order filter's insertion loss reduction at 23 GHz by 40% after 24 h of tumbling with 300–400 μm large glass microbeads.

Index Terms—Additive manufacturing (AM), all-metal waveguide (WG), glass microbeads, mm-wave electronics, power losses, surface polishing, surface roughness, tumbling, WG.

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I. INTRODUCTION

INCREASING demand for wireless communication systems throughput has been observed in recent years, which requires more efficient use of available resources [1], [2]. Extensive research efforts are directed toward the development of novel solutions and technologies allowing highly integrated and energy-efficient front-end production. A wide range of basic microwave building blocks such as: 1) filters [3], [4]; 2) power dividing/combining networks [5], [6]; 3) frequency multiplexers [7], [8]; and 4) antennas and antenna arrays' feeding networks [6], [9], and so on, are realized in strip transmission line and/or substrate integrated waveguide (SIW) techniques, all made with subtractive techniques. Despite many advantages, such circuits suffer from relatively high power loss being a combination of dielectric, conductor, and radiated losses. Low-loss passive devices are commonly produced with the all-metal waveguide (WG) technique [10], [11], where the wave propagates in lossless air, while attenuation is only due to the finite effective conductivity (combination of bulk conductivity, surface roughness, and frequency of operation) of the metallic walls. However, the main drawback of WG techniques relates to the physical production of such circuits, which requires expensive and precise machining. Moreover, the realization of complex structures is limited by subtractive manufacturing technologies such as computer numerical control (CNC) and electro discharge machining (EDM). Such techniques also cannot be used for the realization of various high-frequency WG circuits in a single piece, which allows for avoiding power losses due to electrical contact issues among different parts [12].

To challenge the above issues additive manufacturing (AM) technologies have been proposed as an alternative for the production of various high-frequency electronic circuits [12], [13], [14], [15]. WG components in such a scheme can be printed out of nonconductive material using, e.g., stereolithography process (SLA) [16], [17], [18], fused filament deposition (FFD) [19], [20], and polyjet printing [21], [22], [23]. Nevertheless, such approaches require an additional fabrication step, during which the 3-D-printed

plastic components are metal coated. An alternative solution is direct 3-D printing out of metallic materials [14] such as stainless steel, aluminum, titanium, nickel, or copper using powder bed fusion (PBF) technologies such as binder jetting (BJ) [24], [25], metal selective laser melting (SLM) [26], [27] and direct metal sintering (DMS) [28], [29]. In the context of RF/microwaves, SLM represents one of the most promising AM techniques, as complex geometries with internal features and channels can be easily realized.

One of the most significant issues related to WG circuit production through SLM technology, in comparison with subtractive technologies, is the high surface roughness [30], [31] on manufactured components. Roughness influences the insertion losses of fabricated circuits, especially visible at higher frequencies [32], [33], [34], [35], [36], [37]. In literature, many solutions for smoothing additively manufactured surfaces are presented, but most of them are used on plastic components [38], [39], [40]. For metal parts, on the other hand, mechanical polishing techniques are commonly used where vibration, tumbling, or rotation action forces the abrasive medium against the surface [41]. With these methods, outer surfaces are easily processed, while inner walls are challenging, especially for parts with fine internal features, since usually, the abrasive medium is too large to fit inside or too aggressive. An alternative is the electropolishing process [41], where the current-assisted reverse plating process in an electrolyte bath is used for smoothing the surface. It is well suited for the discussed application as the liquid could easily penetrate the interior. However, chemicals are required, making the process more expensive.

In this article, a novel low-cost polishing method for metal additively manufactured components is presented. This method is applied to WG components operating in the millimeter frequency range. In this regime, a wall roughness of tens of microns is hardly acceptable. Rough internal surfaces impact the electrical performance of components as it results in lower effective conductivity of walls as compared to bulk value. This is of importance since conductor-related losses are the only source of attenuation within the circuits, while in many cases metal type, composition, or bulk conductivity, is technology specific and cannot be altered to some extent. To overcome the challenge of the abrasive medium efficiently penetrating internals with long channels and/or (multiple) cavities having cross section dimensions in the range sub- to a few millimeters, a rotary tumbler is proposed to be used in combination with micro glass beads. An experimental study is carried out on a set of additively manufactured WG sections and bandpass filters operating within cm- and mm-wave frequency range, produced through SLM technology. Polishing impact on both mechanical and electrical properties was studied, showing surface roughness reduction by 18% and sixth order filter's insertion loss reduction at 23 GHz by 40%, after 24 h of tumbling with 300–400 μm large microbeads, hence validating the superior performance and applicability of the proposed method.

II. ADDITIVELY MANUFACTURED ALL-METAL WGS

In this section, the conductor and surface properties of additively manufactured all-metal WGs are studied. The impact

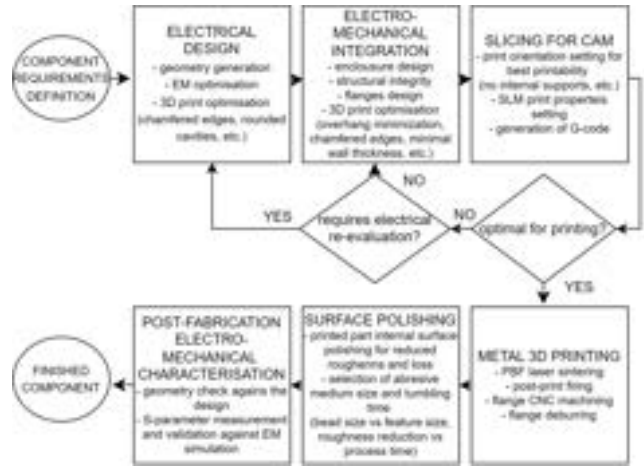


Fig. 1. Flowchart of the design to fabricate the process of microwave components using the proposed approach.

of effective conductivity (influenced by metal conductivity), frequency of operation, and surface roughness on total power loss (TPL) (which is the key electrical performance metric for metal WG circuits), is elaborated. A low-cost surface polishing method for additively manufactured printed WG components postprocessing is introduced. Finally, test vehicles for experimental assessment of the method's performance are developed.

A general flowchart of the employed design to fabricate process is depicted in Fig. 1 where two domains are visible: digital preparation and physical fabrication. First, the component type and related requirements are defined. This is followed by an electrical design where the components' geometry is determined by means of full-wave electromagnetic (EM) simulations and optimization using computer aided desing (CAD) software. At this stage, the geometry optimization may already take advantage of the properties of the intended fabrication technology, i.e., 3-D printing which provides true 3-D flexibility. In addition, features that are helpful for better printability may already be accounted for such as edge chamfering or rounded corners. This is followed by electro-mechanical integration using yet another type of CAD software where mechanical features ensuring structural integrity such as wall thickness, coupling flanges, mounting holes, sub-component integration, and 3-D printing requirements adjustment are done. The next step is file preparation for computer aided manufacturing (CAM) using slicing software. Parameters such as part placement and print orientation are set and the software slices the model into layers and generates the *G*-code for CAM. It provides a report with potential print issues such as print instability, overhangs, internal supports, too-thin walls, and so on. If any problem is indicated, either one or two earlier steps must be re-revisited depending on whether mechanical fixes may or may not affect the electrical performance. Otherwise, the part fabrication begins starting with 3-D printing using the SLM technology out of metal powder. Then the part is cut out of the built platform and if necessary, CNC machined to, e.g., flatten/surface flanges. Finally, the proposed post-processing method is applied to reduce the surface roughness

of the print where glass beads are used as a polishing medium to tumble along with the part leading to reduced roughness and thus reduced power losses. The finished part is then characterized in terms of mechanical (geometry accuracy) and electrical (fulfillment of design requirements). If the objectives are fulfilled then the part is ready, if not the data can be fed back to the design loop and corrected for in the next iteration.

A. Surface Roughness Impact on WG Propagation Characteristics

To investigate the fabrication method's effect on WG components, we first consider wave propagation properties through a section of an air-filled rectangular WG operating in the fundamental TE₁₀ mode. Signal transmission through a matched ($S_{11} = S_{22} = 0$) transmission line section is expressed in terms of scattering parameters (S-parameters) as follows:

$$S_{21} = S_{12} = e^{-\gamma l} = e^{-(\alpha + j\beta)l} \quad (1)$$

where l is the physical length of the lines, while γ is the complex propagation constant, and α and β are, respectively, the attenuation and propagation constants. As provided in [42], the complex propagation constant of the TE₁₀ mode is

$$\gamma = \sqrt{(R' + j\omega L') \left(\frac{1}{R'' + j\omega L''} + G' + j\omega C' \right)} \quad (2)$$

and can be expressed in terms of per- and times-unit-length model resistance R (conductor-related losses), inductance L , conductance G (dielectric-related losses), and capacitance C . Conductivity and surface roughness of the guide's conductor has a significant impact on parameters representing the interaction of magnetic fields with conductive media, namely R' and R'' , and partially L' and L'' . In the case of an ideally smooth surface that represents the abrupt transition from dielectric into conductive media, the skin effect occurs, and thus, the current skin depth can be defined as follows [43]:

$$\delta = \frac{1}{\sqrt{\pi \mu_0 \mu_r \sigma_{dc} f}} \quad (3)$$

where μ_0 and μ_r are the vacuum and relative permeability, σ_{dc} is the conductivity, and f is the frequency. Nevertheless, it needs to be underlined, when the skin depth decreases to a similar range as the variation in surface profile due to roughness (as a reference, skin depth at 30 GHz for copper (60 MS/m) $\approx 0.38 \mu\text{m}$, while for stainless steel (1.5 MS/m) $\approx 2.38 \mu\text{m}$) [42], then (3) is no longer valid. In such a case, the conductivity is a variable parameter continuously depending on the spatial coordinates. It can be determined using, e.g., the Gradient Model [44] that requires only the rms roughness value R_q if the surface roughness follows the normal distribution. The impact of surface roughness is transferred into effective, frequency-dependent material parameters σ_{eff} (lower than σ_{dc}), and $\mu_{r,\text{eff}}$ (higher than μ_r). Then, the skin depths in rough surfaces for current densities δ_c and magnetic fields δ_m can be derived [42], to finally obtain R and L . The ohmic loss R mainly affects the attenuation coefficient, while the inner inductance L is related to the phase coefficient. Moreover, it is necessary to differentiate

between the frequency region close to the cutoff frequency f_c and the usually used transmission region of a WG from $1.25 f_c$ to $1.9 f_c$, since rough surfaces affect the propagation properties with different orders of magnitude. On the one hand, the cutoff frequency decreases with increasing R_q due to an increase in both R and L , but on the other hand, increasing R_q imposes a severe increase of attenuation constant α in the typical frequency range. When R_q is much higher than the skin depth, b is increased at least twofold [42]. The impact of increased L is especially reflected in the phase coefficient β in the frequency region around f_c . Like α , the responses in β also exhibit a shift toward lower frequencies with increasing R_q . In addition, the slope reduces during the transition from attenuation to transmission region. However, in the typical frequency range of the WG, β is relatively insensitive to variations in R_q .

The previous study can be extended to resonators, which are basic building blocks for many circuits such as filters. The quality factor Q of a resonator describes its ability to store energy, and the unloaded quality factor Q_U is the maximum quality factor of the system. When there are only conductor losses, the Q_U value is inversely proportional to α and thus decreases when R_q increases. High Q_U is especially important when high selectivity filters operating at high frequency with low in-band insertion losses IL are to be realized, since [45]

$$\text{IL}(f_0) = 4.343 \frac{N}{Q_U \frac{\Delta f}{f_0}} \quad (4)$$

where N is the filter's order, while $\Delta f/f_0$ and f_0 are the fractional bandwidth and midband frequency, respectively. The impact of conductor properties (including surface roughness) becomes more pronounced as the number of resonators increases. The passband of a filter is also affected since R_q affects the propagation constant and in turn the resonant frequency. However, as indicated earlier within the WG's typical frequency range, this impact is relatively low and is more visible for highly selective filters.

In summary, the overall power losses in WG circuits depend on the frequency range of operation, bulk conductivity, surface roughness, and WG geometry. Therefore, it is crucial for a given fabrication technology and material to provide the smoothest inner surfaces where current flows and magnetic fields propagate, to minimize power losses.

B. Test Vehicles

The analysis presented in Section II-A suggests that two types of circuits might benefit differently from low surface roughness, namely: broadband sections of transmission lines where the attenuation constant is minimized, and narrowband-like filters where the unloaded quality factor is maximized. Therefore, a set of test vehicles was designed to include both types to operate in the cm/mm-wave frequency range.

First, rectangular WGs in various geometries were considered.

1) A 150 mm long straight section in WR-90 geometry (cross section of 22.86×10.16 mm, recommended band between 8.2 and 12.4 GHz).

2) A 100 mm long 90° twisted (five wavelengths) section in WR-42 geometry (cross section of 10.67×4.32 mm, recommended band between 18 and 26.5 GHz).

3) A 100 mm long straight section in WR-28 geometry (cross section of 7.11×3.56 mm, recommended band between 26.5 and 40 GHz).

Each guide was equipped with a flange for mating with the measurement equipment. On top of loss reduction, the usability of the method for complicated internal walls such as spiraled or rotated ones was benchmarked with the WG twist. Relatively long sections were realized to maximize the γ/l term change due to attenuation factor change since the measurement of low losses is limited by measurement system uncertainty. The full-wave models were simulated in *Ansys HFSS* software assuming metal properties and surface finish provided in Section II-C (Grosse model for roughness used). The predicted losses are 0.013 dB/cm for WR-90 at $f_{0_WR-90} = 10.5$ GHz, 0.048 dB/cm for WR-42 at $f_{0_WR-42} = 22.25$ GHz, and 0.073 dB/cm for WR-28 at $f_{0_WR-28} = 33.25$ GHz. Considering the EM-calculated losses of the WG, the unloaded quality factor can be established to be ~ 328 at f_{0_WR-42} .

Second, narrowband filters of different orders and bandwidths were developed.

1) A semi 3-D printing optimized second order bandpass filter, namely BPF2_42 in WR-42 geometry operating at the center frequency of 24 GHz with a relative bandwidth of $\sim 8\%$ for which the (4) yields $IL_{\min}$ of 0.265 dB. The minimal dimension of the resonator cavity is ~ 5.5 mm; the minimal iris opening is ~ 4.4 mm.

2) A fully 3-D printing optimized sixth order bandpass filter, namely BPF6_42 in WR-42 geometry operating at the center frequency of 23 GHz with a relative bandwidth of $\sim 10\%$ for which the (4) yields $IL_{\min}$ of 0.678 dB. The minimal dimension of the resonator cavity is ~ 10.6 mm; the minimal iris opening is ~ 4.4 mm.

3) A downscaled to WR-28 geometry version of the BPF6_42, namely BPF6_28 operating at the center frequency of 38 GHz. The minimal dimension of the resonator cavity is ~ 5.6 mm, while the minimal iris opening is ~ 3 mm.

Advanced layout optimization for AM [27], [46] was applied within a dedicated tool in *InventSim* software [47]. The final layout of both filters is provided in Fig. 2. BPF2 has a filter length of 20.6 mm, BPF6_42 of 73.3 mm, and BPF_28 of 39 mm. Adding the WG sections, the overall component length sums up to 50, 100, and 100 mm, respectively.

C. AM Process

In PBF, the physical and chemical properties of powder particles affect the final quality of additively manufactured components [48]. Fine particles enable high-density parts with good surface quality, while the spherical shape improves flowability and, as a result, mechanical properties. Irregular powder particles can lead to poor surface finish, low density, and increased defects.

In this article, PBF based on SM was adopted, namely the printer [49] was used to 3-D print the test vehicles out of

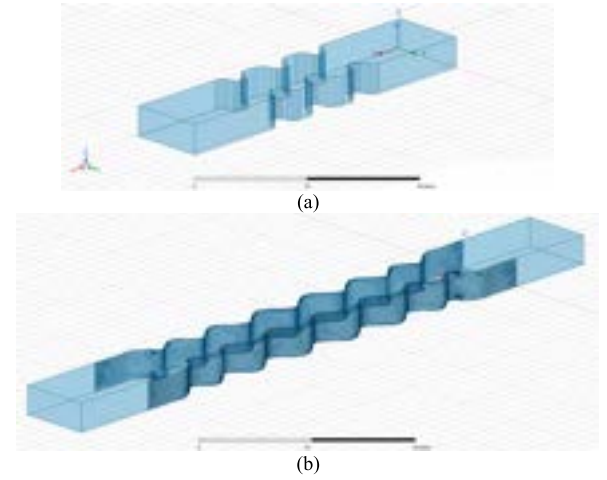


Fig. 2. Layout of the developed (a) second and (b) sixth order bandpass filters in WR-42 geometry featuring 3-D printing optimized geometry.

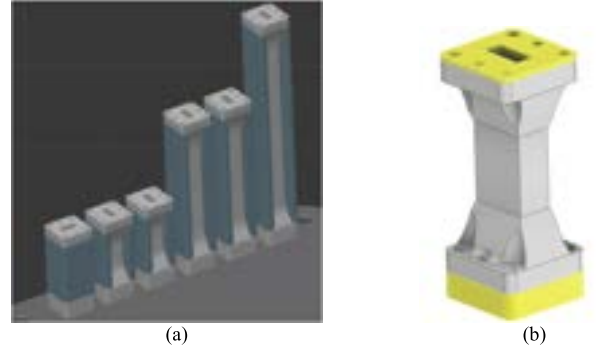


Fig. 3. Test vehicle orientation on the print bed with supports added (a) is set in the slicing software. Exemplary test vehicle view where yellow color indicates stock in the flange region that is removed and machined with CNC after detachment from building plate (b).

Stainless Steel 316L [50]. The average surface roughness (R_a) of SS316L additively manufactured components is commonly $11 \pm 2 \mu\text{m}$ (JIS B 0601-2001 -ISO 97) after bead blasting (assuming Gaussian probability density function (PDF) $R_q = 1.25$, $R_a = 13.75 \pm 2.5 \mu\text{m}$). All test vehicles (having a general form of a long rectangular tube) were oriented on the building plate with flanges parallel to the print bed as shown in Fig. 3. Such orientation results in support only on the outside of the WG to carry the load of the top flange contour, while there are no supports inside the WG cross section. Each printed test piece has the same surface finish with the same lay, i.e., parallel deposited metal layers oriented perpendicular to the direction of wave propagation. Finally, to reduce the number of variables affecting the measured power losses, both flanges were CNC-milled postfabrication so that very good electrical contact is ensured between the measurement instrument's WG flanges and device-under test WG flanges. The fabricated test vehicle set is presented in Fig. 4.

D. Surface Polishing

Additively manufactured metal parts usually have a poor surface finish in comparison to milled and turned components. Many post-processing techniques can ensure that additively manufactured metal parts meet dimensional requirements, performance characteristics, and even esthetic



Fig. 4. Photographs of the fabricated test vehicles.

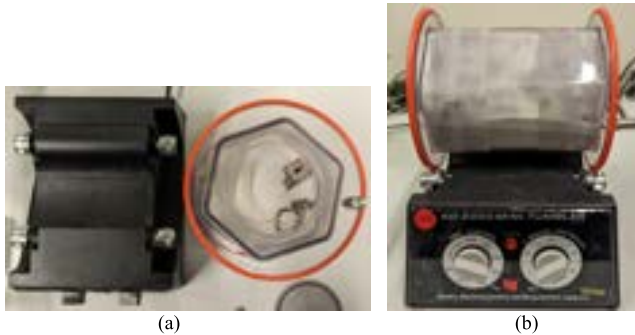


Fig. 5. Photograph of the setup used for polishing of parts. (a) Barrel loaded with glass microbeads. (b) Running machine during polishing cycle.

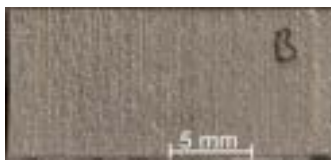


Fig. 6. Photograph of the fabricated sample for mechanical properties analysis. Layers orientation is clearly visible and thus the resulting roughness is orientation dependent. The sample area equals 20×10 mm.

guidelines. These processes include green sanding, media blasting, tumble finishing, pin finishing, vibratory finishing, and electropolishing. The WG components present internal channels, therefore, one of the most promising surface post-processing treatments is electropolishing [51], sometimes referred to as “reverse plating” as it utilizes a phosphoric-based electrolyte bath and rectified current to remove surface material. Electropolishing eliminates micro-peaks (high points within the microscopic surface) to reduce surface roughness to a higher level of smoothness than other finishing methods. Moreover, the electropolishing solution can reach every part of the component surface. This can include internal gaps and cavities that other finishing methods cannot reach. A smoother surface means fewer places for corrosive elements to take hold and thus extend part lifetime. Finally, electropolishing generally delivers a more consistent finish across the entire part surface. In contrast to that, many other polishing systems can leave marks or patterns on the surface. A large drawback of the method is its relatively high cost and careful handling as chemical bath is needed.

In this study, a mechanical barrel-based surface polishing method is introduced and validated. This method presents the advantages of electropolishing in combination with microbead-blasting Barrel polishing [41] is a surface-improving operation in which a mixture of parts, media, and compounds are placed

in a six- or eight-sided barrel and rotated at a predetermined speed to round corners, deburr, grinding, descaling, deflashing, improving surface finish, burnishing, polishing, and radiusing parts in bulk. Tumbling parts are placed in a rotating barrel, and an abrasive media is introduced inside the barrel. Producing good surface finishes depends on the right selection of tumblers, abrasives, lubricating agents, carrying agents, and polishing agents. Very little handling is required for the process and many metal parts can be processed at once while the process can be finished in a tumbling barrel overnight, or in an hour or less in a high-energy machine. Barrel speeds in dry tumbling are generally kept at 28–32 r/min. On the other hand, in the microbead-blasting process, a very fine abrasive media and a miniature nozzle are used to produce a controllable abrasive jet that can target and remove microns of material. The result is a reliable and repeatable method for deburring, texturing, cleaning, stripping, etching, or milling part surfaces. The abrasive medium is easily available in a gamut in hardness (from soft compounds like sodium bicarbonate to hard ceramics like silicon carbide), that suits different materials to be processed as well in different shapes from spherical to blocky or pointy and sizes ranging from a couple of microns to hundreds of microns. Despite advantages, the process itself uses a directed jet of medium and thus internal cavities and channels of additive manufactured components parts cannot be reached.

In this study, the dry tumbling method using the Vevor KD-2000 jewelry grade mini rotary tumbler (5 kg max load, 4.5 l barrel dimensions of 190×180 mm, opening diameter of 90 cm), in combination with 300–400 μm large glass microbeads was employed. Such a combination ensures that the container size is large enough to fit a few parts while the medium size is small enough to easily penetrate the insides of the parts even having openings in the range of single millimeters. Glass microbeads are well suited for processing stainless steel and due to their structure leave the surface a bit satin, uniform, and smooth. The larger the beads the more effective they are and shinier, hence smoother, the surface. For example, 40–70 μm leads to a matte finish whereas 300–400 μm returns very shiny surfaces. On the other hand, too-large beads will not penetrate chambers with small openings and will not affect areas in proximity to sharp corners. An additional benefit of using microbeads in the dry tumbling process is that they provide an extraordinary amount of surface area to carry the dirty residue, thus avoiding having the residue embedded into the surface of the parts. A photograph of the polishing setup is presented in Fig. 5. The tumbler is filled with beads up to roughly three quarters of its volume.

III. EXPERIMENTAL RESULTS

The proposed surface polishing method was experimentally validated within the postprocessing of test vehicles (broadband and resonant ones) operating within the cm/mm-wave frequency range. The impact of polishing on the resulting mechanical and electrical properties of circuits is provided and discussed in detail. Each test vehicle was measured before

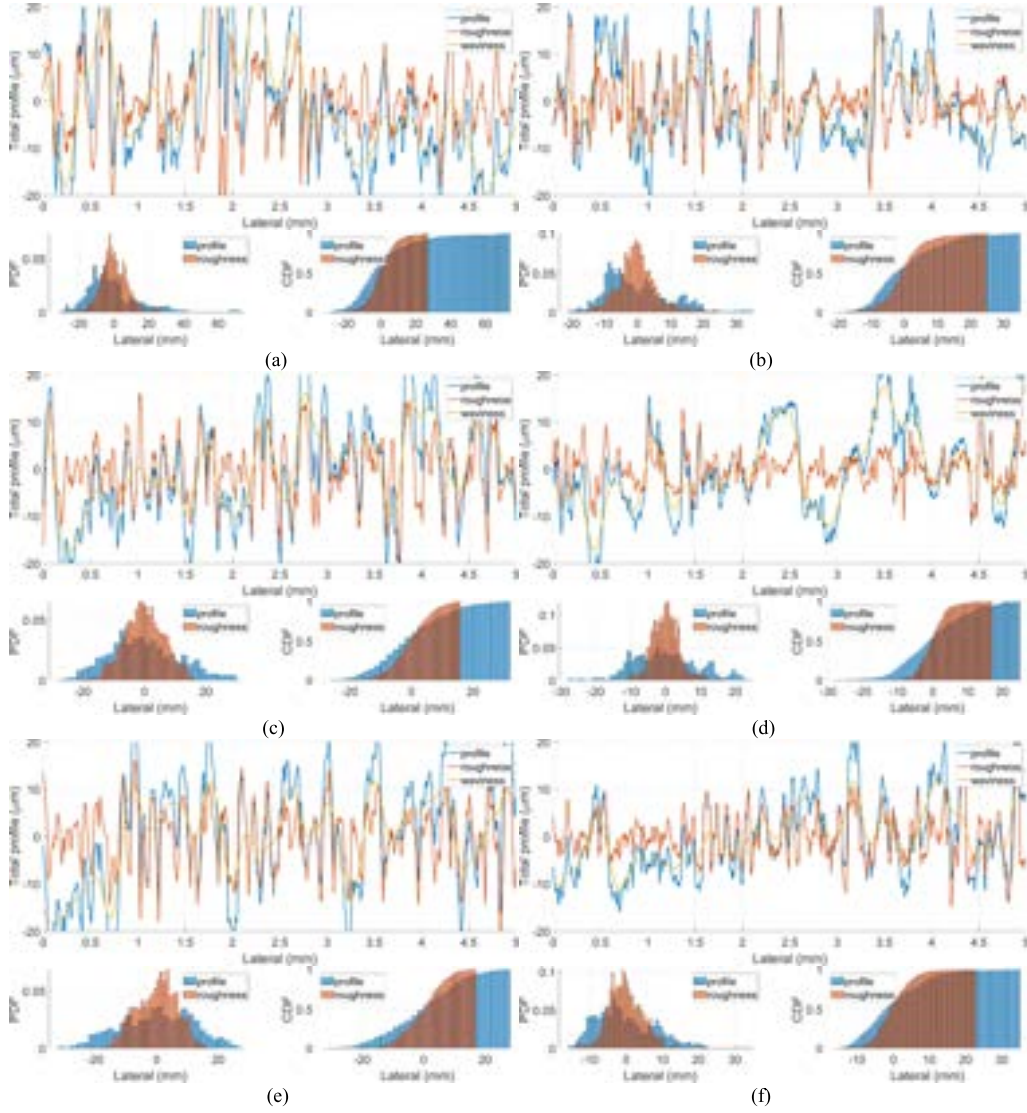


Fig. 7. Measured surface profiles of the printed specimen. (a) and (b) No polishing applied, (c) and (d) after 24 h of polishing, and (e) and (f) after 48 h of polishing. Profiles at orthogonal directions (orientation parallel (a), (c), and (e) and perpendicular (b), (d), and (f) to layer arrangement).

polishing and then after one and two cycles of polishing. Each 24-h (12-h) polishing cycle was carried out at 50 rotation-per-minute speed (to avoid microbeads sticking to internal surfaces due to centrifugal force instead of tumbling).

A. Mechanical Performance

First, the mechanical properties of additively manufactured test vehicles were assessed. Dimensions of the parts shown in Fig. 4 were measured using a digital caliper (10 μ m least significant digit) in two different spots and averaged to assess the fabrication tolerance including post-heat treatment shrinkage. The results are summarized in Table I: the internal cross section of WG components shrunk after all processing steps. Dimension reduction is in the range of 1%–3%.

In the next step, the Bruker Dektak XT stylus profilometer was used to evaluate surface roughness: 1) before polishing; 2) after 24 h of polishing; and 3) after 48 h of polishing. Two parameters were used for comparison purposes, i.e., rms roughness value R_q and the average surface roughness

TABLE I
MEASURED DIMENSIONS OF THE FABRICATED PROTOTYPES AND
CALCULATED DEVIATION Δ FROM THE DESIGN VALUES

	a (mm)	Δa	b (mm)	Δb
WR-90 section	22.68	-0.8%	10.07	-0.9%
WR-42 twist	10.59	-0.8%	4.23	-2.2%
WR-28 section	6.99	-1.8%	3.43	-3.7%
WR-42 BPF2_42	10.61	-0.6%	4.24	-2.0%
WR-42 BPF6_42	10.60	-0.7%	4.24	-2.0%
WR-28 BPF6_28	7.01	-1.5%	3.45	-3.2%

R_a . Since the internal surface profile of test vehicles could not be nondestructively measured between tumbling cycles, a dedicated $20 \times 10 \times 3$ mm specimen shown in Fig. 6 was fabricated and processed in the same way. At every step, an optical inspection was carried out as well, and a 5 mm long surface profile was captured in two orthogonal directions. The resulting roughness and waviness profiles are provided in Fig. 7, the optical images are shown in Fig. 8, while the average and rms roughness values are summarized in Table II.

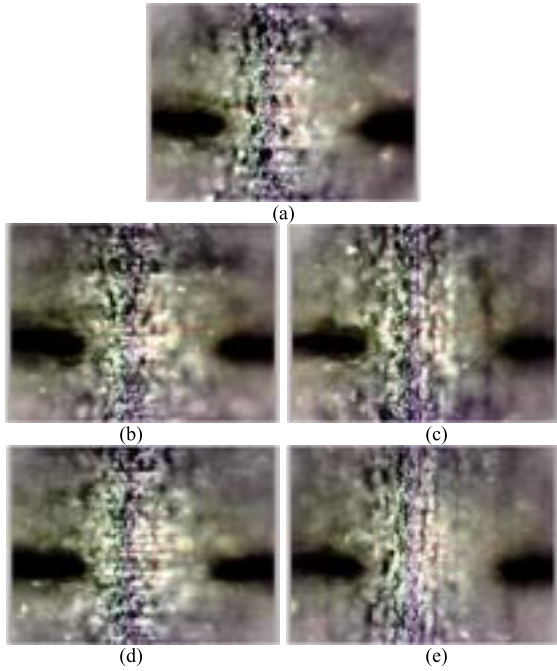


Fig. 8. Optical image of the printed specimen's surface. (a) No polishing applied, (b) and (c) after 24 h, and (d) and (e) after 48 h of polishing. The total picture width and height is 2.25 mm (1× zoom). Photographs at orthogonal directions [(b) and (d) orientation parallel and (c) and (e) perpendicular to layer arrangement]. The layered print pattern is clearly visible.

TABLE II

ESTABLISHED ROUGHNESS OF THE SLM PRINTED SPECIMEN FROM A 5 MM LONG PATH IN TWO ORTHOGONAL DIRECTIONS

Orientation		perpendicular to layer arrangement			
Profile	primary	roughness			
Metrics	R_a (μm)	R_q (μm)	R_a (μm)	R_q (μm)	
Prior polishing	12.05	16.63	6.00	7.85	
After 24h	9.23	11.38	5.19	6.47	
Improvement	23.4%	31.6%	13.5%	17.6%	
After 48h	9.73	11.82	5.37	6.58	
Improvement	-	-	-	-	
Orientation		parallel to layer arrangement			
Profile	primary	roughness			
Metrics	R_a (μm)	R_q (μm)	R_a (μm)	R_q (μm)	
Prior polishing	8.07	9.99	4.48	6.12	
After 24h	7.60	9.46	2.99	3.87	
Improvement	5.8%	5.3%	33.3%	36.8%	
After 48h	6.83	8.45	3.81	4.95	
Improvement	-	-	-	-	

Both the optical and the surface analysis reveal the typical characteristics of part made by SM: irregularities, larger melted grain clusters, and relatively large waviness. The collected profiles must be considered random and as just a local representation of the surface. Nevertheless, they allow us to assess the polishing technique performance to a certain degree.

The highest roughness is found in the planes parallel to the printing plate, while along the direction orthogonal to the printing plate a lower roughness is observed. In both directions, the PDF for roughness is close to the Gaussian distribution. Finally, it is clearly seen that the specimen surface is smoother after 24 h of polishing while an extra 24 h makes no significant difference. Roughness R_q reduction by 17% and

TABLE III

MEASURED IMPACT OF POLISHING ON WG ATTENUATION CONSTANT (EXPRESSED THROUGH TPL METRICS). POSITIVE DIFFERENCE Δ MEANS VALUE REDUCTION WITH RESPECT TO THE BASE ONE

		α @ f_0 (dB)	$\Delta\alpha$ (%)	α @ f_0 (dB/cm)
WR-90	Pre	0.186		0.0124
	After 24h	0.164	11.9	0.0109
	After 48h	0.195	-2.9	0.0130
WR-42 twisted	Pre	0.615		0.0615
	After 12h	0.506	17.7	0.0506
	After 24h	0.499	19.0	0.0490
WR-28	Pre	0.475		0.0475
	After 12h	0.450	5.2	0.0450
	After 24h	0.430	9.5	0.0430

TABLE IV

MEASURED IMPACT OF POLISHING ON BANDPASS FILTER IN-BAND INSERTION LOSS AND EXTRACTED UNLOADED QUALITY FACTOR. POSITIVE DIFFERENCE Δ MEANS VALUE IMPROVEMENT WITH RESPECT TO PRE-POLISHING

		f_0 (GHz)	BW (GHz)	IL @ f_0 (dB)	ΔIL @ f_0 (%)	Q_U @ f_0 (Eq. 4)	ΔQ_U (%)
BPF2-42	EM	23.93	1.95				
	Pre	24.8	1.69	0.67		190.2	
	24h			0.65	9.5	196.1	3.1
	48h			0.62	15.4	205.6	8.1
BPF6-42	EM	23.05	2.4				
	Pre	22.97	2.07	2.02		142.8	
	24h			1.27	39.7	227.4	59.2
	48h			1.24	38.3	233.7	63.6
BPF6-28	EM	38.13	4.35				
	Pre	38.46	4.04	1.42		160.8	
	12h			1.27	10.5	195.3	21.4
	24h			1.2	15.5	207.2	28.8

37% are obtained for orientations perpendicular and parallel to layer arrangement, respectively. Considering the above, various polishing times were applied to the fabricated WG demonstrators for a better assessment of polishing time versus loss reduction assessment.

B. Electrical Performance

Finally, the electrical properties of the additively manufactured test vehicles were assessed. The dc conductivity of the specimen was measured using the Hioki 3522-50 LCR meter to be 0.96 MS/m which is 71% of the bulk value for the SS316L. This was followed by measuring the scattering parameters of the test vehicles using the Agilent PNA N5224A vector network analyzer at an IF bandwidth of 1 kHz. The calibrated reference plane is set at the WG flange, using WR-90, WR-42, and WR-28 Thru-Reflect-Line cal-kits. Measurements were taken before polishing, and once after each 24 h (12 h) polishing run. Since very low loss differences were to be measured, the S_{21} measurement uncertainty of the VNA being 0.02 dB [52] cannot be overlooked, and thus a strict protocol was followed. To ensure repeatable results, each time all the screws that force mating the WG flanges were tightened with the same torque force. Moreover, after calibration, the cables connecting VNA with the coax-to-WG adapters were kept in the same position. The measured frequency response of the pre-polished circuits is provided for the WG section in Fig. 9, while for bandpass filters in

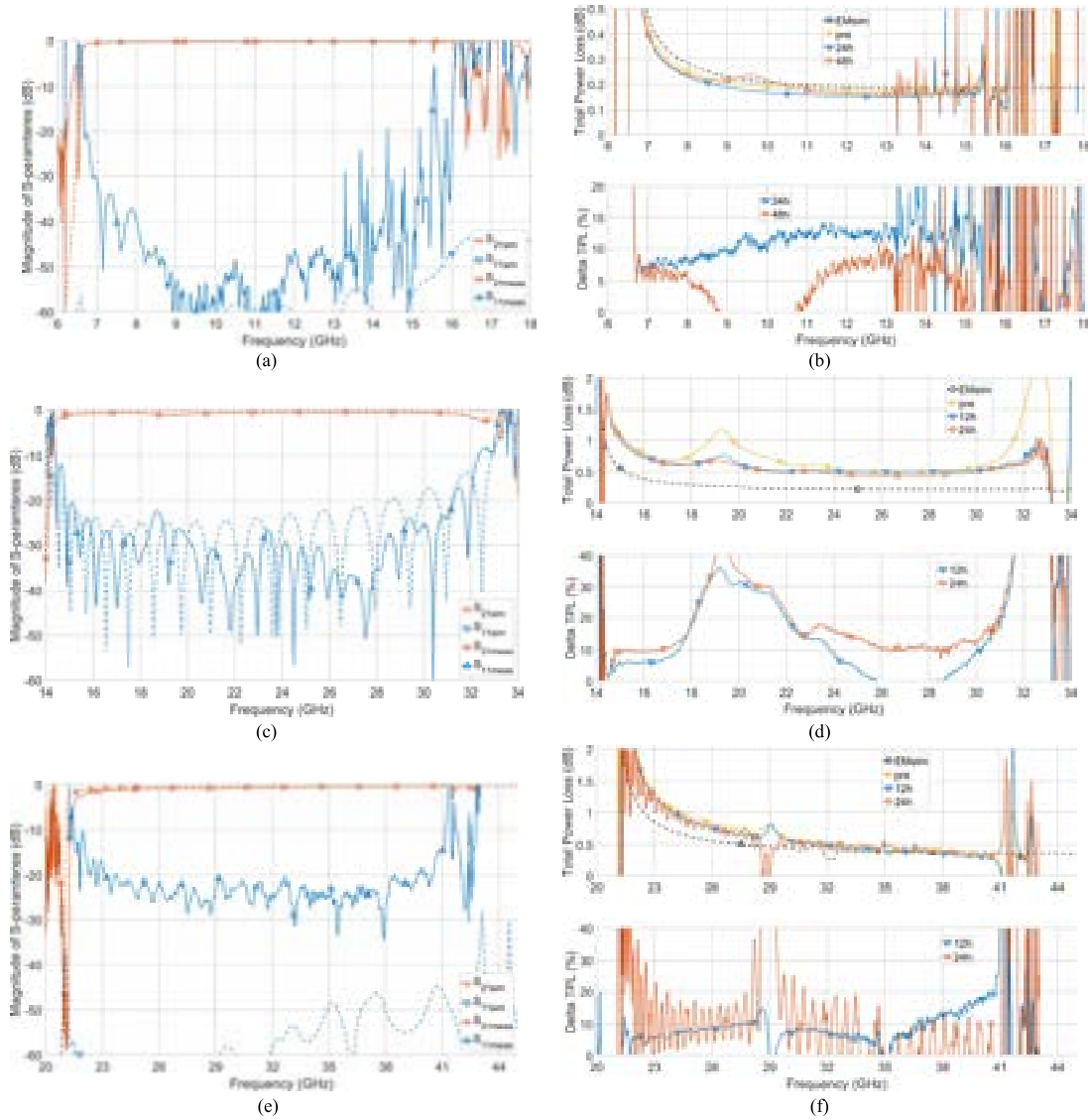


Fig. 9. Measured (solid lines) pre-polishing S-parameters of the SLM-fabricated (a) WR-90, twisted (c) WR-42, and (e) WR-28 sized WG sections along with EM simulated ones overlaid (dashed lines). Measured impact of polishing on TPL of (b) WR-90, (d) WR-42, and twisted (f) WR-28 sized WG sections: pre-polishing, after one and two rounds of polishing. Delta TPL referenced to pre-polishing TPL.

Fig. 10. Moreover, the attenuation constant expressed through the TPL is defined as $TPL = 1 - (|S_{21}|^2 + |S_{11}|^2)$ for both WG sections and BPFs is confronted with the full-wave simulated EM to assess the polishing effectiveness on the power loss. Finally, the percentage improvement after each round of polishing is determined. The results are also summarized in Table III for WGs and in Table IV for BPFs. From the above results, the following conclusions can be drawn. First, the VNA measurement uncertainty has a visible impact on the experimental results which manifests itself in, e.g., erroneous percentage improvement as the in-process TPL values are relatively small. Second, it is seen that a slight dimension reduction is observed due to metal printed part shrinkage. This can be easily counteracted by slightly scaling up the model before printing. Third, for the proposed process, the polishing time is selected as a tradeoff between potential roughness reduction and the time required. It is seen that even after only 12 h of polishing a dozen percent

improvement is seen with longer times the gain being less and less significant. Fourth, glass bead size must be selected considering feature size since the medium must freely penetrate cavities.

The difference is seen, e.g., between loss improvement for WR-42 geometry (even in a twist configuration) versus WR-28. In addition, in BPF6_28 the minimal iris is only an order of magnitude larger than the bead radius. For higher bands smaller beads (e.g., 150–200 μm radius) could yield better results. Fifth, since a rough surface is inevitably bounded by the fabrication technique, the use of different base materials such as aluminum (roughly 30 times better conductivity) powder in combination with the proposed post-processing method would provide a very low loss metal 3-D printed mm-wave components.

Finally, results obtained in this study for narrowband filters were compared with the state-of-the-art. A summary is provided in Table V from which it can be noted that the

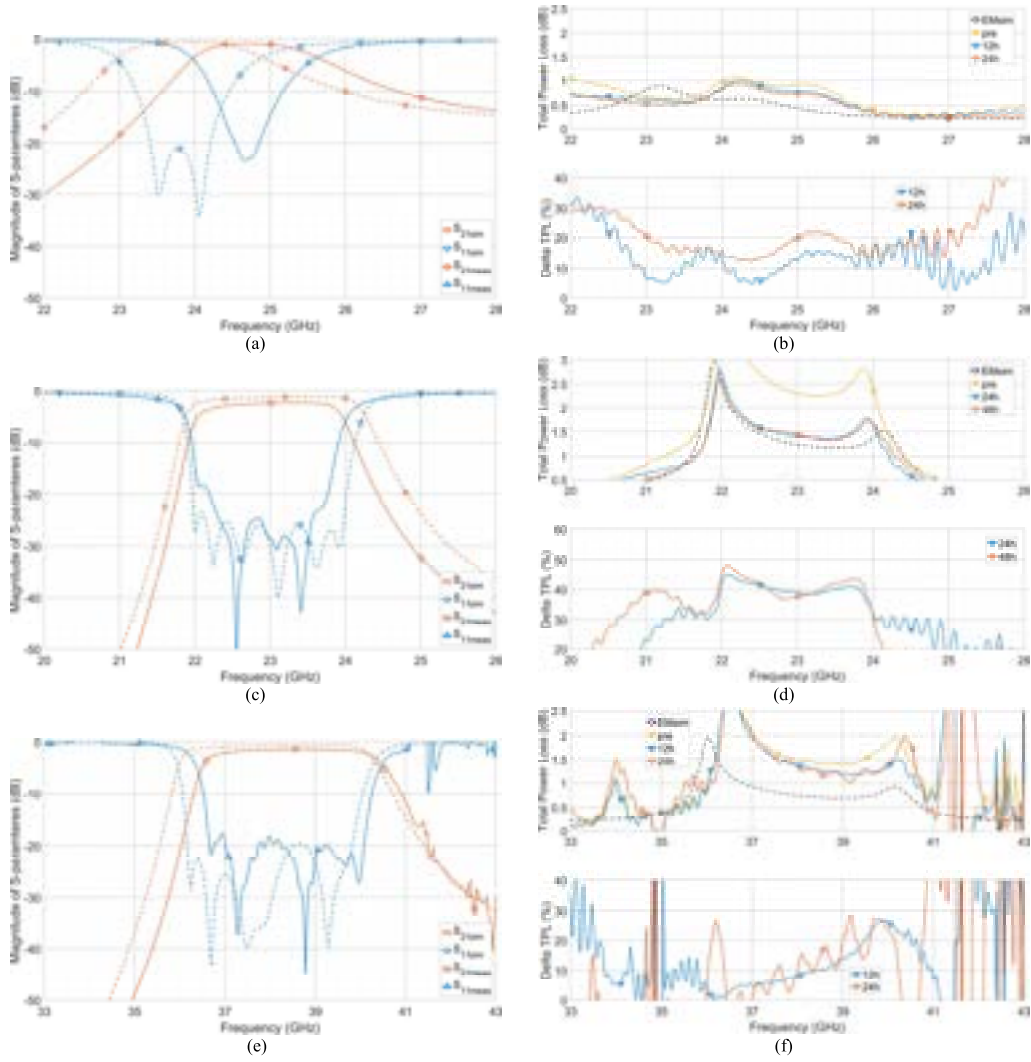


Fig. 10. Measured (solid lines) pre-polishing S-parameters of the SLM-fabricated (a) BPF2_42, (c) BPF6_42, and (e) BPF6_28 filters along with EM simulated ones overlaid (dashed lines). Measured impact of polishing on TPL of (b) BPF2_42, (d) BPF6_42, and (f) BPF6_28 sized WG sections: pre-polishing, after one and two rounds of polishing.

TABLE V
STATE-OF-THE-ART ADDITIVELY MANUFACTURED FILTERS
VERSUS PROPOSED FABRICATION SCHEME

Filter Order/Geometry	3D Technology	f_0 (GHz)	BW (%)	$IL @ f_0$ (dB)
[52] 4 th -order filter/groove gap waveguide	PolyJet	35.65	1.4	~0.5
[53] 5 th -order filter/rectangular waveguide cavities	SLA [#] copper plated	87.5	11.5	~0.5
[54] 9 th -order filter with shaped resonators	SLM Scalmalloy	~12	~5	~1.3
[55] 5 th -order H-plane filter with irises	MLS [*] stainless steel	88	12	~2
This 6 th -order filter with shaped resonators	Polished SLM stainless steel	23	9	~1.2
This 6 th -order filter with shaped resonators	polished SLM stainless steel	38.45	10.5	~1.2

[#]SLA – Stereolithography, ^{*}MLS – Micro Laser Sintering.

proposed fabrication and processing approach provides a very good insertion loss performance even when using a relatively low conductivity metal.

IV. CONCLUSION

A low-cost polishing method dedicated to all-at-once fabricated WG circuits that employ mechanical barrel tumbling of metal additively manufactured parts, using glass microbeads as an abrasive medium, was introduced, and experimentally validated. In contrast to electropolishing, neither chemicals nor sophisticated equipment is necessary, while similar benefits of good internal cavities and channel penetration and polishing are still provided. A set of test vehicles was fabricated in SS316L using the SLM technique. Both mechanical and electrical properties were assessed before polishing and after one and two polishing rounds. In practice, electropolishing reduces a part's R_a by 10%–30% depending on the starting finish. In the case of the proposed method, improvement is in the range of 6%–23% in terms of surface finish and 14%–33% in terms of roughness. The proposed method's results are comparable to those of industry-standard high-cost high-performance techniques. The obtained results translate into a 10%–20% loss reduction within tested WG geometries and finally between 10% and 40% loss reduction for the tested

second- and sixth-order filters. Therefore, the experimental results certified the performance and applicability of the proposed metal 3-D printed WG components post-processing method that the AM technology is well suited for cm- and mm-wave component fabrication.

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Chapter 5

CONCLUSION

5.1. Summary

The main objective of this work was to overcome the existing limitations in EM modeling by enhancing the conventional EM CAD tools with the free-form shape deformation of 3D solids. The need for improvement in this area was identified in Chapter 2, where the current limitations of EM design were outlined against the backdrop of potential opportunities arising from the emerging fabrication technologies, such as additive manufacturing. A detailed review of the latest 3D-printed passive microwave components revealed that, despite extensive research over the last decade, the enormous potential of AM techniques has not yet been fully exploited in the field of microwave electronics. Therefore, in the course of this work, a new method for the design and optimization of passive microwave devices was introduced and applied in practice to address this research gap. The proposed modeling technique based on multivariate radial basis function interpolation was integrated with a full-wave EM field solver and successfully tested on several working examples, as described in Chapter 3. The proposed design toolkit was subsequently used to develop AM-oriented microwave structures having unique shape and smooth surface profile. In total, ten novel microwave filter designs were developed and described in detail in Chapter 4, eight of them were featured in the accompanying publications [MB1-MB5]. All designs were experimentally verified with 3D-printed prototypes. Furthermore, selected practical aspects of the 3D printing process for high-frequency electronics were also addressed in publication [MB6], and summarized in Section 4.5. The obtained results fully confirmed the effectiveness of the proposed techniques, thereby validating the claims put forward in the dissertation:

1. A free-form shape deformation technique based on radial basis function interpolation opens up new design possibilities and offers additional degrees of freedom in the 3D modeling of microwave electronic components,

The usefulness of the proposed shape deformation method to high-frequency electronic design was demonstrated in practice in [MB1, MB2] and described in more detail in Sections 3.2 and 4.1. Based on the design of microwave resonators and filters in [MB1, MB2], it was shown how the new approach based on FFD can expand current design capabilities in EM CAD. The proposed deformation of 3D solids can be easily integrated with established CAD tools and provides increased design flexibility by allowing shapes to be freely modified through arbitrarily defined control points. For

instance, the designs of microwave resonators and filters with squeezed cavities from [MB2] or the deformed dual-mode cavity models in circular [MB3] or rectangular waveguide [MB4] could not have been developed, and their shape optimized, using existing EM modeling software without employing the proposed design framework.

2. Using shape deformation in EM design allows for exploiting the potential of additive manufacturing processes in the fabrication of microwave devices with complex shapes and smooth profiles,

The practical significance of the design possibilities provided by free-form modeling techniques lies in their usability, which in the case of microwave circuits comes down to feasibility of manufacturing. As discussed in Section 2.3, although AM offers great flexibility in printed model complexity, it has not been fully addressed and exploited in EM design up until now. The design technique introduced in this work aims to bridge that gap and facilitate truly AM-oriented design of microwave components, as demonstrated by the 3D-printed components presented in Chapter 4. In particular, the smooth-profile filter designs in a rectangular waveguide [MB4] and air-filled coaxial-line [MB5] should be especially recognized as examples of truly AM-oriented structures, taking full advantage of the capabilities of 3D printing. Moreover, such complex models and geometries could only be fabricated by means of AM, so these two methods (free-form deformation and 3D printing technology) complement each other.

3. Implementing free-form deformation in the design process, combined with fabrication by additive manufacturing, can lead to high-frequency components with improved RF performance,

The validity of this claim has been demonstrated by a number of innovative designs described in Chapter 4. The circuits presented in [MB2, MB3, MB4, MB5] were verified by manufactured prototypes, which confirmed their superior RF performance over conventional designs currently used in the industry. The developed components are characterized by enhanced electrical parameters, such as low loss, high quality factor Q , or wide spurious mode separation. The new AM-oriented devices can also greatly contribute to the miniaturization and higher integration of RF systems, leading to multifunctional designs such as the filtering waveguide twist in [MB2]. This performance improvement is exemplified in article [MB3], where the proposed design methodology was applied to design channel filters for OMUX in satellite communication systems. The study aimed to achieve a wider stopband so that more channels could be accommodated without supplementary lowpass filters. The implemented solution resulted in a 35% improvement in stopband rejection beyond the established state of knowledge. In [MB4], the above-mentioned method was used to design a new class of waveguide filters with dual-mode cavities. The developed smooth-surface structures are characterized by very low losses for components fabricated by 3D printing technology. What is more, the proposed resonant cavity design allows for the coupling of two fundamental modes by means of deformation of the cavity itself, without the need for additional coupling elements (e.g.,

screws). Finally, the effectiveness of the presented design approach was also confirmed in [MB5], where it was verified on a different technology—an air-filled coaxial line. The fabricated component exhibits the widest stopband rejection among the related state-of-the-art solutions, meeting the specification demanded by current Earth observation satellite systems.

4. By combining advanced 3D-printing processes with appropriate post-processing, it is possible to produce microwave hardware whose performance is comparable to that of components manufactured using established processes.

As it was discussed in Section 4.5, one of the key factors affecting the performance of 3D-printed circuits is the roughness of the metal surface. Since the beginning of its use in microwave engineering, additive manufacturing has suffered from relatively high surface roughness, which resulted in limited application of this technology, especially at higher frequencies such as the millimeter-wave range. Therefore, along with the ongoing rapid development of more precise 3D printing processes for specialized applications, the topic of improving surface quality of printed parts has also become the subject of research. In [MB6], it was shown how the surface roughness of 3D-printed waveguide components can be improved in post-processing, using a low-cost mechanical polishing method. Additionally, the latest designs developed in [MB4, MB5] were fabricated using AM processes specifically tailored for RF, which yielded significantly better results in terms of estimated metal conductivity (e.g., 10 MS/m for the eighth-order BPF in [MB4]). The study in [MB6], along with the results obtained with specialized AM techniques, together prove that powder-based AM methods, when combined with suitable post-processing, are capable of producing microwave components of surface quality approaching that of traditional manufacturing techniques such as milling or turning. Moreover, continued advances in high-precision 3D printing and its broader adoption in microwave engineering are expected to further narrow that technological gap in the coming years.

In summary, the result of the research conducted in this doctoral dissertation is a new integrated method for designing passive microwave circuits adapted for 3D printing. The proposed method allows full use of the capabilities of 3D printing by designing circuits with smooth profiles and irregular shapes. The devices described in the accompanying publications exhibit superior electrical parameters (such as stopband rejection and insertion loss) than previously presented solutions. Part of the completed investigation was carried out under ESA-funded research programs and concerned areas with practical application in SATCOM systems. It is also worth noting that a further contribution of this work is the commercial integration of the developed toolkit into a professional full-wave EM simulation environment InventSim. The proposed framework is currently unique in the field of microwave engineering, universal, and can be used to design a wide range of high-frequency electronic components.

5.2. Outlook and prospects for future research

Future research can focus on the development of other types of passive microwave circuits—such as power dividers, couplers, antennas—taking advantage of the flexibility offered by shape deformation and additive manufacturing. Furthermore, as recognized in [5], the unique features of 3D printing combined with free-form modeling can be employed to design highly integrated, multifunctional components, such as *filtennas* (i.e., integrated filtering antennas), up to entire RF front-ends and antenna feed networks for space applications.

Concurrently, as a possible extension of this work, future activities may concentrate on the designs developed within this thesis, such as the novel filter based on a 3D-printed coaxial line [MB5]. Future studies on these solutions should address additional aspects such as thermal behavior, power-handling capability, and possible integration with antennas, in order to better explore the potential of their application in modern SATCOM systems.

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2. **M. Baranowski**, Ł. Balewski, A. Lamecki and M. Mrozowski, "Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators," *IEEE Access*, vol. 12, pp. 78275-78285, 2024,
3. J. Sorocki, I. Piekarz, **M. Baranowski**, A. Lamecki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024,
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5. **M. Baranowski**, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication

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6. **M. Baranowski**, G. Fotyga, A. Lamecki, and M. Mrozowski, "Bayesian Optimization for solving high-frequency passive component design problems", *Bulletin of the Polish Academy of Sciences-Technical Sciences*, vol. 70, no. 4, 2022.

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4. 53rd European Microwave Conference (EuMC), European Microwave Week 2023, Berlin, Germany, title: "Rectangular Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators",
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5. J. Sorocki, I. Piekarz, M. Baranowski, A. Lamecki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024,
6. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," in *IEEE Microwave and Wireless Technology Letters*, vol. 35, no. 9, pp. 1292-1295, Sept. 2025;

I contributed to the publications listed above by providing overall scientific supervision and guidance throughout the research process in my role as his PhD supervisor. My contributions included advising



on research design and methodology, discussing and interpreting results, providing critical feedback during manuscript preparation, and reviewing and editing the final versions of the manuscripts prior to submission. I also contributed to securing funding for the research. The experimental and analytical work was primarily conducted by Mr. Michal Baranowski.

A handwritten signature in blue ink, appearing to read "Adam Lanyon", with a long horizontal flourish extending to the right.

signature

(author name and affiliation)

.....
date

Michał Mrozowski, Gdansk University of Technology

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, Ł. Balewski, A. Lamecki and M. Mrozowski, "Fast Design Optimization of Waveguide Filters Applying Shape Deformation Techniques," *2022 24th International Microwave and Radar Conference (MIKON)*, Gdansk, Poland, 2022, pp. 1-4, doi: 10.23919/MIKON54314.2022.9924969,
2. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication Systems," in *IEEE Microwave and Wireless Components Letters*, vol. 32, no. 12, pp. 1403-1406, Dec. 2022,
3. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques," in *IEEE Transactions on Microwave Theory and Techniques*, vol. 71, no. 7, pp. 3065-3074, July 2023,
4. M. Baranowski, L. Balewski, A. Lamecki and M. Mrozowski, "Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators," in *IEEE Access*, vol. 12, pp. 78275-78285, 2024,

I declare that my contribution to their creation was as follows:

- monitoring of research work
- editing of the draft publication / corrections after receiving review /contribution to the reponses
- Thorough review of the manuscript



Signed by / Podpisano przez:

Michał Piotr Mrozowski.....
Politechnika Gdańska

Date / Data: 2024-11-17 15:06
signature

dr inż. Łukasz Balewski
EM Invent Sp. z o.o.

19.11.2025

date

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, Ł. Balewski, A. Lamecki and M. Mrozowski, "Fast Design Optimization of Waveguide Filters Applying Shape Deformation Techniques," *2022 24th International Microwave and Radar Conference (MIKON)*, Gdansk, Poland, 2022, pp. 1-4, doi: 10.23919/MIKON54314.2022.9924969,
2. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication Systems," in *IEEE Microwave and Wireless Components Letters*, vol. 32, no. 12, pp. 1403-1406, Dec. 2022,
3. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques," in *IEEE Transactions on Microwave Theory and Techniques*, vol. 71, no. 7, pp. 3065-3074, July 2023,
4. M. Baranowski, Ł. Balewski, A. Lamecki and M. Mrozowski, "Low-Loss 3D-Printed Waveguide Filters Based on Deformed Dual-Mode Cavity Resonators," in *IEEE Access*, vol. 12, pp. 78275-78285, 2024,

I declare that my contribution to their creation was as follows:

- implementation of radial basis function interpolation in the InventSim software, integration of shape deformation features with 3D finite-element method, later development of the FFD tool used in the design process as described in the publications.
- editorial and substantial consultation of the publication.

Łukasz Balewski

signature

.....
date

Jaione Galdeano
Microwave engineer
RF Equipment and Technology Section (TEC-EFE)
ESA-ESTEC
Keplerlaan 1, PO Box 299
2200 AG Noordwijk, The Netherlands

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "A Circular Waveguide Dual-Mode Filter With Improved Out-of-Band Performance for Satellite Communication Systems," in *IEEE Microwave and Wireless Components Letters*, vol. 32, no. 12, pp. 1403-1406, Dec. 2022.
2. M. Baranowski, Ł. Balewski, A. Lamecki, M. Mrozowski and J. Galdeano, "The Design of Cavity Resonators and Microwave Filters Applying Shape Deformation Techniques," in *IEEE Transactions on Microwave Theory and Techniques*, vol. 71, no. 7, pp. 3065-3074, July 2023

I declare that my contribution to their creation was as follows:

- Substantive consultation
- editorial corrections

Jaione
Galdeano

Digitally signed
by Jaione
Galdeano
Date: 2025.04.08
09:37:41 +02'00'

.....
signature

05 October 2025

Dr. Alejandro Pons Abenza
Public University of Navarre

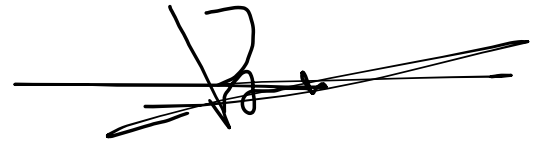
Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- Theoretical synthesis of the initial design.
- Validation of the prototypes.
- Fabrication of the prototypes.
- Measurement of the prototypes.
- Preparation for the publication.

A handwritten signature in black ink, consisting of a large, stylized 'A' followed by a horizontal line and a small flourish.

.....
signature

Germán Andrés Álvarez Botero
Universidad Pública de Navarra (UPNA)
Spain

August 19, 2025.

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- Participation in the idea conception, measurements, and academic discussions throughout the elaboration and review process.

ALVAREZ BOTERO
GERMAN ANDRES
- Y7822461Z

Digitally signed by ALVAREZ
BOTERO GERMAN ANDRES -
Y7822461Z
Date: 2025.08.19 12:28:20
+05'00'

Germán Andrés Álvarez Botero

June 30, 2025

Iván Arregui

Public University of Navarre

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- Participating in the design of a prototype component.
- Assisting with the planning and measurement process.
- Contributing to the drafting and writing of the article.

ARREGUI
PADILLA IVAN -
DNI 72694359F

Firmado digitalmente
por ARREGUI PADILLA
IVAN - DNI 72694359F
Fecha: 2025.06.30
12:43:24 +02'00'

Ivan Arregui

.....July 4, 2025....

(author name and affiliation)

date

...Txema Lopetegi.....

...Public University of Navarre.....

...Pamplona. Navarre. Spain.....

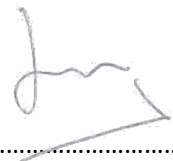
Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- Participation in the synthesis of the initial filter design.
- Participation in the development of the overall strategy to design the new proposed filter.
- Thorough review of the manuscript.



.....
signature

.....June 30th, 2025....

(author name and affiliation)

date

Miguel A. G. Laso, UPNA.....

.....
.....

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- Monitoring of the research work.
- Supervision during the writing of the article.

A handwritten signature in blue ink is written over a circular stamp. The stamp contains the text 'upna' in a stylized font, followed by 'Unidad de Promoción de la Investigación' and 'Departamento de Ingeniería de Telecomunicaciones'.

.....

signature

Leiden, The Netherlands

Petronilo Martin-Iglesias

European Space Agency and Public University of Navarre

20.08.2025

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. M. Baranowski, A. Pons-Abenza, I. Arregui, T. Lopetegi, G. Alvarez-Botero, A. Lamecki, M. Angel Gomez Laso, and P. Martin-Iglesias, "A 3-D-Printing-Oriented Coaxial-Line Filter With Wide Out-of-Band Rejection," IEEE Microwave and Wireless Technology Letters (Early Access), 2025, doi: 10.1109/LMWT.2025.3578244.

I declare that my contribution to their creation was as follows:

- preparation of the publication and substantial consultation.

A handwritten signature in blue ink, appearing to read 'Petronilo Martin-Iglesias', is written over a faint horizontal line.

Jakub Sorocki
Institute of Electronics
AGH University of Krakow, Poland
jakub.sorocki@agh.edu.pl

Krakow, 25.06.2025

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. J. Sorocki , I. Piekarz , M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- Conceptualization of the study
- Coordination of the research group
- Development of the polishing method and experimental setup
- Design of some test vehicles
- Electrical/mechanical measurements
- Data analysis
- Writing the publication draft



signature

Ilona Piekarz
Institute of Electronics
AGH University of Krakow, Poland
ilona.piekarz@agh.edu.pl

Krakow, 25.06.2025

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

1. J. Sorocki, I. Piekarz, M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- Development of the polishing method and experimental setup
- Electrical/mechanical measurements
- Data analysis
- Writing the publication draft



signature

Pavia, 27/06/2025

To whom it may concern

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

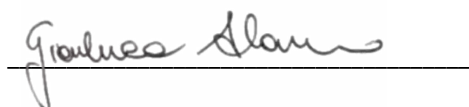
1. J. Sorocki , I. Piekarz , M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- manufacturing of the prototypes
- preparation of the publication

Sincerely,

Gianluca Alaimo



*Scientific and technological sector Officer
Department of Civil Engineering and Architecture,
University of Pavia
Via Ferrata 3, 27100 Pavia, Italy*

*Email: gianluca.alaimo@unipv.it
Phone: +39 0382 985468*

Pavia, 27/06/2025

To whom it may concern

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

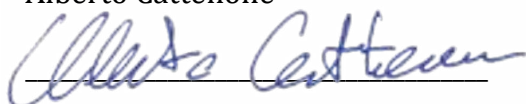
1. J. Sorocki , I. Piekarz , M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- fabrication of the prototypes
- preparation of the publication

Sincerely,

Alberto Cattenone



*Scientific and technological sector Officer
Department of Civil Engineering and Architecture,
University of Pavia
Via Ferrata 3, 27100 Pavia, Italy*

*Email: alberto.cattenone@unipv.it
Phone: +39 0382 985468*

Pavia, 27/06/2025

To whom it may concern

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publications:

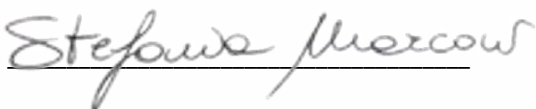
1. J. Sorocki, I. Piekarz, M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- fabrication of the prototypes
- preparation of the publication

Sincerely,

Stefania Marconi



*Associate Professor of Bioengineering
Department of Civil Engineering and Architecture,
University of Pavia
Via Ferrata 3, 27100 Pavia, Italy*

Email: stefania.marconi@unipv.it

Phone: +39 0382 985468



Pavia, Italy

26/06/2025

To whom it may concern.

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publication:

J. Sorocki, I. Piekarz, M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," IEEE Transactions on Microwave Theory and Techniques, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- Preliminary discussion of the problem
- Support in the manufacturing process and measurements

Sincerely,

Nicolò Delmonte

PhD, University of Pavia

Dept. of Electrical, Computer and Biomedical Engineering

Via Ferrata, 5 - 27100 Pavia (Italy)

phone: +39 0382 985223



Pavia, Italy

26/06/2025

To whom it may concern.

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publication:

J. Sorocki, I. Piekarz, M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," IEEE Transactions on Microwave Theory and Techniques, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- Preliminary discussion of the problem
- Support in the manufacturing process and measurements

Sincerely,

Lorenzo Silvestri

PhD, University of Pavia

Dept. of Electrical, Computer and Biomedical Engineering

Via Ferrata, 5 - 27100 Pavia (Italy)

phone: +39 0382 985223



UNIVERSITÀ DI PAVIA
Dipartimento di
Ingegneria Industriale
e dell'Informazione

Prof. Maurizio Bozzi

Pavia, 25 June 2025

Address: Via Ferrata, 5
27100 Pavia, Italy
Phone: +39 0382 985 782
Fax.: +39 0382 422 583
e-mail: maurizio.bozzi@unipv.it

To whom it may concern.

Statement of my contribution to the joint publication with Michał Baranowski

Referring to the following publication:

J. Sorocki, I. Piekarz, M. Baranowski, A. Lamęcki, A. Cattenone, S. Marconi, G. Alaimo, N. Delmonte, L. Silvestri, M. Bozzi, "Low-Cost Method for Internal Surface Roughness Reduction of Additively Manufactured All-Metal Waveguide Components," *IEEE Transactions on Microwave Theory and Techniques*, vol. 72, no. 8, pp. 4519-4529, Aug. 2024.

I declare that my contribution to their creation was as follows:

- Preliminary discussion of the problem
- Preparation of the publication

Sincerely,

Prof. Maurizio Bozzi, Ph.D., IEEE Fellow

Full Professor, University of Pavia
2024 President, IEEE Microwave Theory and Technology Society (MTT-S)

Dept. of Electrical, Computer and Biomedical Engineering
Via Ferrata, 5 - 27100 Pavia (Italy)
phone: +39 0382 985 782 - web: <http://microwave.unipv.it/bozzi>