ESCUELA TÉCNICA SUPERIOR DE INGENIERÍA DE
TELECOMUNICACIÓN
UNIVERSIDAD POLITÉCNICA DE CARTAGENA, REGIÓN DE
MURCIA

Tesis doctoral

**Diseño e implementación de técnicas disruptivas de filtros
resintonizables en banda Ku/Ka**

Design and Implementation of disruptive techniques of Ku/Ka band tunable filters

Autor: Antonio Romera Pérez
Directores: Fernando Quesada Pereira
Alejandro Melcón Álvarez
Supervisora: Lara Arche Andradás
Cartagena, Enero 2020

**CONFORMIDAD DE SOLICITUD DE AUTORIZACIÓN DE DEPÓSITO DE
TESIS DOCTORAL POR EL/LA DIRECTOR/A DE LA TESIS**

D. Fernando D. Quesada Pereira, Director de la Tesis doctoral y Alejandro Melcón Álvarez, Co-director de la tesis doctoral Diseño e implementación de técnicas disruptivas de filtros resintonizables en banda Ku/Ka.

INFORMA:

Que la referida Tesis Doctoral, ha sido realizada por D/D^a. Antonio Romera Pérez, dentro del Programa de Doctorado Tecnologías de la información y las comunicaciones, dando mi conformidad para que sea presentada ante el Comité de Dirección de la Escuela Internacional de Doctorado para ser autorizado su depósito.

- ☒ Informe positivo sobre el plan de investigación y documento de actividades del doctorando/a emitido por el Director/ Tutor (**RAPI**).

La rama de conocimiento en la que esta tesis ha sido desarrollada es:

- ☐ Ciencias
☐ Ciencias Sociales y Jurídicas
☒ Ingeniería y Arquitectura

En Cartagena, a 5 de diciembre de 2019

EL DIRECTOR/A DE LA TESIS

Fdo.: Fernando D. Quesada Pereira

EL CODIRECTOR DE LA TESIS

Fdo.: Alejandro Melcón Álvarez

COMITÉ DE DIRECCIÓN ESCUELA INTERNACIONAL DE DOCTORADO

CONFORMIDAD DE DEPÓSITO DE TESIS DOCTORAL
POR LA COMISIÓN ACADÉMICA DEL PROGRAMA

D/D^a. Jorge Larrey Ruiz, Presidente/a de la Comisión Académica del Programa Tecnologías de la información y las comunicaciones.

INFORMA:

Que la Tesis Doctoral titulada, “Diseño e implementación de técnicas disruptivas de filtros resintonizables en banda Ku/Ka”, ha sido realizada, dentro del mencionado Programa de Doctorado, por D/D^a. Antonio Romera Pérez, bajo la dirección y supervisión del Dr./ Dra. Fernando D. Quedada Pereira.

En reunión de la Comisión Académica, visto que en la misma se acreditan los indicios de calidad correspondientes y la autorización del Director/a de la misma, se acordó dar la conformidad, con la finalidad de que sea autorizado su depósito por el Comité de Dirección de la Escuela Internacional de Doctorado.

- ☒ Evaluación positiva del plan de investigación y documento de actividades por el Presidente de la Comisión Académica del programa (**RAPI**).

La Rama de conocimiento por la que esta tesis ha sido desarrollada es:

- ☐ Ciencias
- ☐ Ciencias Sociales y Jurídicas
- ☒ Ingeniería y Arquitectura

En Cartagena, a 5 de Diciembre de 2020

EL PRESIDENTE DE LA COMISIÓN ACADÉMICA

**JORGE|
LARREY|
RUIZ**

Firmado
digitalmente por
JORGE|LARREY|RUIZ
Fecha: 2019.12.09
12:19:21 +01'00'

Fdo: Jorge Larrey Ruiz

COMITÉ DE DIRECCIÓN ESCUELA INTERNACIONAL DE DOCTORADO

“La regularidad puede influir al cambio positivo. Siempre hay que buscar retos. Hallar nuevas preguntas. Tener el valor de pensar en grande y confiar en tu potencial y en tus decisiones”

Gary Kasparov

Agradecimientos

He de reconocer que cuando entre a la universidad nunca pensé que haría una tesis doctoral, ni siquiera cuando estaba en los últimos años. Esta posibilidad solo me la planteé tras realizar mi proyecto de fin de carrera con mis ahora directores de tesis Fernando y Alejandro. Gracias a ellos me interese por todo lo relacionado con la investigación y por esta rama de las telecomunicaciones. Gracias a sus consejos, sus ideas y las numerosas horas que me han dedicado a lo largo de estos años me ha sido posible realizar este trabajo. Así que por todo esto y por lo que hemos vivido juntos estos años, de verdad muchas gracias por darme la oportunidad de hacer esta tesis con dos grandes personas y profesionales como vosotros.

Esta tesis doctoral ha sido financiada por Thales Alenia Space España. Quiero darle las gracias a Ana y Santiago por darme la oportunidad de realizar la tesis con vosotros, gracias a esto me ha sido posible conocer la perspectiva de la empresa que de otra forma no habría conocido. Y tengo que darle especialmente las gracias a Lara, que ha sido nuestro enlace con la empresa todos estos años y la supervisora del trabajo. Quiero agradecerle las grandes aportaciones que has hecho al trabajo durante las largas reuniones que hemos tenido a lo largo de estos años. Eres una grandísima persona y para mí ha sido realmente un placer trabajar contigo.

He de hacer una mención especial a Alejandro, mi compañero de tesis y sin duda alguna uno de mis mejores amigos. Ha sido un alguien clave en esta etapa de la realización de la tesis, tanto en las largas charlas que hemos tenido hablando de filtros como los ratos de ocio. Muchas gracias por todo lo que hemos compartido.

Por último, pero no menos importante quiero agradecerle a mi familia todo el apoyo que me ha dado y sin el cual no podría haber llegado hasta aquí. A lo largo de estos años han pasado bastantes cosas, pero aun así aquí estamos. Quiero agradecer especialmente a mis padres, ellos me animaron a realizar la tesis cuando yo aun tenía dudas y sin ellos me hubiera sido imposible. También mencionar a mi hermana Mari, a mis sobrinos Patricio y Adrián, y a mi tía Lali por todo el apoyo que me han dado siempre. ¡Muchas gracias!

Resumen

El desarrollo de filtros de microondas resintonizables es un requisito para los nuevos sistemas de comunicación. El mercado demanda sistemas flexibles, que permitan adaptar su funcionamiento a los nuevos requerimientos del servicio. En los sistemas de comunicación, los filtros son una parte fundamental cuya finalidad es transmitir las frecuencias seleccionadas, mientras que el resto se eliminan. Sin embargo, los filtros de microondas sintonizables requieren componentes adicionales que proporcionan la capacidad de cambiar la frecuencia de funcionamiento y las características de la banda de paso. Por lo tanto, en esta tesis doctoral se presenta todo el proceso de diseño, desde el análisis inicial de los resonadores hasta el diseño final de nuevas estructuras de filtro sintonizables.

Las principales tecnologías exploradas en este trabajo son la guía de onda integrada en sustrato (SIW por sus siglas en inglés) y la guía de onda rectangular convencional. El primero se selecciona debido a la fácil integración de componentes electrónicos, como diodos varactores, condensadores, etc., que son necesarios para la sintonía. El segundo es el preferido para aplicaciones que requieren un factor de calidad alto (pocas pérdidas de inserción) y aplicaciones que requieren un manejo de alta potencia, como en los sistemas de comunicación por satélite. Sin embargo, la integración de los componentes de sintonía en las estructuras de las guías de onda es siempre una tarea difícil. Por lo tanto, esto implica añadir algunas estructuras adicionales, por ejemplo, sustratos como soporte para componentes electrónicos. Otra forma de sintonizar los filtros de guía de onda es tener algunos actuadores mecánicos, como motores paso a paso o actuadores piezoeléctricos.

Los filtros desarrollados en esta tesis doctoral se centran principalmente en la sintonía eléctrica para la banda Ku y la sintonía mecánica para la banda Ka. De hecho, esto se debe a las limitaciones de los diodos varactores a altas frecuencias en términos de rango de capacidad, altas pérdida. Para estudiar los filtros con capacitancias variables, hemos encontrado interesante analizarlos en dos etapas. La primera etapa está compuesta por una capacidad distribuida a lo largo de una superficie, que conecta completamente las dos partes metálicas involucradas en la sintonía. Como ventaja

importante, las simulaciones de onda completa presentan una mejor convergencia lo que nos lleva a simulaciones más rápidas, lo que permite un análisis más detallado de los resonadores o filtros en estudio. Además, en esta etapa, problemas tales como frecuencias resonantes espurias cerca de la banda de paso o problemas con la radiación producida en estos filtros pueden ser resueltos o minimizados de manera efectiva. Después de eso, la segunda etapa es considerar las capacidades como superficies rectangulares que modelan de manera más realista una encapsulación de diodo varactor. En este momento, es crucial asegurar que el rango de capacidad requerido para sintonizar el resonador esté disponible en diodos varactores comerciales para construir un filtro realizable.

Mediante el uso de la sintonía electrónica, se investigan tres tipos de filtros en esta tesis. El primer tipo es el de los filtros SIW evanescentes sintonizables, que presentan un amplio rango de sintonía y unas bajas pérdidas por radiación. También se presenta un filtro de guía de onda rectangular que utiliza el mismo tipo de resonadores. El segundo tipo es un filtro de guía de onda rectangular evanescente sintonizable; estos filtros presentan un rango de sintonía moderado y se debe considerar la posibilidad de insertar componentes eléctricos con la ayuda de sustratos dieléctricos introducidos como parte del filtro. Estos sustratos dieléctricos se han utilizado para dos propósitos diferentes: soporte para los componentes electrónicos y para minimizar los problemas de radiación que presentan en un primer momento estos filtros. El tercer tipo de filtros investigados en este trabajo consiste en utilizar conductores magnéticos artificiales sintonizables (AMC por sus siglas en inglés) fabricados con capacidades variables. Su principio de funcionamiento se basa en la modificación de la anchura efectiva de las guías de onda rectangulares para desplazar sus resonancias. El capítulo correspondiente de este documento muestra un estudio en profundidad de la importancia de los parámetros de los AMCs en el diseño de este tipo de filtros sintonizables.

En cuanto a la sintonía mecánica, hay tres áreas principales de interés: filtros con elementos rotativos, filtros con tornillos móviles y filtros con dieléctricos móviles. Los filtros con elementos rotativos son interesantes, ya que algunos autores destacan la robustez y precisión de los motores paso a paso para la sintonía angular, junto con los rangos de sintonía adecuados que se pueden conseguir. En la segunda área, se investigan algunos filtros evanescentes con tornillos ajustables. Para estos filtros, se muestra una configuración con ancho de banda sintonizable utilizando resonadores no resonantes (resonadores utilizados como acoplamientos). Finalmente, se presentan dos resonadores diferentes con cilindros circulares dieléctricos móviles. Estos dos resonadores presentan un alto rango de sintonía y unos factores de calidad altos; por lo tanto, son muy interesantes para aplicaciones espaciales.

Por último, para las aplicaciones espaciales en las que los efectos adversos de alta potencia como multipactor y corona son muy importantes, algunos filtros de paso bajo pueden ser requeridos junto con algunos filtros sintonizables, cuyos rangos libres de espurias no son lo suficientemente amplios como para ser empleados en sistemas de comunicación satelital por sí mismos. El uso de material dieléctrico en áreas donde es más probable que se produzca multipactor, evita el libre movimiento de los electrones en áreas críticas. Este concepto se aplica en los filtros paso bajo, elevando el umbral de multipactor. Del mismo modo, los materiales magnéticos con una constante de baja permeabilidad (que en realidad no se encuentra en la naturaleza) pueden ser empleados teóricamente para diseñar filtros

de paso bajo, siguiendo los mismos principios de trabajo que los materiales dieléctricos. Sin embargo, debido al uso de estos materiales junto al poste metálico, hace resonar los inversores que componen los filtros de paso bajo. Esto, a su vez, provoca que aparezcan algunos ceros de transmisión debido a la interacción con estos inversores resonantes. En consecuencia, estos ceros de transmisión mejoran tanto la selectividad del filtro como el rango libre de espurios.

Abstract

The development of tunable microwave filters is a requirement for new communications systems. The market demands flexible systems, which allow adapting their operation to the new requirements of the service. In communication systems, the filters are a fundamental part whose purpose is to transmit the selected frequencies, while the rest are removed. However, tunable microwave filters require additional components that provide the ability to change their operating frequency and passband characteristics. Therefore, in this Ph.D. dissertation it is presented the whole design process from the initial analysis of the resonators to the final design of novel tunable filter structures.

The main technologies explored in this work are are substrate integrated waveguide (SIW) and conventional rectangular waveguide. The first one is selected due to the easy integration of electronic components, such as varactor diodes, capacitors, etc., that are required for tuning. The second one is preferred for high-quality factor applications and applications that require high power handling, such as in satellite communication systems. However, the integration of tuning components in waveguide structures is always a difficult task. Thus, this involves adding some additional structures, for instance, substrates as support for electronic components. Another way to tune waveguide filters is to have some drive actuators, such as stepper motors or piezoelectric actuators.

The filters developed in this PhD. thesis are mainly focused on electrical tuning for Ku-band and mechanical tuning for Ka-band. In fact, this is due to the limitations of the varactor diodes at high frequencies in terms of capacitance range, high losses and self-resonances. In order to study filters with variable capacitances, we have found interesting analyzing them in two stages. The first stage is composed of a distributed capacitance along a surface, which fully connects the two metal parts involved in the tuning. As an important advantage, the full-wave simulations require less computational time, leading to faster analysis of resonators or filters under study. Moreover, in this stage, problems such as spurious resonant frequencies near to the passband or RF-leakage can effectively be solved or minimized. After that, the second stage is to consider the capacitances

as rectangular surfaces that models more realistically a varactor diode encapsulation. At this moment, it is crucial to ensure that the capacitance range required for tuning the resonator is available in commercial varactor diodes in order to build a realizable filter.

By using electronic tuning, three kinds of filters are investigated in this thesis. The first kind is tunable evanescent SIW filters, which present a wide tuning range and low RF-leakage. Also, a rectangular waveguide filter that uses the same kind of resonators is presented. The second kind is a tunable evanescent rectangular waveguide filter; these filters present a moderate tuning range and a way to insert electrical components with the aid of dielectric substrates introduced as a part of the filter must be envisaged. These dielectric substrates have been used for two different purposes: support for the electronic components and for minimization of RF-leakage. The third type of filters investigated in this work consists in using tunable artificial magnetic conductors (AMC) made with variable capacitances. Their working principle is based on modifying the effective width of rectangular waveguides in order to shift their resonances. The corresponding section in this document displays an in-depth study of the importance of the parameters of the AMCs in the design of this type of tunable filters.

As regards mechanical tuning, there are three main areas of interest: filters with rotating elements, filter with movable screws and filters with movable dielectrics. Filters with rotating elements are interesting, since some authors remark their robustness and precision of stepper motors for angular tuning, together with suitable tuning ranges that can be achieved. The second area, some evanescent filters with adjustable screws are investigated. For these filters, a configuration with tunable bandwidth is shown by using non-resonating resonators (resonators used as couplings). Finally, two different resonators with movable dielectric circular cylinders are presented. These two resonators display a high tuning range and high-quality factors; thus, they are very interesting for space applications, in which high-quality factors are required.

Finally, for space applications where high power adverse effects like multipactor and corona are very important, some lowpass filters can be required alongside some tunable filters, whose spurious free ranges are not wide enough to be employed in satellite communication systems by themselves. The use of dielectric material in areas where multipactor breakdown is more likely to occur, avoids the free electron movement in critical areas. This concept is applied in lowpass filters, raising the multipactor breakdown threshold. Similarly, magnetic materials with low permeability constant (which actually cannot be found in a nature) can be theoretically employed to design lowpass filters, following the same working principles of the dielectric materials. However, due to the use of these materials alongside the metallic post, makes the inverters, that compose lowpass filters, resonant. This, in turn, provoke that some transmission zeros appear due to the interaction with these resonant inverters. Consequently, these transmission zeros improve the selectivity of the filter while the spurious free range is enhanced.

Contents

1	Introduction	1
1.1	State of the art	1
1.2	Motivation	4
1.3	Objetives and working plan	5
1.4	Thesis outline and relevant results	6
2	Design techniques for bandpass filters	9
2.1	Coupling matrix and basic design technique	9
2.1.1	Design example	11
2.2	Step by step design technique	14
2.2.1	Design example	16
2.3	Folded canonical topologies design	19
2.4	Design technique for filters with NRRs	24
2.5	Waveguide filters made using non-resonant resonators	29
2.5.1	Filter with non-resonating resonators operating at two different frequencies	29
2.5.2	Symmetric non-resonating resonators filter	30
2.5.3	Asymmetric non-resonating resonators filter	32
2.6	Conclusions	35
3	SIW tunable filters	37
3.1	Tunable lowpass filter	38
3.1.1	Fundamental mode cutoff frequency variation	38
3.1.2	Lowpass filters basic concepts	39
3.1.3	Tunable lowpass filter design	40
3.2	Tunable evanescent SIW filter	42
3.2.1	Study of the evanescent resonant mode	42
3.2.2	Filter design	42
3.2.3	Tunable coupling mechanisms	44
3.2.4	External Q tunable mechanisms	49
3.3	Waveguide filter using SIW resonator mode	51
3.3.1	Designed filters	51
3.3.2	Couplings reconfiguration	53
3.3.3	Rectangular waveguide evanescent filter with bandwidth reconfiguration	60
3.4	Tunable SIW evanescent filter prototype	61
3.4.1	Design parameters of a substrate integrated waveguide for Ku band	61
3.4.2	Ku band substrate integrated waveguide filter	63

3.4.3	Substrate integrated waveguide with parasitic elements for manufacturing a prototype	65
3.5	Conclusions	68
4	Evanescent waveguide tunable filters	69
4.1	Evanescent filter with horizontal elliptical posts	69
4.1.1	Study of elliptical posts as resonators	70
4.1.2	Filter design	71
4.1.3	Filter shielding to avoid RF-Leakage	72
4.2	Practical prototype filter	74
4.2.1	Prototype resonator analysis	74
4.2.2	Filter prototype	77
4.3	Conclusions	80
5	Tunable filters based on AMCs	81
5.1	Design and simulation of an basic tunable AMC	82
5.2	Proof of concept test	84
5.3	Inline waveguide filter with one AMC acting as side wall	87
5.4	Capacitance tapering in filters with AMCs	89
5.5	Coupling variation mechanism	93
5.6	Relation of spurious resonances and AMC parameters	95
5.7	Filter prototypes with AMC	99
5.7.1	S-band filter prototype	99
5.7.2	Ku band filter prototype	102
5.8	Conclusions	103
6	Mechanically tunable waveguide filters	105
6.1	Tunable resonators with movable arms	106
6.1.1	Rectangular cavity with metal posts	106
6.1.2	Resonators combining movable arms and dielectric materials	108
6.1.3	Rectangular cavities including elements that present angular variations	110
6.2	EVA waveguide tunable filter based on ridge resonators	111
6.2.1	Coupling modification	113
6.2.2	Tunable filter with reconfigurable bandwidth	115
6.3	Tunable filters based on movable dielectric cylinders	117
6.3.1	Tunable filters based on movable lateral dielectric circular cylinders	117
6.3.2	Tunable filters based on movable vertical dielectric circular cylinders	123
6.4	Conclusions	127
7	Novel lowpass filters	129
7.1	Filter with dielectric coated metallic posts	130
7.2	Filter with metallic posts covered with dielectric material	131
7.3	Filter with inductive dielectric posts as inverters	132
7.4	Filter with magnetic materials	133
7.5	Conclusions	135
8	Conclusions	137
	Bibliography	150

List of Figures

2.1	Equivalent circuit represented by the coupling matrix. The circuit represents a normalized lowpass filter, which is composed of parallel capacitors and constant susceptances.	10
2.2	Equivalent circuit, composed of two resonators coupled by three admittance inverters, to compute the value of M_{mn}	11
2.3	Two coupled rectangular waveguide resonant cavities. This structure is adjusted to have a similar response to the circuit depicted in Fig. 2.2. . . .	11
2.4	Structure of a resonator that has been used for evaluating the external quality factor (Q_{ext}).	11
2.5	Structure of a fifth order rectangular waveguide filter.	12
2.6	Scattering parameters magnitude for the filter presented in Fig. 2.5 and the filter described by the coupling matrix (2.4).	13
2.7	Equivalent circuit corresponding to the scaled coupling matrix. (a) Low-pass filter circuit. (b) Bandpass filter circuit.	14
2.8	(a) Open ended $\lambda_{g0}/2$ transmission line resonator. (b) Lumped resonator. Both resonators must have a similar electrical response around the center frequency.	15
2.9	Transmission parameters of the first step.	17
2.10	Transmission parameters of the second step.	17
2.11	Transmission parameters of the third step.	18
2.12	Transmission parameters of the fourth step.	18
2.13	Scattering parameter magnitudes of the filter depicted in Fig. 2.5.	19
2.14	Equivalent circuit of the filter described from the scaled coupling matrix (2.12).	21
2.15	a) Top view of the filter and its geometrical parameters b) Side view of the rectangular waveguide filter. The thickness of all irises is set to $l_i = 1$ mm.	21
2.16	Segments of the circuit shown in Fig. 2.14.	22
2.17	Transmission parameters of the first segment.	23
2.18	Transmission parameters of the second segment	23
2.19	Transmission parameters of the third segment	23
2.20	Transmission parameters of the fourth segment	23
2.21	Scattering parameters magnitudes of three filters: circuit filter, initial filter and optimized filter.	24
2.22	Two equivalent structures. The first one is a resonator with two couplings on both sides. The second one obtains the same coupling level of the first one by replacing the impedance inverters with two impedance inverters and a resonator acting as a coupling.	25

2.23	Structure of a rectangular waveguide resonator and its geometrical parameters. This structure is employed to compute the curves obtained in Fig. 2.24.	25
2.24	Transmission parameters of five different configurations of the structure presented in Fig. 2.23.	26
2.25	Transmission parameter obtained for each width w after adjusting the length of the resonator (see Fig. 2.23) to keep constant the resonant frequency.	26
2.26	Lengths of the coupling resonators adjusted for each width (w) of the structure represented in Fig. 2.23 to keep the same resonant frequency of the NRR.	26
2.27	Equivalent circuit of the structure represented in Fig. 2.23. The NRR is modeled as a common inverter, which can be inductive or capacitive. The transmission lines represent half of the main resonators.	27
2.28	Scattering parameter magnitudes of the filter made of non-resonating resonators as couplings.	28
2.29	Fourth order filter geometry with non-resonating resonators as couplings.	29
2.30	Scattering parameter magnitudes for a fourth order filter made with non-resonating resonators. The filter passband can be seen centered at 16 GHz, while the NRR are located at 13.5 GHz, 14 GHz and 17 GHz, approximately.	30
2.31	Coupling diagram for a symmetric filter with different frequencies for the NRRs.	31
2.32	Eighth order filter geometry with non-resonating resonators as couplings.	31
2.33	Scattering parameter magnitudes of the filter with non-resonating resonators with three different frequencies interspersed between them.	32
2.34	Coupling diagram for an asymmetric filter.	33
2.35	Geometry of the filter with an asymmetrical design.	33
2.36	Scattering parameters magnitude of the asymmetrical filter with non-resonating resonators (see Fig. 2.35). Nine different frequencies was employed to realize the NRR of this filter.	34
3.1	Substrate integrated waveguide structure, where one of its via-holes rows is not connected to the top metallic cover.	38
3.2	Scattering parameters magnitude, obtained when a side row of via-hole posts are disconnected from the top wall and when all posts are connected as usual in the structure shown in Fig. 3.1.	39
3.3	Lowpass filter equivalent circuit. The K values are those corresponding to impedance inverters, whereas Z_0 and θ_c are the impedances and the electrical lengths of the transmission lines, respectively.	39
3.4	Fragments of a lowpass filter. The fragment of the figure (a) is employed to compute the transmission parameter, while the fragment of figure (b) is optimized to have the same response of the previous one.	40
3.5	Structure of a lowpass SIW filter with floating post disconnected to the top wall between capacitive elliptical aperture impedance inverters.	41
3.6	Scattering parameter magnitude of the filter depicted in Fig. 3.5 with lateral posts connected or disconnected to the top wall.	41
3.7	Structure of a resonator in SIW section with a capacitor.	42
3.8	Resonant frequency as a function of the capacitance value of the structure depicted in Fig. 3.7.	43

3.9	Dielectric filled rectangular waveguide evanescent filter. The filter is composed of three resonators of the same type as those introduced in Fig. 3.7.	43
3.10	Scattering parameters magnitude for three different configurations of the filter. These states are obtained by changing the capacitances. The relevant characteristic of these configurations are collected in Table 3.1.	44
3.11	Scattering parameters magnitude for the three states of the filter and their out of band responses. The spurious passband above 3.4 GHz is due to TE_{101} resonating modes that are formed in the SIW sections between each post.	45
3.12	Two resonators coupled in an evanescent SIW structure with floating side post connected through a varactor diode to the top wall. This last varactor diode will control the coupling level between the resonators.	45
3.13	Scattering parameter magnitude of Fig. 3.12 evaluated for capacitances of 5 pF, 7 pF and 9 pF for the varactor that controls the coupling level. . . .	46
3.14	Coupling level (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance for the structure in Fig. 3.12. The coupling level is increased if the spurious free range is decreased, due to the NRR behavior.	47
3.15	Two resonators coupled in an evanescent SIW structure with two open floating side posts connected through varactor diodes to the top wall. . . .	47
3.16	Coupling level (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance (the two varactors take the same capacitance value). The coupling level variation is higher than that achieved in Fig. 3.14, while keeping a similar SFR level.	48
3.17	Two resonators, coupled in an evanescent SIW structure, with three open ended floating posts on each side, connected through varactor diodes to the top metallic wall. The varactor diodes on each side will control the coupling level between the resonators.	48
3.18	Scattering parameter magnitude obtained for the structure depicted in Fig. 3.17.	49
3.19	Structure with one resonator and two open ended floating posts connected through a varactor diode to the top metallic wall.	50
3.20	External quality factor (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance (the four varactor shown in Fig. 3.19 take the same capacitance value).	50
3.21	Structure of a reduced height rectangular waveguide evanescent filter. . . .	51
3.22	Scattering parameter magnitudes corresponding to different configurations of the evanescent filter with reduced height depicted in Fig. 3.21. Each configuration has been obtained by changing the capacitance associated with the resonators.	52
3.23	Structure of the evanescent filter with reduced height shielded by using dielectric boxes on the top cover.	53
3.24	Scattering parameter magnitudes of the evanescent filter with reduced height shielded by using dielectric boxes on the top cover. The structure of the filter is represented in Fig. 3.23.	54
3.25	Structure composed of two resonators and two posts for coupling tuning. . .	54
3.26	Transmission parameter magnitudes as a function of capacitor value for a very small input/output couplings. The structure simulated is represented in Fig. 3.25	55

3.27	2D structure of two coupled resonators and six posts connected through MEMS to the top metallic wall.	55
3.28	Scattering parameter magnitude as a function of the number of connected MEMS (Symmetrically). The structure simulated is represented in Fig. 3.27.	56
3.29	2D structure of a resonator and twelve posts connected through MEMS to the top metallic wall.	57
3.30	Scattering parameter magnitudes as a function of the number of MEMS connected (Symmetrically). The structure simulated is presented in Fig. 3.29.	57
3.31	Structure employed to compute the coupling between two resonators by using a scattering parameters simulation.	58
3.32	Coupling levels as a function of the position of the posts (pos), depending on the MEMS state, for the structure shown in Fig. 3.31.	58
3.33	Structure employed to compute the Q_{ext} as a function of the location of the tuning posts.	59
3.34	External quality factor as a function of the position of the posts (pos), depending on the MEMS state, for the structure shown in Fig. 3.31.	59
3.35	Third order evanescent rectangular waveguide filter with two possible states.	60
3.36	Scattering parameter magnitude for the configuration with center frequency at 14 GHz. This is the lower center frequency configuration of the filter depicted in Fig. 3.35.	61
3.37	Scattering parameter magnitude for the configuration with center frequency at 17 GHz. This is the higher center frequency configuration of the filter depicted in Fig. 3.35	61
3.38	Sketch of the substrate integrated waveguide designed for a filter at Ku band, and the microstrip input port and its transition to it. The employed substrate is ROGER 5880 with $\epsilon_r = 2.2$ and the substrate thickness is 0.127 mm.	62
3.39	SIW resonator with four rectangular surfaces (two above the top wall and two below the bottom wall) acting as capacitances.	62
3.40	Ku-band SIW resonator as a function of the dielectric substrate thickness and the varactor diode capacitance.	63
3.41	Substrate integrated waveguide filter at Ku band.	64
3.42	Scattering parameter magnitudes of the SIW filter depicted in Fig. 3.41, for three different states. These states have been obtained by changing the capacitor value as shown in the legend.	64
3.43	Sketch of the resonator with non-filled via holes. The dimension P_c is 1.2 mm, P_w is 0.5 mm and $R = 0.9$ mm. The size of the varactor diode is 0.9 mm $\times$ 0.5 mm.	65
3.44	Ku-band SIW resonator as a function of the dielectric substrate thickness and the varactor diode capacitance.	66
3.45	Layout and dimensions of the designed filter, where $l_1 = 2$ mm and $l_2 = 5.6$ mm. The board dimensions are $b_l = 31.2$ mm $\times$ $b_w = 27$ mm. The board also includes the pads to place the bias network required the polarization of the varactors. The SIW parameters are depicted in Fig. 3.38 and are collected in Table 3.4.	67
3.46	Scattering parameter magnitudes of the filter presented in Fig. 3.45. These states are obtained by varying the capacitances from 0.31 pF to 0.41 pF.	67

4.1	Evanescent rectangular waveguide section with a side post attached to the waveguide as resonator. The elliptical cylinder is not touching the top and bottom walls, existing an air gap in those areas.	70
4.2	Structure of the proposed evanescent rectangular waveguide tunable resonator. The red surfaces on the top and bottom walls are distributed capacitances, being the plates, in the middle, isolated from the rest of the structure.	71
4.3	Scattering parameters and RF-leakage for different configurations computed by sweeping the distributed capacitance of the tunable resonator depicted in Fig. 4.2	71
4.4	Structure of a third order filter realized with floating plates (marked surfaces) connected through capacitors.	72
4.5	Scattering parameter magnitudes of the filter, illustrated in Fig. 4.4, realized with floating plates connected through capacitors. The center frequency is shifted by changing the capacitances for each resonator.	73
4.6	Filter with dielectric shielding boxes. The dimensions of this filter are the same that the previous one presented in Fig. 4.4. The only difference is the dielectric boxes located above and under the waveguide where tuning mechanism are placed.	73
4.7	Scattering parameter magnitudes of the filter with dielectric shielding boxes, depicted in Fig. 4.6 for four different tuning states.	74
4.8	Evanescent resonator with discrete capacitances.	75
4.9	Transmission parameters of the evanescent resonator with discrete capacitances depicted in Fig. 4.8 for six different widths of the capacitors.	75
4.10	Resonator with discrete capacitances an SIW structure acting as top and bottom walls.	76
4.11	Layout of the SIW structure located above/under the resonator. The biasing network will be included in the SIW structure when needed.	76
4.12	Resonant frequency as a function of the depth of the elliptical post (h). The figure shows two curves of resonant frequency (left axis) that correspond to the limits that were imposed to the varactor.	77
4.13	Resonant frequency of the resonator depicted in Fig 4.10 as a function of the semiaxes w_y and w_z	78
4.14	Structure of the electronically tunable third order evanescent rectangular waveguide filter.	78
4.15	Scattering parameter magnitudes of four configurations of the filter. These configurations are computed by only changing the capacitance associated with each varactor diode.	79
4.16	Two extreme configurations of the filter with $C = 0.4$ pF and $C = 1.1$ pF alongside with their out of band responses.	80
5.1	a) Layout of the proposed unit cell for a periodic surface. b) Front view of the waveguide and the periodic surface acting as side wall.	82
5.2	Simulation schemes for an unit cell in ANSYS HFSS.	83
5.3	ANSYS HFSS model for simulation of periodic structures. The periodic structure is composed of two strips connected to a capacitor printed on a dielectric substrate.	83

5.4	Reflection coefficient phase of a unit cell of the periodic structure for three states 0.1 pF, 1 pF and 50 pF. These results corresponds to the unit cell presented in Fig. 5.1 and using the setup displayed in Fig. 5.3.	84
5.5	Evanescent rectangular waveguide structure with an artificial magnetic conductor acting as a side wall.	85
5.6	Scattering parameter magnitudes of the evanescent rectangular waveguide with an artificial magnetic conductor in one of its sides represented in Fig. 5.5.	86
5.7	Electric field in a evanescent rectangular waveguide with an AMC acting as lateral wall represented in Fig. 5.5. These fields are evaluated at 14.5 GHz with the configuration determined by $C_{AMC} = 0.1$ pF. The fields corresponding to the fundamental mode at this particular frequency are propagative and at this frequency the artificial surfaces behaves close to PEC.	86
5.8	Electric field in a evanescent rectangular waveguide with an AMC acting as lateral wall represented in Fig. 5.5. These fields are evaluated at 14.5 GHz with the configuration determined by $C_{AMC} = 1$ pF. The fundamental mode fields corresponding to this configuration are propagative at this particular frequency, being the maximum amplitude located on the side wall, where the AMC is placed.	86
5.9	Structure of a rectangular waveguide filter with an AMC acting as side wall.	88
5.10	2D structure of the filter of Fig. 5.9.	88
5.11	Scattering parameter magnitudes for the two extreme configurations of the filter illustrated in Fig. 5.9.	89
5.12	Capacitance tapering functions in terms of normalized distance.	90
5.13	Capacitor values as a function of their position in the filter depicted in Fig. 5.10 using a half-sine tapering function.	91
5.14	Scattering parameters magnitudes of the filter depicted in Fig. 5.10, for two configurations with a half-sine tapering function applied to the tunable AMC.	91
5.15	Scattering parameter magnitudes of the filter depicted in Fig. 5.9. Four tuning states are plotted: the baseline and three with half-sine tapering, with the maximum T_j (5.2) amplitudes: 0.2 pF, 0.5 pF and 1 pF.	92
5.16	Scattering parameter magnitude of the configuration with maximum amplitude of the tapering 1 pF(half-sine tapering function) for the filter depicted in Fig. 5.9.	93
5.17	Fourth order filter with two artificial magnetic conductors acting as side walls. The posts are connected to the top wall through varactor diodes, instead of being short-circuited.	94
5.18	Scattering parameters of the filter, depicted in Fig. 5.17, with a half-sine tapering configuration with T_j set to 0.2 pF.	94
5.19	Artificial magnetic conductor structure with a metallic strip. The geometry of the strip is defined by its width (W) and its length (L). The dielectric has relative permittivity $\epsilon_r = 2.2$ and its thickness will be varied in different designs.	95
5.20	Reflection phase coefficient for the five configurations listed in Table 5.2.	96
5.21	Third order filter structure with non-tunable AMCs acting as side walls.	97
5.22	Test of filters with different configurations of AMCs.	98

5.23	Layout of the proposed unit cell for a practical periodic surface with $f_{PMC} = 3.7$ GHz, realized in a dielectric substrate.	99
5.24	Reflection coefficient phase response of the AMC designed unit cell. The PMC boundary condition is set at 3.7 GHz by using a 1.2 pF capacitance. The phase variation slope of the PMC is $-545^\circ/\text{GHz}$, measured near that frequency point.	100
5.25	Structure of the filter in S-band. The side walls have been replaced by the previously calculated AMC (see Fig. 5.23).	101
5.26	Scattering parameters of the filter presented in Fig. 5.25. The baseline configuration is designed with all the capacitors set to 0.3 pF.	101
5.27	Reflection coefficient phase response of the designed AMC unit cell.	102
5.28	Structure of the filter in Ku-band with two AMC acting as side-walls.	103
5.29	Scattering parameter magnitudes of the filter presented in Fig. 5.28.	103
6.1	Rectangular waveguide cavity with a moving arm placed in the middle.	106
6.2	Rectangular waveguide cavity with multiple post alongside a moving arm placed inside.	107
6.3	Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.2.	107
6.4	Scattering parameter magnitudes of the resonator represented in Fig. 6.2, when the ports are placed in the front and side walls.	108
6.5	Resonant cavity with two dielectric bricks placed touching the center of the top and bottom wall, together with a movable arm centered in the structure.	109
6.6	Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.5.	109
6.7	Resonant cavity with an elliptical cylinder in the middle. The elliptic post rotates from 0° to 90°	110
6.8	Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.7.	110
6.9	Evanescent rectangular waveguide filter. a) Top view of the evanescent rectangular waveguide filter. The relevant dimensions are shown. b) Side view of the filter.	111
6.10	Scattering parameters for the filter presented in Fig 6.9. The filter center frequency tuning range goes from $f_c = 12.5$ GHz to 13 GHz. The lowest frequency configuration has a reduced bandwidth of 80 MHz.	112
6.11	Structure of two ridge resonators coupled in an evanescent rectangular waveguide. Between these posts it is placed a third circular cylinder post which constitutes a third resonator, employed as NRR for controlling the coupling between the other.	113
6.12	Coupling level between the two main resonators as a function of the penetration of the post for two different values of the radius of the horizontal post.	114
6.13	Scattering parameter magnitudes of the structure depicted in Fig. 6.11 for different depths of the horizontal post, namely, $t = 7$ mm, 8 mm and 9 mm.	114
6.14	3D view of the evanescent rectangular waveguide filter with non-resonating resonators.	115
6.15	Tunable evanescent rectangular waveguide filter with horizontal post resonator as NRRs.	116

6.16	Scattering parameters of the structure depicted in Fig. 6.14. The bandwidth is maintained for the three tuning states of the filter. Also, the resonances produced by the non-resonating resonators can be seen in the graph.	116
6.17	3D view of the resonator with a dielectric circular ring shaped cylinder. The frequency variation is achieved by modifying the dimension Offset_x . . .	118
6.18	Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.17.	118
6.19	Electric field distribution of the third resonant mode for the structure of Fig. 6.17 at 19 GHz.	119
6.20	Test coupling structures consisting in two resonators coupled by an iris (centered or attached to a side wall). The dimensions for the two dielectric ring shaped resonators and its cavities are the same as the one presented in Fig. 6.17.	119
6.21	Coupling level as a function of the depth of the posts in the cavity (Offset_x) represented in Figs. 6.20(a) and 6.20(b). The computed values are cross and circular markers, while the interpolated values are lines. The center frequencies that have been use to evaluate the couplings, are indicated for each marker.	120
6.22	Tunable filter built using lateral dielectric ring shaped cylinders.	121
6.23	Scattering parameter magnitudes of the filter depicted in Fig. 6.22. The baseline filter is designed for $\text{offset}_x = 1$ mm.	122
6.24	Scattering parameter magnitudes of the filter designed with lateral dielectric ring cylinders.	122
6.25	3D view of the proposed resonator with a dielectric circular cylinder that can vary its depth into the rectangular cavity.	123
6.26	Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.25.	124
6.27	Electromagnetic fields of the third resonant mode of the cavity depicted in Fig. 6.25.	124
6.28	3D view of a sixth order filter with dielectric circular cylinders that can vary its depth into the cavity.	125
6.29	Scattering parameter magnitudes of the filter depicted in Fig. 6.28 for four different tuning states. Each configuration is obtained by only changing the offset_y parameter, , according to Table 6.4.	126
6.30	Scattering parameters magnitudes of the filter depicted in Fig. 6.28 for the two extreme configurations ($\text{offset}_y = 1$ mm and $\text{offset}_y = 3$ mm). This figure also shows the spurious band that appears below the frequencies of the filter main passbands.	126
7.1	Structure of a sixth order lowpass filter implemented by using posts coated by dielectric material acting as impedance inverters.	130
7.2	Scattering parameters of the filter presented in Fig. 7.1. The lowpass filter is implemented by using metallic posts coated by a dielectric material with constant thickness, as shown in Fig. 7.1.	131
7.3	Structure of a sixth order lowpass filter implemented by using posts covered by dielectric material as impedance inverters.	131

7.4	Scattering parameter magnitude of the lowpass filter presented in Fig. 7.3. The filter is composed of rectangular posts covered with dielectric material from the bottom wall to the top wall of the rectangular waveguide.	132
7.5	Structure of a sixth order lowpass filter implemented by using dielectric material post as impedance inverters.	132
7.6	Scattering parameter magnitudes of the filter presented in Fig. 7.5. The scattering parameters show spurious resonances produced by the dielectric posts between 7.5 GHz and 9.5 GHz.	133
7.7	Structure of a sixth order lowpass filter implemented by using magnetic material slabs as part of the impedance inverter.	134
7.8	Scattering parameter magnitudes of the filter presented in Fig. 7.7. The filter is composed of impedance inverters realized with capacitive irises and magnetic materials near them.	134
8.1	Coupling-routing diagram, of a parallel-connected two-port network. This topology represents a fourth order filter.	139
8.2	Second order filter with an AMC acting as side-wall. The couplings are implemented with floating posts connected though MEMS to the top and bottom walls.	140

List of Tables

2.1	Initial dimensions for the filter. These dimensions are obtained after optimizing the structures presented in Figs. 2.2 and 2.4 to achieve required inter-resonator couplings and Q_{ext}	13
2.2	Dimensions obtained in the different steps of the design procedure, following the scheme of Fig. 2.5. Dimensions are given in mm.	19
2.3	Dimensions of the filter. This table displays the parameters optimized in all steps and the values of the dimensions, after a global optimization of the whole structure. The last column shows the relative error between the final dimensions (obtained after a global optimization) and the dimensions of step 4.	23
2.4	Transmission parameters required by the scaled coupling matrix (2.13), iris widths and resonator lengths to associated to each NRR.	27
2.5	Resonator lengths obtained after optimizing the phase of the transmission parameter.	28
2.6	Dimensions of the filter depicted in Fig. 2.29. The dimensions are given in mm. The structure of the filter is symmetrical from input to output. . . .	30
2.7	Dimensions of the filter depicted in Fig. 2.32. The dimensions are given in mm. The structure of the filter is symmetrical from input to output. . . .	32
2.8	Dimensions of the asymmetric eighth order filter depicted in Fig. 2.35. The dimensions are given in mm. The numbering is made from one port to another; however, these dimensions are not marked in the sketch due to the large size of the structure.	34
3.1	Center frequency and absolute bandwidth for each configuration of the filter. In addition, the table also includes the percentages of tunability range and bandwidth variation, taking the second configuration as baseline.	44
3.2	Capacitances for four different tuning stated for the filter with reduced height, represented in Fig. 3.21. The table also includes the center frequency, bandwidth, constant return loss level and the unloaded quality factor obtained for each tuning state.	52
3.3	Capacitances for four different tuning stated for the filter with reduced height, depicted in Fig. 3.23. The table also contains the center frequency, bandwidth, constant return loss level and the unloaded quality factor obtained for each tuning state.	53
3.4	Parameters of a SIW structure and its transition from a microstrip input line. These parameters are graphically represented in Fig. 3.38.	62
3.5	Filter center frequency, bandwidth and capacitance values for each state of the filter depicted in Fig. 3.41. The four column corresponds to the return losses level used as reference for measuring the bandwidth.	65

5.1	T_j values for the tapering distribution. Each variable T_j controls the tapering in a resonator or the input/output port.	92
5.2	Dimensions and phase variation rate of the AMCs, depicted in Fig. 5.19, designed to study their effects in a rectangular waveguide filter.	95
6.1	Specifications of the states of the filter shown in Fig. 6.9 and their corresponding ridge post heights.	112
6.2	Reconfigured dimensions and center frequency for each tuning state of the filter.	115
6.3	Tuning states of the tunable filter, depicted in Fig. 6.22, by changing the depth of the lateral dielectric cylinders.	121
6.4	Tuning states of the tunable filter, depicted in Fig. 6.28, by changing the depth of the dielectric circular cylinders.	125

Chapter 1

Introduction

1.1 State of the art

Nowadays, communications systems need to be dynamically changed and be flexible [Matthaei and Yones, 1980]. The flexibility will provide the ability to respond faster to market demands. Microwave filters are key components in communication systems, whose function is to remove undesired RF signals and transmit the desired one within a frequency range. In order to provide some flexibility, these systems have filter banks to provide many channels with different bandwidths. However, these channels are fixed at design time, and they cannot be modified to other configurations in their lifetime. Reconfigurable filters can be used to replace these filter banks, offering a solution to adapt their specifications to the requirements of the communication system at certain moments. Although their design time and cost can be higher than a single non-tunable filter, the footprint and mass are considerably reduced as compared to filter banks [Mansour et al., 2014].

Within the available microwave technologies, microstrip is the most common one for designing tunable filters. It is an obvious election, due to its cost and small size, together with the fact that it allows a direct integration of electrical components, such as resistors, inductors, capacitors and diodes. For instance, in [Cho and Rebeiz, 2015], a fourth-order filter is designed by using varactor diodes as tuning mechanism. The particularity of this design is that the structure of the filter has eight resonators, which are divided into two branches. Each branch is a four order filter with a different center frequency. This, in turn, has a tuning varactor diode for each resonant strip, resulting in a tunable fourth order filter. The advantage of this implementation is to achieve a considerable tuning range by placing two different filters in a common structure and being able to disconnect one of the branches of the circuit. On the other hand, other materials can be used as tuning mechanism like ferrites. One option is to make a thin film with a nickel ferrite like in [Yang et al., 2013], which presents a filter based on this concept in Ku-band. However, ferrite materials are tuned by applying a magnetic field, which usually are placed in a complex biasing structure difficult to integrate with microwave components. In addition, these materials lead to high ohmic losses. Nevertheless, the filter presents almost 1 GHz of tuning range, which is an outstanding result at Ku-band. A similar approach is performed in [Peng et al., 2015], where the substrate is realized by using a metamaterial that changes the center frequency of the filter as a function of applied DC current. Other authors propose similar structures by using liquid crystal as, for instance, in [Shen et al., 2015].

Liquid crystal has the property of changing its dielectric permittivity as a function of the applied electric field. Thus, using liquid crystal as part of the substrate employed for a microstrip filter, its center frequency can be shifted.

Other planar technologies, such as substrate integrated waveguides (SIW) or coplanar waveguides are also employed for realizing tunable filters. These planar technologies share the benefits of cost, small form factor, and direct integration of electronic components. Moreover, these technologies present better quality factors than microstrip. There exist many implementations built with varactor diodes [You et al., 2016, Anand and Liu, 2015, Yang et al., 2014, Anand et al., 2013, Sekar and Entesari, 2012] achieving wide tuning ranges. A similar solution is to employ MEMS to achieve electrical tuning as shown in [Sekar et al., 2011], where a cavity with three MEMS switches is designed. This paper presents a two-pole filter that has 14 different states. Another popular option is to employ piezoelectric actuators to tune cavities where a resonating mode with a strong capacitive behavior is excited. For instance, in [Lee et al., 2015, Saeedi et al., 2014, Liu et al., 2012], the resonant cavities are built with one or more posts (can be dielectric or metallic) where the dielectric material inside the substrate has been removed. Thus, the capacitance formed between the posts and the piezoelectric actuator is varied in order to obtain a tunable resonator. An alternative to piezoelectric actuators is to employ screws that are introduced into a circular cavity, performing a mechanical change in the structure, as was presented in [Huang et al., 2016]. There exist other less common tuning methods like making a tunable filter by introducing liquids, as presented in [Barrera and Huff, 2013, Dang et al., 2015]. However, these solutions are not widely used, due to the complex setup for the tunable system. For SIW structures, the liquid crystal option is also used in the same way as microstrip. For instance, in [Jiang et al., 2019], it is implemented a SIW structure by employing three layers of dielectric substrates: one for the bottom wall, the intermediate layer, where the inner part has the dielectric material removed and filled with liquid crystal, and acting as the top wall with complementary split-ring resonators. Then, the center frequency of the filter is modified by applying an external electric field, changing the liquid crystal electrical properties. Finally, it is interesting to mention the SIW design presented in [Adhikari et al., 2013], which employs ferrite materials and varactor diodes mixed for tuning, center frequency and bandwidth of the filter at the same time. The Ferrite slabs replace the via holes acting as lateral walls of the resonant cavities. These slabs are tuned by a magnetic field, while the resonant frequency of the cavities is controlled by a varactor diode. This design is very interesting, since it operates at Ku-band frequencies, using SIW and leading to reasonable insertion losses.

If higher quality factors are required, waveguide technology is the most appropriate. However, it is challenging to integrate electronic components, such as varactor diodes or MEMS in filters built in waveguide technology. There exist some works like [Pelliccia et al., 2015], where MEMS are introduced in dielectric substrates placed inside rectangular waveguide filters, connecting or disconnecting metallic strips. This implementation allows to achieve high-quality factors, but as drawback, the filter only has two possible states. Other works use MEMS as switches for a capacitor bank, as presented in [Mansour, 2015]. There are also publications with varactor diodes. For instance, it is described in [Aouidad et al., 2015] how a dielectric substrate is introduced as part of a stepped impedance resonators. On the dielectric substrate external wall, varactor diodes are placed. Their feeding network can be used to tune the resonant

frequency of each resonator independently. Again, another option is to employ materials like ferrites or liquid crystal that change their properties as a function of an imposed static magnetic or electric field respectively, as it presented in [Acar et al., 2015, Polat et al., 2019, Franke et al., 2014]. However, their use is still not extended in the industry. The most common tuning mechanism in waveguide structures is to apply a mechanical deformation to the structure by inserting adjustable screws or employing movable walls. In [Arnold et al., 2014a, Arnod et al., 2015], two filters are presented by using a concept of using resonators as coupling structures. These resonators, used as coupling structures, are called non-resonating resonators. The use of non-resonating resonators is an a useful alternative, since tuning resonators is always easier than tuning couplings without generating other adverse effects. They achieve an excellent tuning range at Ka-band with moderate insertion losses. The tunability is obtained by employing movable plungers in the main resonators and in the non-resonating resonators. Other approach is to employ piezoelectric actuators to tune combline resonators as presented in [Yang et al., 2018]. Also, combline resonators tuned by screws are employed in [Yuceer, 2016]. This last tunability example is interesting, due to the way that two filters are working together as if they were one. We find this concept promising because it can be used to tune each filter independently and vary the bandwidth. In [Rosenberg et al., 2016, Yassini et al., 2012], two waveguide filters at Ka-band composed of circular cavities and movable walls are presented. These designs provide low insertion losses, since the cylindrical resonating modes usually have high quality factor values. Another option with cylindrical cavities is to use a dielectric slab rotating inside the cavity to tune the center frequency as it was presented in [Nam et al., 2018]. By using a low loss dielectric material, the overall quality factor is not affected in this case. Finally, we want to mention the filter presented in [Basavarajappa and Mansour, 2018], where a standard rectangular waveguide filter, implemented with the fundamental resonating mode TE_{101} with a movable wall, is designed. The particularity of this work is that their authors present an algorithm to make a coupling structure, in which the coupling is not altered when the filter is shifted up or down in frequency. Thus, they achieve an almost constant absolute bandwidth.

Space communications systems require lowpass filters to remove adjacent channels or undesired spurious passbands. These undesired spurious bands can be produced by the same communication system, for instance, intermodulation products generated by high-power amplifiers [Aitchison et al., 2001]. Therefore, lowpass filter requirements are: high selectivity, wide spurious free-range responses [Levy, 1973], and high-power handling capabilities [Arregui et al., 2010]. However, under vacuum conditions, some adverse effects such as Corona or Multipactor can appear [Vaughan, 1988], which may result in a significant damage in the device. A typical implementations of lowpass filters consist in employing capacitive irises as impedance inverters, and small gaps are needed when the designed reflection level is high. Consequently, the high-power handling capabilities are critical [Gonzalez-Iglesias et al., 2012]. A possibility to increase the multipactor breakdown threshold is to employ dielectric materials as reported in [Torregrosa-Penalva et al., 2010].

The design of microwave filters is usually a time-consuming task, especially for those with complex topologies. Full-wave simulation software requires a high computational cost to analyze the whole filter structure. Thus, employing efficient design techniques is one of the best options to find an initial approximation of the microwave filter. There exist many

techniques that can be used to obtain initial dimensions for a filter [Matthaei and Yones, 1980, Hong and Lancaster, 2001]. However, they usually require a global optimization of the structure in order to match the requirements. Although these techniques may be suitable for low order filters or simple topologies, for complex topologies it could be a difficult task to optimize the whole structure if the initial response is far from the desired one.

One of the most employed design techniques [Cameron et al., 2007] computes the resonators and couplings separately, joining all the individual pieces at the end. On one hand, this process is usually performed using eigensolvers (provided by the full-wave software), which do not require a long simulation time. On the other hand, after assembling the structure of the filter, the response is usually far from the target. Thus, global optimization is mandatory from the initial dimensions. However, these optimizations are commonly performed by using gradient descent algorithms, which can get trapped into a local minimum if the target response is not close enough to the initial response of the filter.

In order to solve the problem of local minima, there exist the possibility to employ genetic algorithms for microwave filter designs [Goldberg, 1989, Lai and Jeng, 2006, Goudos and Sahalos, 2010, Wang et al., 2005]. These algorithms prevent getting trapped in minima as it happens in gradient algorithms. Nevertheless, it requires a high number of iterations until the filter is refined, leading to a high computational cost. Another possible option is to use space mapping techniques, which uses the correspondence of a coarse model to a fine model for a faster design, in order to save computational designing time as reported in [Hinojosa et al., 2015, Hinojosa et al., 2011, Sans et al., 2014, Bandler et al., 1994].

Some design techniques are based on equivalent circuits. For instance, in [Cogollos et al., 2012], the authors employ an equivalent circuit for designing a dual-mode circular waveguide filter (quartet topology with one cross-coupling). This technique independently computes the iris and resonators along the main path, while the cross-couplings are finally adjusted by using the equivalent circuit. This technique results in initial dimensions really close to the final ones. However, it is not clear how this design technique can be extended for larger topologies or larger or different topologies, with increasing higher number of cross-couplings.

In this context, a step by step procedure can be used to find the dimensions of the filter with a lower computational effort [Guglielmi and Melcon, 1993, Guglielmi, 1994]. The technique consists of adding pieces of the structure sequentially and optimizing the newly added dimensions in each step. Thus, the parts added in each step will be taken into account during the optimization of the rest of the structure. At the same time, they are considered in further steps. This process leads to a filter response, which is always close to the target one. This technique was also expanded in [Martínez-Martínez et al., 2018] to dual-mode using all-inductive waveguide technology.

1.2 Motivation

At this moment, satellite communication payloads are evolving to be more flexible. The current passive microwave devices that compose the actual systems must be designed specifically to a given application. However, there is a demand for new communication systems that allow adapting their operation for the services that the market demands. Actually, this situation requires large filter banks with passive filters working at fixed

frequencies. However, the use of reconfigurable filters may be a better choice in terms of size, cost and required low power. Moreover, these new reconfigurable systems are not fixed to a particular frequency band. This is a significant advantage in order to adapt them to the environmental conditions, new regulations of the electromagnetic spectrum or new customer demands.

The most common technologies for microwave systems are microstrip, substrate integrated waveguide (SIW) and classical bulky waveguides. The election of technology usually depends on the trade-off between cost, volume and quality factor (related to ohmic losses). For reconfigurable devices, planar technologies (microstrip and SIW) are more appropriate, due to the easy integration of the components required for the tunability, such as varactor diodes, DC block capacitors and RF choke inductors. On the other hand, traditional waveguide technology is the most employed option for high-frequency bands due to its lower losses and high power handling capabilities. However, introducing tunability to a waveguide filter is a complex task, because it requires additional parts to include electronic components (varactor or MEMS and their associated feeding elements). Alternative methods include the use of actuator drives, such as stepper motors, plungers, etc.

In this PhD. Thesis, our goal is to find novel reconfigurable microwave filters to work at high-frequency bands like Ku-band or Ka-band. Our focus is the electronic reconfiguration of filters in Ku-band, where available commercial components can be used. On the other hand, mechanical actuators will be employed for filters at Ka-band, where electronic components fail to operate correctly, or for applications needing high-quality factors.

1.3 Objectives and working plan

The main objective of this PhD. dissertation is to investigate reconfigurable filters for satellite communication systems working at Ku-band and/or Ka-band. At the same time, to carry out studies related to reconfigurable filters with electronic tunability in high-frequency bands. Due to the limitations of the current commercial electronic components, the designs at Ka-band are based on performing mechanical deformations of the structure, while filters at Ku-band are realized by using electronic components for achieving reconfiguration. On the other hand, the technologies that will be used are substrate integrated waveguide (SIW) and traditional rectangular waveguides. In this work, the SIW technology is employed only for Ku-band filters, while rectangular waveguides are used for designing filters in both bands. The SIW technology is employed due to its compactness, leading to compact sizes, and low cost. However, the quality factors that this technology offers are reduced if compared to filters realized in conventional rectangular waveguide technology. In spite of these disadvantages, the easy integration of the electrical components (surface mount devices) represents an essential property for designing tunable filters. Also, it should be remarked that due to the shielding nature of SIW structures, the RF-Leakage, always present in electrically tunable filters, is minimized. By contrast, classical waveguide technology offers the highest quality factors for microwave filters. However, the structures are bulkier when compared to planar technologies. Moreover, introducing electronic components is a challenge in this waveguide technology. Thus, an important challenge will be how to deal with this problem in the resulting guided structure.

The working plan for achieving these objectives can be summarized as follows:

- **Study of the state of the art in reconfigurable filters.** A bibliography research was conducted; this provides a background of the current situation of reconfigurable filters and their relevance in new communication systems. Moreover, this research was focused in SIW and traditional waveguide technologies, which were selected due to their low cost in case of SIW and the high quality factors from classical waveguide. Even though the vast majority of scientific papers do not present filters at high frequencies, it provides interesting and relevant information about the technologies employed to add tunability in filters an integration with other microwave components, such as varactor diodes, microelectromechanical systems (MEMS) and mechanical actuators.
- **Study of filters with ideal capacitances.** In order to determine if a filter is suitable for tunability, the first step is to analyze the behavior of this filter with ideal capacitors. This allows the investigation of tunable filters without the restrictions imposed by physical components. These initial studies provide useful information about the capacitance required for a certain application and the spurious frequency band that can result in this ideal structure.
- **Realize tunable filters with varactor diodes.** Firstly, this task will be carried out by a composition of varactor diodes that modify the resonant mode in the same way as the ideal capacitors. Secondly, the filter is optimized and adjusted, taking into account the parasitic inductance introduced by varactor diodes, which is significant for high-frequency filters.
- **Develop mechanically tunable filters.** These filters have the clear advantage that their design is conducted as using electromagnetic simulators. Some test are performed to the proposed resonators to establish their tuning range in terms of a certain mechanical variation inside the structure. Through these tests, it is possible to select better options for designing tunable filters. On the other hand, these filters require an additional analysis of their coupling structures, in order to produce a high coupling level variation between the different tuning states of the filter.

1.4 Thesis outline and relevant results

This section provides a general overview of the eight chapters that compose this work.

- **Chapter 1: Introduction.** This is the present chapter, and provides a general background of the importance of reconfigurable microwave filters and their relevance in new flexible payloads. Moreover, a state of the art is presented, which shows the most relevant tunable filters reported in recent scientific papers.
- **Chapter 2: Design techniques for bandpass filters based on the coupling matrix.** This chapter is mainly focused on reviewing design techniques that are useful for developing tunable filters. The first technique is based on computing couplings from the scattering parameters, which is extremely useful for tunable filters, where using eigensolutions is not always possible or is a more complex task. The second one is a step-by-step technique that achieves an extremely good initial

dimensions for inline filters with a low computational cost. This design technique is extended for the first time to the design of rectangular waveguide folded canonical filters. It should be remarked that this kind of filters is usually employed in the industry, due to the transmission zeros (real or complex) that can be implemented in these topologies. Therefore, this design technique offers a great advantage over traditional techniques [Guglielmi and Melcon, 1993, Martínez-Martínez et al., 2018]. Also, in this chapter, the study of filters whose couplings are made using non-resonating resonators is performed, which is a promising concept for tunable filters. For designing these filters, the scaled coupling matrix is convenient since it allows us to compute the exact coupling level in terms of the transmission parameter. Then, it is easy to match that level using a non-resonating resonator. The importance of these designs is to obtain information about the problems that can come from placing non-resonating resonators as couplings in microwave filters.

- **Chapter 3: Substrate integrated waveguide tunable filters.** This chapter introduces a novel reconfigurable structure SIW filter. The resonant mode employed for this filter provides a wide tuning range if compared to other SIW filters reported in scientific databases. In order to excite this resonant mode, a via hole is placed centered in the SIW structure and disconnected from the top and bottom metallic walls. To introduce tunability, some varactor diodes are connected to the walls and the floating via hole. Moreover, some test were performed by using non-resonant resonators in this SIW structure for the coupling elements. The advantage of this non-resonant resonators is that they can be implemented on a side post. Thus, it is not necessary to increase the size of the filter to introduce bandwidth tuning to the design. In this chapter, it is also included an equivalent design in rectangular waveguide. This last rectangular waveguide filter shares the pros and cons that are present as compared to the SIW design.
- **Chapter 4: Evanescent waveguide tunable filters.** Evanescent filters are interesting to realize reconfigurable filters. Evanescent filters have their passbands in a waveguide section where the fundamental mode is propagative at a higher frequency. Thus, the operation mode is under the cutoff in that section, allowing to control the coupling between resonators by acting on their separation. Moreover, evanescent filters usually have associated a wide spurious free range, because of the strong attenuation of the fundamental in a wide frequency band. This is especially relevant for tunable filters, which are usually inappropriate in terms of spurious free range. Finally, this chapter proposes a reconfigurable rectangular waveguide filter based on electronic tuning (MEMS and varactor diodes) at Ku-band. In order to introduce the physical components in a rectangular waveguide, two additional SIW structures are included in the physical design. They act as the top and bottom walls, and at the same time, they provide additional shielding to the structure, avoiding RF-Leakage in these filters.
- **Chapter 5: Tunable artificial magnetic conductors applied to rectangular waveguide filters.** Artificial magnetic conductors are typically employed in antennas as ground planes due to their unique properties. In this chapter, we apply for the first time to add tunability in a rectangular waveguide filter. The first part of this chapter is devoted to provide a general background in the design of tunable artificial magnetic conductors. For this application, a capacitance is integrated in the structure of the artificial magnetic conductor. A proof of concept test is performed, where the artificial magnetic conductor replaces a sidewall of an evanescent

rectangular waveguide section. This test shows an intuitive interpretation of the electromagnetic field. Finally, this concept is applied to some filter examples replacing one or both side walls by tunable artificial magnetic conductors. Various filters are designed to demonstrate the effects produced due to the introduction of artificial magnetic conductors. To the author knowledge, this is the first time that tunable artificial magnetic conductors have been used as a tuning mechanism in microwave filters.

- **Chapter 6: Mechanically tunable rectangular waveguide filters.** This chapter presents a novel waveguide resonators using movable parts inside the structure. This chapter is divided into three sections. The first section analyses resonators, whose tuning method consists of a movable arm. This arm provides tunability by providing rotation around its center. Depending on the configuration this movable arm can spin around up to 360° . The resonant modes, presented in this section, have associated a good tuning range, and many of them maintain their quality factor almost constant in their configurations. In the second section, an evanescent filter is designed, achieving a high tuning range and a large spurious free range. This filter is modified to accommodate non-resonant resonators to implement the couplings, resulting in a tunable filter with constant bandwidth in its different configurations. Finally, the third section presents two reconfigurable filters made with movable dielectric circular cylinder resonators. These filters are suitable for applications where a high tuning range and a high-quality factor is required. These filters achieve quality factors over $2 \cdot 10^4$ at Ka-band, almost in all their tuning range.
- **Chapter 7: Novel lowpass rectangular waveguide filters using dielectric and magnetic materials.** This chapter aims to develop novel lowpass filters for satellite communication systems, where high-power adverse effects can reduce the performance or even damage the filter. The main concept is to employ dielectric materials in areas of the filters, where an electron avalanche is more likely to occur, due to the vacuum conditions. Because free electron movement is not possible inside a dielectric material, this will avoid that the multipactor effect occurs in areas filled with dielectric materials. Moreover, a full-inductive lowpass filter is introduced by using dielectric materials as impedance inverters. This kind of configuration avoids the fringing fields typically associated with commonly used capacitive irises. On the other hand, a lowpass filter realized with magnetic materials is presented. The magnetic material causes resonances in the impedance inverters, so some transmission zeros can be introduced to the lowpass response. I demonstrate that this properly allows for enhancing the selectivity of the filter.

Chapter 2

Design techniques for bandpass filters based on the coupling matrix

This chapter aims to review the different design procedures that will be used along the Ph.D. thesis. The first section introduces briefly the basic concepts of the coupling matrix, which is the most employed resource for designing microwave filters [Atia and Williams, 1972]. In the same section, it is reviewed a design technique that is particularly useful for realizing tunable filters [Cameron et al., 2007].

The second section will explain a design procedure based on finding equivalent circuits of RF filters, and segmenting them to build filters following a step by step procedure [Guglielmi, 1994, Martínez-Martínez et al., 2018]. Moreover, it will be demonstrated, for the first time, that this procedure can also be extended to complex topologies like quartets (implementing transmission zeros) [Romera-Pérez et al., 2019a].

The third section shows how to design filters with non-resonating resonators (NRR), which are a useful resource to achieve bandwidth tuning, as was demonstrated in [Arnod et al., 2015]. These NRRs are regular resonators, whose resonant frequencies are not placed in the passband of the filter. Physically they are placed between the primary resonators of the structure controlling all couplings of the filter. It should be mentioned that modifying couplings is a complex task as compared to introduce tuning in the resonators. Thus, this is a feasible option to achieve bandwidth reconfiguration. This section is also focused on realizing rectangular waveguide filters whose coupling structures are TE_{101} resonant modes. At last, the fourth section provides some examples of filters with NRR in order to analyze the related problems and their possible solutions.

2.1 Coupling matrix and basic design technique

The coupling matrix is a mathematical representation of a lowpass filter circuit. This formality was introduced for the first time in [Atia and Williams, 1972], and was used to design dual-mode filters. After that, the coupling matrix has been employed to design microwave filters. Furthermore, many design techniques have been developed by employing the coupling matrix as the starting point. A detailed explanation about how to obtain the coupling matrix from the order, return loss level, and finite transmission zeros positions of a certain filter transfer function can be found in [Cameron et al., 2007, Ch. 8].

A lowpass filter circuit that is represented by an equivalent coupling matrix is composed of

admittance inverters, parallel capacitors and constant susceptances, as shown in Fig. 2.1. The fixed values of capacitors are normalized to 1 F, and input/output ports impedances are also normalized to 1 Ω . On the other hand, the values of the admittance inverters of the circuit are the coupling matrix off-diagonal elements (M_{mn}), while the constant susceptances are the main diagonal elements of the coupling matrix ($B_m = M_{mm}$).

Figure 2.1: Equivalent circuit represented by the coupling matrix. The circuit represents a normalized lowpass filter, which is composed of parallel capacitors and constant susceptances. These components are coupled by admittance inverters, whose values correspond to the coupling matrix elements. The coupling matrix can also represent the dual network composed of series inductances, impedance inverters and constant reactances. The constant susceptance values are the main diagonal of the coupling matrix. The input and output ports impedances are normalized to 1 Ω .

In [Cameron et al., 2007, Ch. 14] a design technique is presented where the coupling (M) between two resonators is computed as:

$$M = \frac{f_c}{BW} \cdot \frac{f_e^2 - f_o^2}{f_e^2 + f_o^2}, \quad (2.1)$$

being f_c the center frequency of the desired filter, BW the bandwidth, f_e the even mode resonant frequency and f_o the odd mode resonant frequency.

For a microwave structure with two coupled resonators, the fastest way to compute the resonant mode frequencies is to employ an eigensolver (resonant mode frequency analysis provided in commercial software packages) and adjust the coupling level with the selected admittance inverter. However, this procedure is not appropriate for tunable filters with varactor diodes, due to the multiple modes that could be excited in this kind of structure. An alternative is to employ a circuit like the one presented in Fig. 2.2, whose resonators are coupled by an admittance inverter of value M_{mn} . In this circuit, the input and output couplings are set to a low value that does not affect the resonant frequencies of those resonators. Particularizing the technique for rectangular waveguide technology, a rectangular waveguide structure with two coupled resonators (see Fig. 2.3) is used to compute the normalized coupling coefficient given in (2.1) as a function of a relevant geometrical parameters (windows width w_{ij} in Fig. 2.3).

In order to obtain the input (M_{S1}) and output (M_{nL}) couplings, the external quality factor is computed as:

$$Q_{ext} = \frac{f_c}{M_{S1}^2 BW}. \quad (2.2)$$

It should be noted that Q_{ext} can be either calculated with M_{S1} or M_{nL} , considering a symmetric filter. To calculate Q_{ext} , the bandwidth of a resonator with two identical irises measured at -3 dB reference level (BW_{-3dB}) is measured, and then we compute the resonance is at center frequency f_c presents an external quality factor:

Figure 2.2: Equivalent circuit, composed of two resonators coupled by three admittance inverters, to compute the value of M_{mn} . The input/output couplings M_{S1} and M_{2L} of this circuit are set to arbitrary low values. Thus, the load effect produced by the input/output couplings is minimal.

Figure 2.3: Two coupled rectangular waveguide resonant cavities. This structure is adjusted to have a similar response to the circuit depicted in Fig. 2.2. The dimensions l_i and w_{ij} are adjusted for this purpose, while w_{in} is small enough to achieve very low input/output couplings.

$$Q_{ext} = \frac{2f_c}{BW_{-3dB}}. \quad (2.3)$$

After equating (2.2) and (2.3) the bandwidth at -3 dB level can be obtained. After that, the structure presented in Fig. 2.4 is adjusted to match the center frequency f_c and -3 dB bandwidth BW_{-3dB} obtained from (2.2) and (2.3).

Figure 2.4: Structure of a resonator that has been used for evaluating the external quality factor (Q_{ext}). The dimensions w_1 and l_1 are optimized to achieve the Q_{ext} .

2.1.1 Design example

In order to show the described design technique, a fifth order filter with 20 dB of return losses at center frequency $f_c = 14$ GHz and bandwidth $BW = 100$ MHz is designed. The coupling matrix synthesized with these parameters is:

$$\mathbf{M} = \begin{pmatrix} 0 & 1.0137 & 0 & 0 & 0 & 0 & 0 \\ 1.0137 & 0 & 0.8653 & 0 & 0 & 0 & 0 \\ 0 & 0.8653 & 0 & 0.6357 & 0 & 0 & 0 \\ 0 & 0 & 0.6357 & 0 & 0.6357 & 0 & 0 \\ 0 & 0 & 0 & 0.6357 & 0 & 0.8653 & 0 \\ 0 & 0 & 0 & 0 & 0.8653 & 0 & 1.0137 \\ 0 & 0 & 0 & 0 & 0 & 1.0137 & 0 \end{pmatrix}. \quad (2.4)$$

The filter is designed in standard rectangular waveguide technology (WR-62) ($a = 15.78$ mm and $b = 7.89$ mm). Fig. 2.5 illustrates the geometry of the filter. This filter is implemented by using the fundamental resonant mode TE_{101} . The sketch also shows the dimensions that need to be calculated to implement the filtering function.

Figure 2.5: Structure of a fifth order rectangular waveguide filter. This filter is implemented in standard rectangular waveguide (WR-62) whose dimensions are $a = 15.7899$ mm and $b = 7.899$ mm. The thickness corresponding to all irises is 1 mm.

By manually adjusting the structures presented in Figs. 2.3 and 2.4, the dimensions of Table 2.1 are obtained. However, the resonant cavity lengths should be readjusted in order to correct their resonant frequencies due to loading effects of coupling windows. This refinement could be performed by using a structure similar to the one presented in Fig. 2.4. In this way, the irises are set to the dimensions calculated before. Then, the lengths of the rectangular cavity are adjusted to match the center frequency of the filter. This refinement leads to the following lengths: $l_1 = 13.4$ mm, $l_2 = 14.2$ mm and $l_3 = 14.25$ mm. It is evident from these lengths that this is a necessary step, due to the significant change, specially in the first resonator length.

The scattering parameters magnitudes retrieved for this filter are shown in Fig. 2.6. It is clear from the results that the filter is neither at the desired center frequency nor it is equalized. However, it provides starting dimensions that can be employed for a fine tuning of the filter. Even though the design process can be cumbersome, we should remark its convenience to design tunable filters, where it is easier to employ scattering parameter analysis than eigenmode solutions. This is due to multiple resonant modes computed at frequencies close to the employed resonant mode of the filter.

	Length (mm)
w_1	6.1
w_2	3.22
w_3	2.93
l_1	12.67
l_2	14.03
l_3	14.07

Table 2.1: Initial dimensions for the filter. These dimensions are obtained after optimizing the structures presented in Figs. 2.2 and 2.4 to achieve required inter-resonator couplings and Q_{ext} .

Figure 2.6: Scattering parameters magnitude for the filter presented in Fig. 2.5 and the filter described by the coupling matrix (2.4).

2.2 Step by step design technique and the scaled coupling matrix

This design technique [Guglielmi, 1994] aims to obtain a filter following a step by step procedure. As the main advantage of this technique, the elements of the structure introduced in each step are taken into account in the subsequent steps. This leads to better initial dimensions for the filter as compared to the previously presented design technique. A fine optimization can be performed at the end of the procedure if it is required.

The coupling matrix, employed in the previous section, models a lowpass filter circuit. The key idea is to find another coupling matrix that corresponds to a bandpass filter. Furthermore, this bandpass filter circuit must be composed of resonators, whose frequency response is similar to the real physical resonators that will implement the filter. After that, this technique produces a realistic equivalent circuit of the desired filter. Then, this circuit will be segmented in some fragments. Once the original circuit has been divided in several pieces, evaluating these fragments will produce partial responses, that will be later used as a goal function to optimize the physical structure. Therefore, the segments of the real filter that corresponds to that fragments of the equivalent circuit will be optimized, using the previously obtained goal functions. The key concept is to concatenate segments one by one, until all dimensions of the filter are optimized. In order to clarify the nomenclature, this new coupling matrix for bandpass filters will be known as scaled coupling matrix, while the other one will be called normalized coupling matrix. Also, it should be mentioned that the theory presented in this section is focused on rectangular waveguide filters.

The scaled coupling matrix describes a lowpass filter, as presented in Fig. 2.7(a). The main differences with the circuit of Fig. 2.1 are the scaled input/output admittances (Y_S, Y_L), and the scaled capacitors (C_m). In order to maintain the same lowpass response, the values of the admittance inverters (off-diagonal elements of the coupling matrix (M'_{sm})) and the constant susceptances (main diagonal elements $jB'_m = M'_{mm}$) must be scaled. The bandpass circuit of Fig. 2.7(b) is obtained by simply applying a lowpass to bandpass transformation described in [Cameron et al., 2007].

(a)

(b)

Figure 2.7: Equivalent circuit corresponding to the scaled coupling matrix. (a) Lowpass filter circuit. (b) Bandpass filter circuit. Comparing the circuit in (a) to the circuit of Fig. 2.1, the input/output admittances are now scaled to Y_S/Y_L , while the capacitors are scaled to C_m . (b) Circuit obtained after applying the lowpass to bandpass transformation.

The key to the scaling process is to find an adequate capacitor (C_m) to scale the lowpass filter. To do this, the equivalence between a lumped resonator and the half wavelength

open-ended transmission line, displayed in Fig. 2.8, is employed. Fig. 2.8(a) illustrates an equivalent rectangular waveguide resonator for the fundamental resonant mode TE_{101} , which is the mode employed for the filter example presented in Fig. 2.5. And Fig. 2.8(b) represent the equivalent lumped resonator. It should be noted that the length of the rectangular waveguide resonator is set to:

$$l = \frac{\lambda_{g0}}{2}, \quad (2.5)$$

being λ_{g0} the guided wavelength inside the rectangular waveguide at center frequency. The lumped resonator illustrated in Fig. 2.8(b) is calculated to be similar to the physical waveguide resonator.

Figure 2.8: (a) Open ended $\lambda_{g0}/2$ transmission line resonator. (b) Lumped resonator. Both resonators must have a similar electrical response around the center frequency.

This equivalent lumped resonator must have the same resonant frequency and slope parameter as the waveguide resonator at the center frequency. Thus, the lumped capacitor and inductor are calculated as:

$$C_{r_m} = \frac{\beta_1}{\omega_0 \omega_{c1} \mu_0} \tan(\beta_1 l) \left(\frac{\omega_{c1}}{\omega_0} - \frac{\omega_0}{\omega_{c1}} \right)^{-1}, \quad (2.6)$$

$$L_{r_m} = \frac{1}{\omega_0^2 C_{r_m}}, \quad (2.7)$$

being β_1 the propagation constant of the fundamental mode in the rectangular waveguide, computed at the lower cutoff frequency of the passband (ω_{c1}), ω_0 the angular center frequency of the filter and l the physical length of the rectangular waveguide resonator modeled as an open ended transmission line.

Finally, applying a bandpass to lowpass transformation, the lowpass capacitor can be computed as:

$$C_m = \omega_0 F_B C_{r_m}, \quad F_B = \frac{\omega_{c2} - \omega_{c1}}{\omega_0}, \quad (2.8)$$

being F_B the fractional bandwidth of the filter and ω_{c2} the upper cutoff frequency of the filter. After that, the new admittance inverters can be scaled, taking into account the

lowpass capacitors and the input/output admittances as:

$$M_{Sm}'^2 = Y_S C_m M_{Sm}^2, \quad (2.9a)$$

$$M_{mn}'^2 = C_m C_n M_{mn}^2, \quad (2.9b)$$

$$M_{nL}'^2 = C_n Y_L M_{nL}^2. \quad (2.9c)$$

In case $m = n$ (main diagonal of the coupling matrix), (2.9b) is also valid. Thus, $M_{mm}' = B_m'$ are scaled applying that same formula. With these steps, the scaled coupling matrix and their equivalent circuit are fully determined. Now, the final step is to segment the equivalent bandpass circuit and design the physical rectangular waveguide filter. However, the segments of the circuit have to be chosen in a way that allows adding parts of the structure sequentially, until all filter dimensions are completely determined.

2.2.1 Design example

In order to compare to the previous design technique, a filter with the same specifications that the one presented in Section 2.1.1 is synthesized. Thus, the normalized coupling is the same as matrix (2.4), and also the structure of the filter is shown in Fig. 2.5. By employing the design parameters of the filter and the rectangular waveguide in (2.6) and (2.7), the values of the parallel resonator are found ($C_{rm} = 64.44$ fF and $L_{rm} = 2$ nH). The lowpass capacitor is obtained applying the bandpass to lowpass transformation of (2.8), resulting in $C_m = 40.51$ μF . Finally, the coupling matrix is scaled with (2.9), leading to the following result:

$$\mathbf{M} = 10^{-3} \cdot \begin{pmatrix} 0 & 0.2851 & 0 & 0 & 0 & 0 & 0 \\ 0.2851 & 0 & 0.0351 & 0 & 0 & 0 & 0 \\ 0 & 0.0351 & 0 & 0.0258 & 0 & 0 & 0 \\ 0 & 0 & 0.0258 & 0 & 0.0258 & 0 & 0 \\ 0 & 0 & 0 & 0.0258 & 0 & 0.0351 & 0 \\ 0 & 0 & 0 & 0 & 0.0351 & 0 & 0.2851 \\ 0 & 0 & 0 & 0 & 0 & 0.2851 & 0 \end{pmatrix}. \quad (2.10)$$

The first step is to optimize the input coupling. Fig. 2.9 shows the transmission parameters of the rectangular waveguide (blue line) and the circuit response utilized as goal function (red circles). In this step, the response is optimized to match the corresponding value at the center frequency of the filter. The full-wave software FEST3D has been employed for evaluating and optimizing all the steps of this filter. The inset illustrates the 3D structure that was optimized, which is composed of only one coupling window (the input coupling). It is important to mention that an ideal impedance inverter has a response that is constant as a function of the frequency. However, the response of a waveguide iris cannot be constant as a function of the frequency, due to the dispersion of the waveguide. Nevertheless, this is not a major problem for small bandwidths. After this step, the dimension w_1 is known.

The second step is to add the first resonator (l_1) and the second coupling iris (w_2) (see Fig. 2.5). Fig. 2.10 represents the transmission parameter from the full-wave simulation

Figure 2.9: Transmission parameters of the first step. The circuit response is used as a goal function. The inset shows the optimized waveguide structure corresponding to the first segment of the circuit.

(blue line) and the circuit response (red circles) employed as a goal function. As can be observed, the response of the physical structure is very similar to the one obtained from the circuit. The inset depicts the optimized structure corresponding to the current segment (a resonator and two coupling windows). In this step, the dimensions l_1 and w_2 are optimized.

Figure 2.10: Transmission parameters of the second step. The circuit response is used as a goal function. The inset illustrates the first resonator coupled with two different irises.

As the third step, the second resonator (l_2) and the third iris (w_3) are added to the structure. Fig. 2.11 displays the transmission parameters of the full-wave simulation of 3D structure and the circuit response, used as a goal function. The inset illustrates the structure under analysis in this step. The dimensions l_1 and w_2 were included after the first optimization in order to obtain an accurate match to the goal function. This fine optimization of the previous geometrical parameters can be performed in all the steps if needed.

Figure 2.11: Transmission parameters of the third step. The waveguide response is evaluated from the structure of the inset, where the second resonator and third iris have been added.

Finally, the last step is to calculate the remaining dimensions corresponding to the third resonator (l_3) and the fourth coupling iris (w_4). However, because the symmetry of the filter structure, this last iris (w_4) will be discarded. Fig. 2.12 displays the transmission parameters obtained for this fourth segment. As it is evident, these transmission parameters last step diverges only a bit at the lower and higher frequencies represented.

Figure 2.12: Transmission parameters of the fourth step. The waveguide response is evaluated from the structure of the inset, where the third resonator and a fourth iris have been added.

At this point, all the dimensions of the filter are known. Table 2.2 collects the dimensions of the filter obtained in each step. There are some remarkable facts that can be deduced from the analysis of these data. In the first place, the input coupling controlled by w_1 does not change from the optimized value in the first step. Secondly, in the third step, the values of l_1 and w_2 were optimized after the first optimization of the new parameters in that step. However, this optimization did not produce a significant change in the previously obtained value. Thus, the optimization does not require a high computational cost. It should be reminded that the coupling matrix theory is a mathematics model

for narrow band filters. Therefore, for wideband filters a higher discrepancy between the steps and the final values of the filter is expected. Finally, with the dimensions obtained in the fourth step, the scattering parameter magnitudes plotted in Fig. 2.13 are retrieved. As it is evident from the figure, this is an exceptional result, where the full-wave response almost matches perfectly the circuit response.

	Step 1	Step 2	Step 3	Step 4
w_1	6.028	6.028	6.028	6.08
l_1		13.426	13.421	13.421
w_2		3.076	3.078	3.078
l_2			14.211	14.211
w_3			2.791	2.791
l_3				14.247

Table 2.2: Dimensions obtained in the different steps of the design procedure, following the scheme of Fig. 2.5. Dimensions are given in mm.

Figure 2.13: Scattering parameter magnitudes of the filter depicted in Fig. 2.5. The dimensions of the filter are the ones obtained in the fourth step of Table 2.2. A global optimization was not needed in this case.

2.3 Step by step procedure for folded canonical topologies

As was demonstrated in Section 2.2, the presented step by step design technique achieves a rectangular waveguide filter structure in only four steps, performing optimizations of two variables for each step. Then, this section describes a strategy to apply a step by step procedure for folded canonical topologies, such as trisections, quartets, etc. This strategy is published in a journal [Romera-Pérez et al., 2019a]. Folded canonical topologies, whose particularity is there are couplings between non-adjacent resonators, can be

designed following a small modification of the design technique presented before. This small modification consist in using multiple ports in order to consider all paths that the signal can follow. This allows optimizing in each step the direct couplings and the cross-couplings that are required by the topology. It should be mentioned that the synthesis of the scaled coupling matrix is performed following the same methodology presented in the previous section. Here, a fourth order filter with a quartet topology is designed in order to show a practical example. The quartet topology presents two transmission zeros (real or complex depending on the configuration) due to the cross-coupling from resonator 1 to resonator 4. The design parameters are center frequency $f_c = 11$ GHz, bandwidth $BW = 300$ MHz and return losses = 15 dB. These parameters synthesize the following normalized coupling matrix:

$$\mathbf{M} = \begin{pmatrix} 0 & 0.9017 & 0 & 0 & 0 & 0 \\ 0.9017 & 0 & 0.7411 & 0 & 0.2483 & 0 \\ 0 & 0.7411 & 0 & -0.7576 & 0 & 0 \\ 0 & 0 & -0.7576 & 0 & 0.7411 & 0 \\ 0 & 0.2483 & 0 & 0.7411 & 0 & 0.9017 \\ 0 & 0 & 0 & 0 & 0.9017 & 0 \end{pmatrix}. \quad (2.11)$$

The standard rectangular waveguide WR-75 ($a = 19.05$ mm and $b = 9.535$ mm) is employed for this filter example. Thus, applying the equations (2.6) and (2.7), the equivalent lumped resonator is $C_r = 0.086$ pF and $L_r = 2.42$ nH. After that, the lowpass capacitor is obtained from equation (2.8), giving $C_m = 163.05$ nF. Finally, the scaled coupling matrix is calculated from (2.9) giving:

$$\mathbf{M}' = 10^{-3} \cdot \begin{pmatrix} 0 & 0.4959 & 0 & 0 & 0 & 0 \\ 0.4959 & 0 & 0.1208 & 0 & 0.0405 & 0 \\ 0 & 0.1208 & 0 & -0.1235 & 0 & 0 \\ 0 & 0 & -0.1235 & 0 & 0.1208 & 0 \\ 0 & 0.0405 & 0 & 0.1208 & 0 & 0.4959 \\ 0 & 0 & 0 & 0 & 0.4959 & 0 \end{pmatrix}. \quad (2.12)$$

This matrix corresponds to the equivalent circuit illustrated in Fig. 2.14. This equivalent circuit will be segmented to optimize, in each step, a small number of geometrical parameters. The top and side view of the proposed filter with a quartet topology is presented in Fig. 2.15. This figure shows all the geometrical parameters that need to be obtained.

At this point, a strategy is defined to add sequentially only a few parameters to optimize in each step. Fig. 2.16 illustrates the segments of each step. The first step, shown in Fig. 2.16(a), is used to optimize the input coupling. The second and third steps (Figs. 2.16(b) and (c)) require to define three ports and to monitor two different transfer functions. In these steps, the third port is employed to monitor the path that goes through the cross-coupling M_{14} , while the second port is at the end of the main path. The two transfer functions S_{21} and S_{31} of these steps are optimized at the same time. This fact is the key difference that we propose to efficiently design filters in folded canonical topologies. The last step needs only two ports and optimizes one transfer function (goal from the circuit), as can be seen in the fourth step (see Fig. 2.16(d)).

The figures from 2.17 to 2.20 present the transmission parameters for all steps after optimizing. The markers represent the circuit responses, used as a goal functions to

Figure 2.14: Equivalent circuit of the filter described from the scaled coupling matrix (2.12).

Figure 2.15: a) Top view of the filter and its geometrical parameters b) Side view of the rectangular waveguide filter. The thickness of all irises is set to $l_i = 1$ mm.

Figure 2.16: Segments of the circuit shown in Fig. 2.14. a) This circuit allows optimizing the input coupling of the physical structure. b) With this circuit, the first resonator and the coupling 1-2 and the cross-coupling 1-4 are added to the physical structure and optimized. c) The circuit adds the second resonator and the coupling 2-3. d) The last segment of the circuit, employed for refining the couplings 2-3 and 1-4. To do this, the third resonator and the coupling 3-4 is added to the circuit of the previous iteration.

optimize the waveguide geometry. Also, the insets display the 3D structures that have been optimized in each step. As the main difference with respect to the previously designed inline filter, the responses in the second and third steps (Figs. 2.18 and 2.19) of the parameters $|S_{21}|$ and also the $|S_{31}|$ (cross-coupling) are optimized. It is remarkable that even the most complex steps (3 and 4) show an excellent agreement with the circuit response.

The dimensions of the filter are collected in Table 2.3. The table also contains the values of the dimensions obtained after each iteration. It should be remarked that the relative error between the dimensions, obtained after step 4, and the dimensions of the filter after a global optimization is always less than 0.6 %. In Fig. 2.21 it is displayed the scattering parameters of the filter for three different configurations: the circuit model, the initial configuration obtained with the design procedure, and the response after a global optimization. As it is evident from the figure, the initial structure almost matches the desired specifications, resulting in a slight mismatch in the return loss level. Global optimization, based on gradient algorithms, is performed to obtain the final dimensions of the filter. Due to the exceptional starting dimensions, the optimization is not trapped in local minima.

In conclusion, this extension of the step by step design technique, described in this section, allows designing folded canonical filters. The multiple ports enable the possibility of monitoring the different paths of the signal, being a key factor in the design process of folded canonical filters. Therefore, this design technique is really interesting to design filters, which require transmission zeros in their response (real or complex). The main reason, as it has been shown, is that the dimensions, optimized by the technique, almost match initial design specifications.

Figure 2.17: Transmission parameters of the first segment (see Fig. 2.16(a)) of the filter, depicted in Fig. 2.15. The inset shows the 3D geometry corresponding to this step.

Figure 2.18: Transmission parameters of the second segment (see Fig. 2.16(b)) of the filter, shown in Fig. 2.15. The panel shows the optimized 3D structure of step 2.

Figure 2.19: Transmission parameters of the third segment (see Fig. 2.16(c)) of the filter. The inset shows the 3D geometry after including the third resonator.

Figure 2.20: Transmission parameters of the fourth segment (see Fig. 2.16(d)) of the filter. The inset shows the structure for this step. This structure introduces the iris connecting the third and fourth resonator.

	Step 1	Step 2	Step 3	Step 4	Final	Error (%)
a_1 (mm)	9.114	9.038	9.038	9.038	9.056	0.20
a_2 (mm)		5.574	5.517	5.517	5.548	0.60
a_3 (mm)			9.407	9.359	9.382	0.24
a_4 (mm)		3.923	3.923	3.923	3.929	0.20
l_1 (mm)		17.228	17.249	17.249	17.226	0.13
l_2 (mm)			18.565	18.553	18.542	0.05

Table 2.3: Dimensions of the filter. This table displays the parameters optimized in all steps and the values of the dimensions, after a global optimization of the whole structure. The last column shows the relative error between the final dimensions (obtained after a global optimization) and the dimensions of step 4.

Figure 2.21: Scattering parameters magnitudes of three filters. The circuit response, obtained from the scaled coupling matrix. The initial filter is built with the parameters that were optimized in the steps of our design technique. And the final response of the filter, obtained from a global optimization of the whole structure.

2.4 Design technique for filters with Non-Resonating Resonators (NRR)

The filters, that have been previously discussed, are composed of resonators and couplings. An alternative is to also use resonators as couplings [Arnold et al., 2014b], as shown in Fig. 2.22. This concept is really interesting for designing tunable filters, where modifying the couplings is a challenge. These common elements (NRR) resonate at a central frequency out of the passband of the filter. In other words, these resonators, used as couplings, have to be tuned out of the passband, otherwise they will act as a traditional resonator. Tunable filters are usually based on dynamically modifying the central frequency of their constituent resonators. The main issue, that presents this kind of filters, is the difficulty to keep a constant absolute bandwidth instead of a relative bandwidth as the center frequency is varied. To maintain a constant absolute bandwidth or being able to adjust the filter bandwidth, dynamically tunable couplings are needed. Then, the use of resonators as couplings could be a solution to this problem. This concept allows us to adjust the central frequency of the filter by tuning the main resonators and corrects the bandwidth by tuning the resonators that act as couplings. Tuning the coupling resonators will affect to the adjacent resonators, but this effect can be compensated by modifying the main resonators.

In order to design this kind of filter, we can use the scaled coupling matrix that characterizes the couplings and provides the transmission parameter $|S_{21}|$ level required to implement them. For our first approach, a rectangular waveguide resonator (using the fundamental resonant mode TE_{101}) is used as NRR. Fig. 2.23 depicts the rectangular waveguide resonator and their geometrical parameters. With the aid of the aforementioned MATLAB-HFSS API, we can compute some curves of resonators fixed at the same frequency, but with different input couplings. Thus, the curves are obtained by sweeping the irises width (w) from a small value to the width of the employed standard

Figure 2.22: Two equivalent structures. The first one is a resonator with two couplings on both sides. The second one obtains the same coupling level of the first one by replacing the impedance inverters with two impedance inverters and a resonator acting as a coupling.

waveguide, while the length of the resonator (l) is corrected by optimizing to maintain the resonant frequency at the desired one.

Figure 2.23: Structure of a rectangular waveguide resonator and its geometrical parameters. This structure is employed to compute the curves obtained in Fig. 2.24.

To illustrate the design process, a fourth order filter with center frequency $f_c = 16$ GHz, bandwidth $BW = 100$ MHz and return losses $RL = 15$ dB is designed. The standard waveguide is WR-62 ($a = 15.799$ mm and $b = 7.899$ mm) and the NRRs are set to be centered at frequency 14 GHz. The filter is implemented with the fundamental resonant mode TE_{101} for both, main and non-resonating resonators. Fig. 2.24 plots some of the curves obtained after running the MATLAB script, for this particular example. Thus, each curve is linked to a pair (w, l) in order to implement the required NRR, giving different coupling levels for a fixed resonant frequency of the NRR.

By using this data, it is possible to obtain the graphs of Figs. 2.25 and 2.26. The first one represents the transmission parameter of this non-resonant resonator as a function of the width of the input/output coupling evaluated at the center frequency of the desired filter (16 GHz). It should be reminded that the transmission parameter level can be correlated to a coupling level through the scaled coupling matrix explained in Section 2.2. Thus, this calculation can be performed mathematically by calculating the transmission parameter of an inverter (or obtained from a circuit simulation). Then, knowing the required $|S_{21}|$ for the inverters, in Fig. 2.25 can be found the iris widths (w) that implement the inverters for our filter. The second one is the length of the resonator l associated with each value of the width w , maintaining the resonant frequency at 14 GHz (f_{NRR}).

Figure 2.24: Transmission parameters of five different configurations of the structure presented in Fig. 2.23.

Figure 2.25: Transmission parameter obtained for each width w after adjusting the length of the resonator (see Fig. 2.23) to keep constant the resonant frequency. This transmission parameter is evaluated at the filter center frequency (16 GHz).

Figure 2.26: Lengths of the coupling resonators adjusted for each width (w) of the structure represented in Fig. 2.23 to keep the same resonant frequency of the NRR.

Using the filter specifications and waveguide sizes, the following the procedure of Section 2.2, scaled coupling matrix is synthesized:

$$\mathbf{M}' = 10^{-3} \cdot \begin{pmatrix} 0 & -0.2486 & 0 & 0 & 0 \\ -0.2486 & 0 & 0.0285 & 0 & 0 \\ 0 & 0.0285 & 0 & 0.0285 & 0 \\ 0 & 0 & 0.0285 & 0 & 0.2486 \\ 0 & 0 & 0 & 0.2486 & 0 \end{pmatrix}. \quad (2.13)$$

By calculating the required $|S_{21}|$ to implement the couplings of (2.13) (computing the equivalent circuit in a software or using the formula of the transmission parameter of an admittance inverter), and using the data of Figs. 2.25 and 2.26, we can find the geometrical parameter of the non-resonating resonators. In Table 2.4 are collected the geometrical parameters which implement the filter, alongside the required $|S_{21}|$ for each coupling.

Coupling	required $ S_{21} (\text{dB})$	$w(\text{mm})$	$l(\text{mm})$
M'_{S1}, M'_{3L}	-9.88	7.42	10.45
M'_{12}, M'_{23}	-25.46	5.44	10.93

Table 2.4: Transmission parameters required by the scaled coupling matrix (2.13), iris widths and resonator lengths to associated to each NRR.

At this point, only remains to calculate the length of the main resonators of the filter. The circuit model of Fig. 2.27 represents the structure of Fig 2.23. In this circuit, the NRR resonator is acting as a common inverter and the electrical lengths of the input and output ports are 90° , which is the half of the main resonator electrical length ($\lambda_g/2$ evaluated at the center frequency of the filter, for the fundamental resonant mode TE_{101}). It should be remarked that the inverter phase can be either -90° (inductive coupling $f_{NRR} > f_c$) or $+90^\circ$ (capacitive coupling $f_{NRR} < f_c$) depending on the type of coupling.

Figure 2.27: Equivalent circuit of the structure represented in Fig. 2.23. The NRR is modeled as a common inverter, which can be inductive or capacitive. The transmission lines represent half of the main resonators.

Therefore, the physical lengths of the input and output ports of the NRR structure (see Fig. 2.23) are optimized to match the transmission parameter phase, obtained from adding the phase of the transmission lines plus the inverter phase as:

$$\angle S_{21} = -90^\circ - 90^\circ - 90^\circ = +90^\circ \text{ if } f_{NRR} > f_c, \quad (2.14)$$

$$\angle S_{21} = -90^\circ + 90^\circ - 90^\circ = -90^\circ \text{ if } f_{NRR} < f_c, \quad (2.15)$$

being the obtained lengths the half electrical lengths of the main resonators. Finally, the filter can be built by joining the NRR and their optimized input/output ports in order. Table 2.5 collects the lengths of the main resonators for this filter example.

Main resonators	Resonator length (mm)
1,3	21.39
2	21.86

Table 2.5: Resonator lengths obtained after optimizing the phase of the transmission parameter.

Finally, building the filter with these dimensions, the response of the filter is presented in Fig. 2.28. The filter passband is located at 16 GHz as it should. However, it is evident that the NRRs have been combined in a spurious passband at 14 GHz, resulting in 2 GHz of spurious free range (SFR)¹.

Figure 2.28: Scattering parameter magnitudes of the filter made of non-resonating resonators as couplings. The response also shows the spurious passband produced by the combination of all the non-resonating resonators.

Summarizing, this technique can be performed following these steps:

1. Execute the optimization between MATLAB and HFSS in order to obtain the design curves to measure the couplings.
2. Interpolate the results and choose the parameters that match with the levels that have been obtained from the scaled coupling matrix.
3. Perform a sweep or optimization to obtain the phase at the central frequency that meets $\angle S_{21} = -90^\circ$ for an inductive coupling ($f_{NRR} > f_c$) or $\angle S_{21} = +90^\circ$ for a capacitive coupling ($f_{NRR} < f_c$).

¹The spurious free range (SFR) is the range of frequencies between the filter passband and the nearest spurious bands or resonances near it. This parameter is a measure of the cleanliness of the filter response around its passband. A large SFR is always desirable for a microwave filter.

4. Finally, we can gather each previously calculated coupling in order to obtain an initial dimensions for the filter.

Furthermore, these coupling resonators can provide positive and negative couplings in order to design more complex topologies, as it is presented in [Arnod et al., 2015].

2.5 Rectangular waveguide filters made using non-resonant resonators

Following the same procedure to design filters with non-resonating resonators (NRR), we are going to investigate about the possible configurations that can be used. The first configuration was presented in the previous section, where all NRR were designed for the same frequency.

2.5.1 Filter with non-resonating resonators operating at two different frequencies

An alternative is to take different frequencies, trying to avoid coupling between non-resonating resonators. As an example, a fourth order filter is designed. The center frequency of the filter is $f_c = 16$ GHz and the resonators that act as couplings are placed at 14 GHz and 17 GHz. Even though it is possible to use this configuration in an inline filter, for a topology like quartet, trisections or other complex topologies, the frequency of the NRR cannot arbitrarily be chosen. Thus, for these complex topologies, the frequency of the NRRs will be determined to satisfy the signs imposed by the coupling matrix, being the frequency of the NRR above the center frequency for a positive couplings or below the center frequency for a negative couplings. Nevertheless, it is still useful to study how it would affect to the spurious response of the filter. Fig. 2.29 shows the structure of the designed filter and Table 2.6 collects the dimensions for each resonator in the structure.

Figure 2.29: Fourth order filter geometry with non-resonating resonators as couplings. The main resonators are marked in red color while the NRR are marked in blue and green for NRR at 14 GHz and 17 GHz, respectively.

NRR 1	l	11.87
	w	7.29
MR 1	l_1	9.96
NRR 2	l	10.05
	w	4.08
MR 2	l_2	10.67
NRR 3	l	13.36
	w	4.66

Table 2.6: Dimensions of the filter depicted in Fig. 2.29. The dimensions are given in mm. The structure of the filter is symmetrical from input to output.

As it is shown in Fig. 2.30, this filter does not present a spurious band, as was noticed in the filter presented in the previous section. However, there exists a spurious resonance near 17 GHz. Despite this spurious, its bandwidth is considerably reduced if compared with the obtained response for the filter in Fig. 2.28.

Figure 2.30: Scattering parameter magnitudes for a fourth order filter made with non-resonating resonators. The filter passband can be seen centered at 16 GHz, while the NRR are located at 13.5 GHz, 14 GHz and 17 GHz, approximately.

The conclusion is that, even if it is not possible to completely avoid the spurious resonances, the use of multiple frequencies for implementing the NRRs leads to narrower spurious bands.

2.5.2 Symmetric non-resonating resonators filter

Since the previous attempt only shows a narrow resonance that reaches a considerable transmission level, the next effort is to choose additional different frequencies for the NRR in order to finally avoid the spurious resonances near the passband. Three different

frequencies (12, 12.5 and 13 GHz) are employed to implement the NRRs. The filter specifications are: order $N = 8$, return losses $RL = 15$ dB, center frequency $f_c = 11$ GHz and bandwidth $BW = 500$ MHz. The filter is realized in standard rectangular waveguide (WR-75) (width $a = 19.05$ mm and height $b = 9.525$ mm) and uses the TE_{101} resonant mode for both main and coupling resonators. The coupling diagram and the frequencies used for them are displayed in Fig. 2.31. The circles correspond to the main resonators, whereas the lines represent the couplings, which are implemented by non-resonating resonators. Some considerations have to be taken into account. First, the frequencies are selected to be relatively far from the center frequency. Secondly, the frequencies are interspersed to minimize the coupling between them. The 3d structure of this filter is displayed in Fig. 2.32, and Table 2.32 contains geometrical parameters of the filter.

Figure 2.31: Coupling diagram for a symmetric filter. The circles represent the resonators that compose the filter (main resonators), source and load, while the lines represent the couplings. The numbers above the lines that represent the couplings are the employed frequencies for implementing NRRs.

Figure 2.32: Eighth order filter geometry with non-resonating resonators as couplings. The main resonators are marked in blue color while the NRR are marked in red, green and yellow, depending on their resonant frequencies. The NRRs frequencies follows the diagram shown in Fig. 2.31.

However, as the scattering parameters drawn in Fig. 2.33 show, there exist spurious passbands where the non-resonating resonators are defined. It should be pointed out that the obtained return loss level in the filter passband does not exactly match with the target. This occurs because the NRRs were designed separately, and they do not take into account the interaction between NRRs. Therefore, it can lead to a slightly different coupling value in all the NRRs.

NRR 1	l	11.5032	MR 1	l_1	12.3669
	w	7.779	MR 2	l_2	14.8009
NRR 2	l	8.7277	MR 3	l_3	14.9879
	w	13.0542	MR 4	l_4	15.302
NRR 3	l	8.6672			
	w	12.2537			
NRR 4	l	8.8281			
	w	11.0934			
NRR 5	l	8.1339			
	w	13.7988			

Table 2.7: Dimensions of the filter depicted in Fig. 2.32. The dimensions are given in mm. The structure of the filter is symmetrical from input to output.

Figure 2.33: Scattering parameter magnitudes of the filter with non-resonating resonators with three different frequencies interspersed between them.

2.5.3 Asymmetric non-resonating resonators filter

From previous designs, it is evident that the couplings made with non-resonating resonators interact between them, resulting in spurious passbands close to the filter passband. As it was demonstrated in section 2.5.1, the use of different frequencies for each coupling and the wide difference between the NRR frequencies minimize the spurious bands generated by them. This issue can be, even more, minimized if the filter is designed asymmetrically choosing a different frequency for each coupling present in the filter. This option should prevent the majority of interaction between the NRRs. This design has the same parameters as those used for the symmetric design, except for the return losses that were set at 20 dB. The coupling diagrams with the frequencies for each coupling are displayed in Fig. 2.34. It can be observed in the figure, that the election of the coupling frequencies was done to achieve at least 1.5 GHz difference with respect

to the nearby couplings, changing them asymmetrically along the whole filter. Also, Fig. 2.35 shows the geometry of the filter, while Table ?? contains the dimensions of each resonator and NRR that compounds the filter.

Figure 2.34: Coupling diagram for an asymmetric filter. For this filter, each NRR is resonating at a different frequency from the others in order to avoid the interaction between them. The numbers above the lines that represent the couplings are the employed frequencies for implementing NRRs.

Figure 2.35: Geometry of the filter with an asymmetrical design. The NRR are colored in green and the main resonators are colored in blue. The first and last NRR and main resonator are marked in the figure.

In Fig. 2.36, the scattering parameter magnitudes of the asymmetrical filter are shown. These parameters have spurious as expected, but there are no passbands formed by them, resulting in a moderate spurious free range. Thus, non-symmetrical filters are better than symmetrical ones for avoiding spurious passband as it is demonstrated.

NRR 1	l	8.45	MR 1	l_1	12.1
	w	11.37	MR 2	l_2	14.67
NRR 2	l	6.84	MR 3	l_3	15.32
	w	9.66	MR 4	l_4	15.40
NRR 3	l	8.86	MR 5	l_5	15.26
	w	9.09	MR 6	l_6	15.20
NRR 4	l	15.32	MR 7	l_7	14.81
	w	8.15	MR 8	l_8	12.42
NRR 5	l	7.95			
	w	9.07			
NRR 6	l	12.01			
	w	8.52			
NRR 7	l	8.15			
	w	9.12			
NRR 8	l	5.97			
	w	9.66			
NRR 9	l	5.09			
	w	12.08			

Table 2.8: Dimensions of the asymmetric eighth order filter depicted in Fig. 2.35. The dimensions are given in mm. The numbering is made from one port to another; however, these dimensions are not marked in the sketch due to the large size of the structure.

Figure 2.36: Scattering parameters magnitude of the asymmetrical filter with non-resonating resonators (see Fig. 2.35). Nine different frequencies was employed to realize the NRR of this filter.

2.6 Conclusions

This chapter provides three design processes that have been used for designing filters in this PhD. dissertation. These design processes are based on the coupling matrix, which is the most known resource for microwave filters. The first design process only uses scattering parameter simulations in order to find the dimensions of the physical couplings and resonators. This is a useful design process for tunable filters because analyzing the resonant modes with an eigensolver can lead to many modes depending on the capacitances employed (for electronic tuning). Even though the initial dimensions are not especially close to the final ones, they are good enough to refine the filter with them.

The second design process consists of a step-by-step technique. This process starts from a mathematical approximation of the physical resonator to a lumped resonator. And, finding a coupling matrix adapted to the physical resonator that is used for the real filter. The filter is built by adding small segments of the structure and optimizing them. This leads to optimizations of a small number of variables for each segment. At the end of this process, the obtained response is really close to the specifications of the filters, as it was demonstrated. Moreover, this design technique was extended for folded canonical topologies.

The third design technique presented in this chapter is employed to design filters with NRRs. As it was discussed, the use of NRRs is a suitable coupling tuning mechanism that can be used in many tunable filters. By using this design technique, several filters with NRRs have been designed in this chapter. The main purpose was to study the properties of these filters, such as spurious produced by the NRRs and alternatives to minimize their impact. As it was shown in that section, the filters with NRRs cannot avoid the spurious resonances generated by them. However, it is possible to have narrower resonances if the NRRs is designed to be at different frequencies. Although design asymmetrical filters is a hard task, it could solve the problem of the spurious resonances. Thus, choosing the appropriate frequencies for the NRR can avoid that they are coupled between them, not generating a spurious band.

Chapter 3

Substrate integrated waveguide tunable filters

Substrate integrated waveguide (SIW) [Wu et al., 2003, Deslandes and Wu, 2003, Rayas-Sanchez and Gutierrez-Ayala, 2008] is one of the most attractive technologies for tunable filters. Its low cost and small form factor, together with the obtained higher quality factor when compared to planar technologies, such as, microstrip, coplanar, etc., makes it one of the most generally used alternatives. For reconfigurable structures, it is specially a good option, since it is easy to integrate Surface Mount Devices (SMD) [Bozzi et al., 2011], that are required if a electronically tunable filter is desired. Due to this fact, there are many SIW tunable filters in scientific papers [You et al., 2016, Anand and Liu, 2015, Lee et al., 2015, Yang et al., 2014, Sekar and Entesari, 2012, Anand et al., 2013, Sirci et al., 2012, Liu et al., 2012, Sekar et al., 2011].

This chapter is focused on three main topics:

- **Tunable lowpass filter:** some test have been carried out in this kind of filters. The interest of this structure is to independently control the lowest and the highest cutoff frequencies by acting on the side wall that composes the SIW structure. For the tests carried out on this kind of structures, we have not used any specific component like varactor diodes [Lu et al., 2010] or MEMS [De Los Santos, 2002]. These test are carried out by short-circuiting post of the structure or disconnecting them from the metallic surfaces of the structure.
- **Tunable evanescent mode substrate integrated waveguide filters:** Evanescent mode tunable filters show a reasonable trade-off between tuning range, spurious free range and bandwidth reduction. A novel SIW evanescent tunable filter is proposed based on varactor diodes. The resonators are implemented as via holes disconnected from the top wall through a capacitance, placing a varactor diode to connect the via hole to the top wall. A first design is implemented in S-band achieving a 1 GHz of center frequency tuning range. Moreover, this filter was scaled to Ku-band and a prototype was finally designed by using commercial components. In addition, some tuning mechanisms are studied in order to modify the couplings.
- **Tunable waveguide filters using the studied SIW resonator modes:** This filter consist of an alternative implementation of the resonating mode, used in the previous point, in waveguide technology. The utilization of this resonant mode in waveguide requires a complex setup as compared to its SIW implementation. In

addition, the setup needs to include a solution to avoid radiation problems. However, its advantage is the wide tuning range achieved by this kind of implementation.

3.1 Tunable lowpass filter

Usually, tunable filters exhibit bandpass responses, being composed of tunable resonators. However, a lowpass filter with tuning elements can be envisaged to modify the upper cutoff frequency. Additionally, it is possible to introduce tuning elements in the walls, allowing them to reconfigure the lower cutoff frequency of the filter. These two tuning mechanisms together produce a tunable filter with bandwidth control.

3.1.1 Fundamental mode cutoff frequency variation

An analysis of a SIW structure with lateral posts, connected and disconnected to the top metal cover, is performed to study how it will affect to its lower cutoff frequency. Fig. 3.1 shows the simulated structure and its parameters. The dielectric substrate is Arlon AD1000 ($\epsilon_r = 10.2$) and its thickness is 0.65 mm. Using these parameters listed in the figure, the SIW structure presents a cutoff frequency of 3 GHz (controlled by the distance between the two rows of lateral posts W_{SIW}). The test will show the difference when the row of posts is connected or disconnected to the top metallic cover. A capacitor could be placed connecting the post and to the top metal cover to switch states between short-circuit and open circuit.

Figure 3.1: Substrate integrated waveguide structure, where one of its via-holes rows is not connected to the top metallic cover. The diameter of the side via-holes is $d = 1$ mm and the period between them is $p = 2$ mm. The width is $W_{SIW} = 20.25$ mm, the length of the SIW section is $l_{SIW} = 16$ mm and the thickness of the substrate is $h = 0.635$ mm. The gray surface is configured as PEC boundary condition (in Ansys HFSS) for connecting the post (short-circuit). On the contrary for the disconnected state (open circuit), this surface is not present in the structure, letting the post disconnected from the top wall.

As we can see in Fig. 3.2, the configuration with disconnected posts provides a cutoff frequency of the SIW near to 2 GHz, in contrast to the connected posts simulation leading to a cutoff frequency of 3 GHz.

Figure 3.2: Scattering parameters magnitude, obtained when a side row of via-hole posts are disconnected from the top wall and when all posts are connected as usual in the structure shown in Fig. 3.1.

3.1.2 Lowpass filters basic concepts

In this section, a brief review of the basic concepts related to microwave lowpass filters is performed. The objective is to provide a general background in lowpass filters (a most detailed explanation can be found in [Cameron et al., 2007]). Also, the design technique for lowpass filter is presented in [Vera-Castejón et al., 2014], while in [Máximo-Gutiérrez et al., 2019] is proposed the procedure for their implementation in SIW technology.

The circuit depicted in Fig. 3.3 represents a lowpass filter. This circuit is composed of impedance inverters, whose values K are obtained through the synthesis process for the selected design parameters and transmission lines of characteristic impedance Z_0 and electrical length θ_c . These K values are obtained from the synthesis explained in [Cameron et al., 2007]. The design parameters to start this synthesis procedure are the order of the filter N , return losses (RL) and the electrical length of the transmission lines (θ_c) that composes a stepped impedance filter.

Figure 3.3: Lowpass filter equivalent circuit. The K values are those corresponding to impedance inverters, whereas Z_0 and θ_c are the impedances and the electrical lengths of the transmission lines, respectively.

In order to design the microwave filter structure, the circuit represented in Fig. 3.3 is segmented in several pieces, like the one shown in Fig. 3.4(a). First, this fragment is evaluated in order to retrieve the S_{21} in magnitude and phase. Secondly, the lowpass segment in SIW technology, represented in Fig 3.4(b) is optimized to match the S_{21} , calculated from the circuitual model, in magnitude and phase at the cutoff frequency of

the lowpass filter. For this task, the elliptical axes (w_x and w_z) are first adjusted to match the transmission parameter magnitude $|S_{21}|$, then l is modified to match the transmission parameter phase $\angle S_{21}$. Finally, all the fragments of SIW are placed altogether to make the lowpass filter.

(a) Fragment of the lowpass filter circuit presented in Fig. 3.3. This circuit is composed of two transmission lines of electrical length $\theta_c/2$ (half of the length of the whole transmission line that joins two inverters in Fig. 3.3) connected to an impedance inverter.

(b) Fragment of a SIW lowpass filter. The ellipse in the top wall acts as an impedance inverter and its geometrical parameters control the inverter value, while l controls the transmission parameter phase. This SIW structure has the same geometrical parameters and substrate used in the one of Fig. 3.1. Other parameters are variable to implement the required inverter value.

Figure 3.4: Fragments of a lowpass filter. The fragment of the figure (a) is employed to compute the transmission parameter, while the fragment of figure (b) is optimized to have the same response of the previous one.

3.1.3 Tunable lowpass filter design

The first test that can be performed over a SIW lowpass filter is to let floating posts (disconnected to the top metallic surface) in the SIW sections that interconnects the impedance inverters, as shown in Fig. 3.5. Figure 3.5 shows the structure, including the post disconnected to the top wall on one side wall of the filter. The main idea is to control the phase of these SIW transmission line sections, placed between impedance inverters, to modify the θ_c parameter (see Fig. 3.3). The test is conducted considering floating post (as is shown in Fig. 3.5) for one state and with the post connected to the top metallic wall for the second state. The filter realized in [Máximo-Gutiérrez, 2015] has been used to prove the concept of how to modify the lowpass filter response. In [Máximo-Gutiérrez et al., 2019], a detailed explanation of the design of this kind of filters can be found. Posts, near to elliptical apertures (that act as impedance inverters), are connected to the top metal cover in order to avoid modifying the value originally obtaining during the design process.

Fig. 3.6 represents the responses of the lowpass filter in two states, the synthesized filter and the one with the post disconnected to the top wall. As can be observed comparing the responses, let the floating posts affect to the lower cutoff frequency more than to the upper cutoff frequency. Thus, to modify the upper cutoff frequency, the impedance

Figure 3.5: Structure of a lowpass SIW filter with floating post disconnected to the top wall between capacitive elliptical aperture impedance inverters. The lengths between ellipses are; $l_1 = 9.219$ mm, $l_2 = 9.438$ mm, $l_3 = 9.519$ mm and $l_4 = 9.842$ mm. The diameter of the lateral via holes is $d = 1$ mm and the period between them $p = 2$ mm. The width is $W_{SIW} = 20.25$ mm and the thickness of the substrate is $h = 0.635$ mm with relative permittivity $\epsilon = 10.2$. The major semiaxis of the elliptic impedance inverters are: $wx_1 = 1.610$ mm, $wx_2 = 2.462$ mm, $wx_3 = 2.98$ mm and $wx_4 = 3.092$ mm. The ratio between semiaxes is 1.5 for all the ellipses.

inverters need to be adjusted. However, this is not an easy task, due to the lack of tunable parameters. Therefore, a new capacitive impedance inverter needs to be introduced to have a better control of the filter response.

Figure 3.6: Scattering parameter magnitude of the filter depicted in Fig. 3.5 with lateral posts connected or disconnected to the top wall. The lower cutoff frequency has been displaced, approximately, 200 MHz, while the upper cutoff frequency is only shifted (also approximately) 80 MHz.

3.2 Tunable evanescent mode substrate integrated waveguide filters

3.2.1 Study of the evanescent resonant mode

In SIW technology, a post in the middle of the structure connected through a capacitor to the top metal cover behaves as an evanescent mode resonator. For our designs, the post is placed in a evanescent SIW section. The corresponding resonant mode has a wide range of tuning and, thanks to the dielectric-vacuum interface, RF-leakage is minimum. Fig. 3.7 displays one SIW resonator, implemented as described before, couple by the input and output ports. A capacitor in HFSS [Ansoft, 2019] is modeled as a surface, touching two metallic conductors. In our case, we have modeled the varactor diodes used as capacitor using a concentric circle surface, since it leads to a better convergence in the computational simulations and less computational cost. However, for analyzing commercial varactor diodes will be necessary to consider the shape of the physical component, and this it will be studied in final designs. Thus, we are going first to perform the study of filters with concentric circular surfaces to model capacitor, and later adapt them to rectangular surfaces with the size of commercial varactor diodes.

Figure 3.7: Structure of a resonator in SIW section with a capacitor (red surface). The employed SIW is the same as the same used for the lowpass filter design (see Fig. 3.5). The radius of the centered via hole is $d_i = 1$ mm and the surface that contains the capacitor condition is $offset = 1$ mm width. The length of the SIW section is $l_{SIW} = 39$ mm.

For the characterization of the proposed resonant mode, the behavior of this resonance is studied as a function of the capacitance. In Fig. 3.8 is represented the resonant frequency of the proposed resonant mode, the resonant frequency of this mode vary from 2 GHz to 3 GHz for the range of analyzed capacitances. Due to the SIW structure, used for this resonator, is evanescent in frequencies below 3 GHz, from 3 GHz to higher frequencies spurious resonances may appear, because the fundamental mode of the designed SIW becomes propagative.

3.2.2 Filter design

With the aid of the HFSS eigensolver information, a filter can be designed by using traditional methods. In order to simplify the structure to perform a faster simulation,

Figure 3.8: Resonant frequency as a function of the capacitance value of the structure depicted in Fig. 3.7.

the SIW structure could be analyzed by using its equivalent dielectric filled rectangular waveguide [Wu et al., 2003, Rayas-Sanchez and Gutierrez-Ayala, 2008]. This equivalent model reduces the complexity of the filter, because the posts on each side of the SIW are replaced by simple metallic walls. Finally, an structure of a third order filter, with the proposed resonators in SIW technology, is represented by Fig. 3.9. This structure keeps the evanescent waveguide section cutoff at 3 GHz, being similar to the employed SIW in the resonator analysis.

Figure 3.9: Dielectric filled rectangular waveguide evanescent filter. The filter is composed of three resonators of the same type as those introduced in Fig. 3.7. The width of the input section is $W_{SIW} = 23.48$ mm and the width of the evanescent section is $W_{eva} = 20$ mm. The substrate parameters are the same as those employed in the previous designs: $h = 0.635$ mm and $\epsilon_r = 10.2$. The length between the impedance step and the first and last resonator is $l_1 = 15$ mm, whereas the length with the second resonator is $l_2 = 35$ mm.

This structure leads to a design with a 44 MHz bandwidth, at a center frequency of 2.48 GHz, by using capacitors of approximately 7 pF. By tuning the capacitor value, the passband of the filter is shifted to lower frequencies, as can be observed in Fig. 3.10, where three different configurations of the filter are displayed. Apart from the central frequency shift, the bandwidth of the filter is also modified. This bandwidth change is due to higher couplings when frequency increases. Table 3.1 collects the capacitances,

center frequencies and bandwidths for each response of this filter. As can be observed, this configuration provides a 1 GHz tuning range.

C (pF)	f_c (GHz)	BW (MHz)	f_c variation (%)	BW variation (%)
4	2.81	102	13.26%	+31.81%
7	2.48	44	-	-
10	2.22	28	10.65%	-36.36%

Table 3.1: Center frequency and absolute bandwidth for each configuration of the filter. In addition, the table also includes the percentages of tunability range and bandwidth variation, taking the second configuration as baseline.

Figure 3.10: Scattering parameters magnitude for three different configurations of the filter. These states are obtained by changing the capacitances. The relevant characteristic of these configurations are collected in Table 3.1.

In Fig. 3.11, the out of band response of the tunable filter is presented for the three different states of the filter. There exist some spurious resonances near 3.4 GHz produced by the fundamental resonating mode of the dielectric waveguide TE_{101} (remember that the cutoff of this SIW is 3 GHz). These resonances are formed in the SIW sections between the three posts of the filter.

In conclusion, the evanescent mode resonator exhibits a high tuning range for S-band, and its RF-leakage is negligible. Moreover, the implementation of the proposed structure is easy to manufacture and can be easily integrated with electronic.

3.2.3 Tunable coupling mechanisms

The filter, that is under study, is designed in an evanescent mode SIW. The couplings in evanescent filters are typically determined by distance between resonators, assuming that the fundamental mode is attenuated, because it operates below its cutoff frequency. The cutoff of the fundamental mode in a SIW structure is controlled by the width between the side posts. Then, our propose is to find a mechanism that virtually increases the width of a SIW close to the coupling sections. In this way, a virtually wider SIW will produce

Figure 3.11: Scattering parameters magnitude for the three states of the filter and their out of band responses. The spurious passband above 3.4 GHz is due to TE_{101} resonating modes that are formed in the SIW sections between each post.

a higher coupling. The structure proposed to control the coupling is shown in Fig. 3.12. The varactor diodes, used to tune both resonances, are marked in yellow and the varactor diode, employed to control the coupling, is marked in red. Basically the coupling level would be increased when the post is not short-circuited, since it implements an equivalent wider structure in that area. On the contrary, if the capacitance of the varactor diode is high enough, it behaves as a short-circuit, thus reducing the coupling (the equivalent width becomes smaller). Intermediate values of the varactor diode capacitances gives rise to intermediate states, resulting in a continuous tuning of the coupling.

Figure 3.12: Two resonators coupled in an evanescent SIW structure with floating side post connected through a varactor diode to the top wall. This last varactor diode will control the coupling level between the resonators. The yellow surfaces are the varactor diodes used for the resonators, while red surface is the varactor diode for connecting the floating post to the top wall. The parameters of the SIW structure and the resonators are those used in Fig. 3.7. The total size of the SIW section is $l_{SIW} = 74$ mm, which is divided in the input length $l_{in} = 24.5$ mm and the distance between the two resonators is $l_{res} = 25$ mm.

This structure is used to measure the coupling level between the two resonators. The input/output coupling is set to a low value to reduce its influence over the coupling, leading to a more accurate test. Thus, the coupling level will be altered only by the variation of the third varactor that were introduced on the floating side post for this characterization. In Fig.3.13, the scattering parameter magnitudes for three states are presented. These states are obtained by changing the capacitance associated to the varactor diode on the side post. As can be observed from the scattering parameters, there exists a spurious resonance apart from the two resonances from our resonators. Furthermore, this spurious resonance is controlled by the third varactor. As can be seen in Fig. 3.13, the coupling between the resonators is altered if the capacitance is changed as expected. However, the coupling variation suggests that the change is produced by the spurious resonance. This conclusion is consistent with previous studies of non-resonating resonators that were presented in Chapter 2. Thus, the coupling level can be controlled by the same NRR mechanism. Unfortunately, at the same time, a spurious resonance is placed nearby.

Figure 3.13: Scattering parameter magnitude of Fig. 3.12 evaluated for capacitances of 5 pF, 7 pF and 9 pF for the varactor that controls the coupling level.

It is important to quantify the variation in the coupling level generated by this mechanism. Fig. 3.14 illustrates the coupling level obtained as a function of the capacitance associated to the varactor diode. Also, the spurious free range, between the main resonant frequency (around 2.5 GHz, approximately) and the spurious resonance produced by a NRR, is plotted in the graph following the right ordinate axis. This result is extracted from the results of Fig. 3.13. The coupling level is increased if the capacitor takes higher values, while the spurious resonance approaches the main resonances. This simulation provides a total 0.0407 of k-coupling¹ level variation. If we take, for instance, a 75 MHz bandwidth filter placed at center frequency 2.5 GHz, the variation of the coupling matrix level is 1.36 (remember that $M = k \cdot f_c / BW$, being k the electromagnetic coupling). This variation is more than enough to change the inner couplings of a filter for many applications. At this point, it should be pointed out that the behavior of this structure was not the expected in a first place. Initially, it was predicted a higher coupling level if the capacitance is reduced.

¹The electromagnetic coupling is a measure used to quantify the power transmitted from one resonator to a second one. For an structure with two coupled resonators, the coupling k can be computed using the formula $k = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2}$, being f_1 the lower resonance and f_2 the upper resonance.

However, the coupling level, represented in Fig 3.14, indicates the opposite trend. This is due to the fact that the coupling is actually controlled using the NRR mechanism.

Figure 3.14: Coupling level (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance for the structure in Fig. 3.12. The coupling level is increased if the spurious free range is decreased, due to the NRR behavior.

The previous structure (see Fig. 3.12) shows NRRs, which are the main reason for the coupling level change. However, the proximity of spurious resonances is not a desirable feature. Then, an alternative structure, with two posts (one on each side) disconnected from the top wall and connected through varactor diodes, is proposed in Fig. 3.15. The layout of the varactor diodes follows the same criteria as the previous one (see Fig. 3.12). This structure is proposed to study if the addition of an additional NRR can improve the coupling level variation offered by the structure with only one NRR.

Figure 3.15: Two resonators coupled in an evanescent SIW structure with two open floating side posts connected through varactor diodes to the top wall. The yellow surfaces are the varactor diodes used for the resonators, while red surfaces are the varactor diodes for connecting the floating post to the top wall. The varactor diodes on each side will control the coupling level between the resonators. The dimensions of this structure are the same as those used in Fig. 3.12, the only difference is the second capacitor.

Performing the same study as the one realized for the previous structure (Fig. 3.12), the same response is obtained in this case, as it is shown in Fig. 3.16. It is obvious from the data, that the coupling variation follows the same principle as the previous structure. However, it can be pointed out that this simulation reports a total 0.0627 k -coupling variation, which is a 54% higher than previously reported, being the SFR nearly the

same. This increase could be generated by the contribution of the two non-resonating resonators that are present in this structure. Thus, it is an advantage to use this solution instead of the previous one. The main advantage is the on the possibility to achieve the same level of coupling with a better SFR than the previous one.

Figure 3.16: Coupling level (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance (the two varactors take the same capacitance value). The coupling level variation is higher than that achieved in Fig. 3.14, while keeping a similar SFR level.

Taking into account the previous simulations, a configuration with multiple posts is now studied in order to study if the coupling variation can be improved. Consequently, a structure with three posts on each side disconnected from the top wall and connected through varactor diodes is evaluated. The structure under study is presented in Fig. 3.17. The color code follows the same criteria as the first design (Fig. 3.12).

Figure 3.17: Two resonators, coupled in an evanescent SIW structure, with three open ended floating posts on each side, connected through varactor diodes to the top metallic wall. The varactor diodes on each side will control the coupling level between the resonators. The dimensions of this structure are the same as those used in Fig. 3.12, and the only difference is the five additional capacitors on the side walls.

If we set all the varactor diodes of the lateral posts to 1 pF (the lower capacitance value that it is considered for this test), the scattering parameters depicted in Fig. 3.18 are obtained. The two main resonances are the one located at lower frequencies, being the lowest one very attenuated because its close to the cutoff of the structure. This response

suggest that the resonators, produced by the lateral posts, are strongly coupled between them. At the same time one of their spurious resonances is placed close to the main resonances. This fact is not desirable for designing tunable filters, since the spurious resonances are quite close to the main resonances, and thus, it will not allow to shift the filter bandpass up in frequency. In addition, the coupling level is considerably high. Consequently, to adjust the coupling level to small values, typically used for channel filters, the separation between resonators should take a significantly high value.

Figure 3.18: Scattering parameter magnitude obtained for the structure depicted in Fig. 3.17. All the varactors are set to the same capacitance value. This response shows that a strong coupling between the lateral post resonators, introduces a spurious resonance near the main resonators.

3.2.4 External Q tunable mechanisms

Apart from the internal couplings, it is important to measure how the input/output couplings can be changed by using a similar strategy as that the one followed for the internal couplings. Since the last structure (Fig. 3.17) shows that a high number of open ended floating posts is not suitable for this application, only two posts are used for tuning in this study (see Fig. 3.19). This setup should modify the external quality factor (Q_{ext}), if the varactor diode capacitances are modified.

As can be observed in Fig. 3.20, there exist two spurious resonances due to the coupling posts. When the spurious resonances are near to the main resonance, the Q_{ext} is increased from 120 ($C = 1$ pF) to 260 ($C = 9$ pF). This coupling variation is outstanding, taking into account that a bandpass filter width center frequency $f_c = 2.5$ GHz requires Q_{ext} variation around of 25 to double its bandwidth.

Figure 3.19: Structure with one resonator and two open ended floating posts connected through a varactor diode to the top metallic wall. The yellow surface represents the varactor diode for the resonator, while the red rectangles represent the varactor diodes used to tune the input/output couplings. The SIW and resonator parameters remains the same as those employed in Fig. 3.12. For this structure the length of the SIW section is $l_{SIW} = 49$ mm, whereas the resonator is located in the middle of the section.

Figure 3.20: External quality factor (left ordinate axis) and spurious free range (right ordinate axis) as a function of the varactor diode capacitance (the four varactor shown in Fig. 3.19 take the same capacitance value).

3.3 Tunable waveguide filters using the studied SIW resonator modes

SIW and rectangular waveguides present similar properties. Thus, it is possible to implement the resonant mode, used in the previous design (Fig. 3.10) to an equivalent mode in a rectangular waveguide. This resonant mode strongly depends on the height of the rectangular waveguide. Therefore, the rectangular waveguide height needs to be significantly reduced and a height tapering will be required [Levy, 1973]. Moreover, in contrast of the SIW, the waveguide mode is not confined in it, thus an additional shielding structure is needed to avoid RF-leakage. that can be produced. Finally, it should be remarked that the great advantage is the great range of tuning that can be achieved.

3.3.1 Designed filters

Our first design is a third order filter with center frequency at 14 GHz. The three vertical posts that act as resonators are connected through a capacitor to the top metal cover. In order to solve the RF-leakage issue, some vacuum filled boxes are placed above the resonators to avoid radiation on that areas. The structure of the filter is shown in Fig. 3.21. The standard waveguide (WR-62), which is 15.79 mm width and 7.89 mm height, has been selected for the input and output ports. However, this last height needs to be reduced to 1 mm to reach the desired center frequency.

Figure 3.21: Structure of a reduced height rectangular waveguide evanescent filter. The red surfaces are the capacitances in each resonator, the slot is 1 mm. The dimensions illustrated in the figure are $l_1 = 5.3$ mm, $l_2 = 14.5$ mm, $l_{in} = 10$ mm, $a_{eva} = 7$ mm and the circular cylinder radius is 4 mm. The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height).

Finally, three configurations are built from the baseline configuration. Table 3.2 shows characteristic of each passband and related and unloaded quality factor (Q_U) of the filter (considering silver plating but without the varactor diode losses). It should be noticed that the capacitance values need a small variation to shift the passband. As we can see. In Fig. 3.22, this filter can be used along the usable frequencies of the standard rectangular waveguide (WR-62).

Figure 3.22: Scattering parameter magnitudes corresponding to different configurations of the evanescent filter with reduced height depicted in Fig. 3.21. Each configuration has been obtained by changing the capacitance associated with the resonators.

C_1 (pF)	C_2 (pF)	f_c (GHz)	BW (MHz)	RL (dB)	Q_U
0.8403	0.8485	11.45	100	10	1212.95
0.5602	0.566	13.09	100	18	1297.23
0.3255	0.33	15.08	120	25	1401.42
0.1262	0.126	17.51	190	26	1604.86

Table 3.2: Capacitances for four different tuning stated for the filter with reduced height, represented in Fig. 3.21. The table also includes the center frequency, bandwidth, constant return loss level and the unloaded quality factor obtained for each tuning state.

In order to solve the radiation problem in a more practical way, it could be useful to use a dielectric substrate, which, in addition, allow us to place the components needed to implement the tuning mechanism. Thus, an alternative implementation could be used for a practical implementation of these filters. In Fig. 3.23, it is shown the structure of the new filter. The dielectric substrate, placed just above the resonator, is Roger 5880 with a dielectric relative permittivity of $\epsilon_r = 2.2$ and thickness 3 mm (the dielectric material losses are not being taken into account for the simulations). Although the insertion of these dielectric boxes decreases the center frequency of the filter, it could be corrected by reducing the diameter of the circular cylinder acting as resonators, as is shown in Fig. 3.23. Thus, the size of the dielectric boxes provides a degree of freedom to design our filter with posts so that they are not extremely big or small.

Figure 3.23: Structure of the evanescent filter with reduced height shielded by using dielectric boxes on the top cover. The dimensions are $a = 15.79$ mm, $b = 1$ mm, $a_{eva} = 7$ mm, $l_{in} = 10$ mm, $l_1 = 4.5$ mm, $l_2 = 13$ mm, $l_{box} = 2.2$ mm, $h_{box} = 3$ mm. The post radius is 0.5 mm, and the red surface around them represent the capacitors (the slot for the capacitances is 1 mm).

The characteristics of the passbands for several tuning states are shown in Table 3.3 and their responses in Fig. 3.24. From the table, it is clear that the capacitance values need to be smaller than before.

C_1 (pF)	C_2 (pF)	f_c (GHz)	BW (MHz)	RL (dB)	Q_U
0.27	0.2712	12	50	12	1142.36
0.1608	0.1617	14.04	60	21	1107.63
0.088	0.0885	16.05	100	25.5	1259.63
0.042	0.0419	17.7	110	25.3	1205.58

Table 3.3: Capacitances for four different tuning stated for the filter with reduced height, depicted in Fig. 3.23. The table also contains the center frequency, bandwidth, constant return loss level and the unloaded quality factor obtained for each tuning state.

3.3.2 Couplings reconfiguration

Achieving coupling reconfiguration in an evanescent filter consists in modifying the width of the evanescent section to make it more propagative, if one requires higher coupling. Since we are developing a fully electronically tunable filter in this design, posts connected through capacitors or MEMS are proposed.

Figure 3.24: Scattering parameter magnitudes of the evanescent filter with reduced height shielded by using dielectric boxes on the top cover. The structure of the filter is represented in Fig. 3.23.

Our first test is to place two posts (one on each side), connected through capacitors, to the top metal cover in a structure with two resonators with a weak input/output coupling, as can be seen in Fig. 3.25. The configuration of the structure employs vacuum boxes above the resonators to avoid the radiation problem. However, over the post used to control the couplings (in the middle of the structure) we set a radiation boundary condition (modeled in Ansys HFSS) to measure if shielding is required in that zone in order to avoid RF-leakage.

Figure 3.25: Structure composed of two resonators and two posts for coupling tuning. Each post of this structure has associated one capacitor (red surfaces). The test consists in changing the capacitance associated to the posts in the middle of the two resonators. The relevant dimensions are: $a = 15.79$ mm, $b = 1$ mm, $a_{eva} = 7$ mm, $l_{in} = 10$ mm, $l_1 = 7$ mm, $l_2 = 10$ mm, $l_{box} = 5.2$ mm, $h_{box} = 1$ mm. The post radius is $r = 2$ mm for the resonators (the capacitor slot is 1 mm), whereas the coupling posts radius is $r_i = 0.3$ mm (and the slots for capacitances associated to them is 0.1 mm). The two posts, that controls de couplings, are separated from the side-walls a distance $pos = 0.2$ mm.

This option presents a variation of the coupling coefficient k from 0.0362 to 0.0557. This difference between the highest and lowest coupling level is 0.0195. However, as it is shown in Fig. 3.26, this mechanism of tuning introduces a new resonance near the passband and its behavior is like a non-resonating resonator. Thus, the same conclusions

that were extracted from Section 3.2.3 can be applied for this kind of coupling tuning mechanism.

Figure 3.26: Transmission parameter magnitudes as a function of capacitor value for a very small input/output couplings. The structure simulated is represented in Fig. 3.25

Other solution to realize tunable couplings is to use MEMS [De Los Santos, 2002], instead of capacitors. An important restriction to be taken into account is that each MEMS has only two different states. Thus, they provide up to 2^n possible states, being n the number of MEMS used in the design. To measure the coupling variation, the structure represented in Fig. 3.27 is simulated, where the posts between resonators are now connected through MEMS. In order to measure the RF-leakage, a radiation boundary condition is set above the MEMS.

Figure 3.27: 2D structure of two coupled resonators and six posts connected through MEMS to the top metallic wall. The waveguide dimensions are $a = 15.79$ mm and $b = 1$ mm. The dimensions used for this test are: $a_{eva} = 7$ mm, $l_{eva} = 29$ mm, and $l_{post} = 7.5$ mm. The size of the posts are $r = 2$ mm (and the slot around them (the capacitor) is 1 mm), whereas the smaller posts radius is $r_i = 0.3$ mm (the slot around them is 0.1 mm, set to PEC or not depending on the corresponding MEMS state). The distance from the nearest posts center to the side-walls is $pos = 0.5$ mm and the periodicity between them is 0.9 mm (in groups of three as shown in the figure). There are vacuum boxes above the resonator posts as well as the ones represented in the 3D structure represented in Fig. 3.21 (The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height)).

As can be observed in Fig. 3.28, the coupling variation is now within the range from $k = 0.0024$ to $k = 0.0467$ (the difference between them is 0.0443). Thus, this coupling tuning mechanism produces twice the coupling variation offered by the mechanism made with varactor diodes (based on NRRs). The MEMS combinations are connected in pairs from the side to the center in these simulations. However, they allow sixty-four states that provide a fine tuning of the coupling, even though half of them are redundant. It should be noticed that the state with none of the MEMS connected present a significant RF-leakage in the lower frequency resonance. Thus, it is still recommended to shield the structure if high levels of couplings are required.

Figure 3.28: Scattering parameter magnitude as a function of the number of connected MEMS (Symmetrically). The structure simulated is represented in Fig. 3.27.

A similar result is obtained if the external Q (Q_{ext}) is measured. Fig. 3.29 shows a structure with a single resonator and several post connected through MEMS to the top metal cover. In Fig. 3.30 the scattering parameters for two states are presented, the first one with all the MEMS disconnected (the post are disconnected to the top metallic wall) and the second one with all the MEMS connected (short-circuiting the post to the top metallic wall). It is evident from the figure that the bandwidth of the resonance can be very narrow (low input/output coupling), if the MEMS are in their ON state. On the other hand, the OFF state leads to wider bandwidth resonances.

From these results, it is clear that this tuning mechanism offers a significant coupling and Q_{ext} variation. A second test is performed in order to compute coupling curves by using the structure illustrated in Fig. 3.31. This structure shows two resonators in an evanescent section, and there are two posts between them. These posts are connected or disconnected to the top metallic wall depending on the MEMS state. The position of these posts (pos) is swept along the waveguide from the lateral to the center, and the corresponding coupling is then calculated.

Figure 3.29: 2D structure of a resonator and twelve posts connected through MEMS to the top metallic wall. The waveguide dimensions are $a = 15.79$ mm and $b = 1$ mm. The dimensions used for this test are: $a_{eva} = 7$ mm and $l_{eva} = 14$ mm. The size of the post is $r = 2$ mm (and the slot around it (the capacitor) is 1 mm), whereas the smaller posts radius is $r_i = 0.3$ mm (the slot around them is 0.1 mm, set to PEC or not depending on the configuration). The distance from the nearest posts center to the side-walls is $pos = 0.5$ mm and the periodicity between them is 0.9 mm (in groups of three as shown in the figure). There are a vacuum boxes above the resonator post as well as Fig. 3.21 (The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height)).

Figure 3.30: Scattering parameter magnitudes as a function of the number of MEMS connected (Symmetrically). The structure simulated is presented in Fig. 3.29.

Figure 3.31: Structure employed to compute the coupling between two resonators by using a scattering parameters simulation. The waveguide dimensions are $a = 15.79$ mm and $b = 1$ mm. The dimensions used for this test are: $a_{eva} = 7$ mm, $l_1 = 10$ mm and $l_2 = 20$ mm. The size of the resonant posts is $r = 2$ mm (and the slot around it (the capacitor) is 1 mm), whereas the tuning posts radius is $r_i = 0.3$ mm (the slot around them is 0.1 mm, set to PEC or not depending on the configuration). The distance from the nearest posts center to the side-walls is pos , which is varied to find the needed coupling. The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height)).

In Fig. 3.32 are presented the couplings obtained for five center frequencies (13, 14, 15, 16 and 17 GHz) as a function of the position of the posts (pos), depending on the MEMS state. The two figures show that the electromagnetic coupling (k) decreases for lower frequencies as it is expected. However, there are two major differences, on one hand, while in Fig. 3.32(a) the coupling level increases as the posts position (pos) increases, in Fig. 3.32(b) the coupling level follows the opposite trend. On the other hand, the couplings levels of Fig. 3.32(b) are higher than those obtained in Fig. 3.32(a).

(a) Coupling level as a function of the position of the posts (pos) for five center frequencies (13, 14, 15, 16 and 17 GHz) when MEMS are on short-circuiting the posts to the top wall (ON state).

(b) Coupling level as a function of the position of the posts (pos) for five center frequencies (13, 14, 15, 16 and 17 GHz) when the posts are disconnected to the top metallic wall (MEMS is on OFF state).

Figure 3.32: Coupling levels as a function of the position of the posts (pos), depending on the MEMS state, for the structure shown in Fig. 3.31.

The same test is now performed for studying the Q_{ext} as a function of the post position. For this test, a structure of one resonator and four post with its respective MEMS is studied, as it is shown in Fig. 3.33. For this test, the posts position (pos) will be swept and the Q_{ext} will be computed for the two states of the MEMS (ON and OFF).

Figure 3.33: Structure employed to compute the Q_{ext} as a function of the location of the tuning posts. The waveguide dimensions are $a = 15.79$ mm and $b = 1$ mm. The dimensions used for this test are: $a_{eva} = 7$ mm and $l_{eva} = 11.6$ mm. The size of the resonant posts is $r = 2$ mm (and the slot around it (the capacitor) is 1 mm), whereas the tuning posts radius is $r_i = 0.3$ mm (the slot around them is 0.1 mm, set to PEC or not depending on the configuration). The distance from the nearest posts center to the side-walls is pos , which is varied to find the needed coupling. The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height)).

In Fig. 3.34 present the Q_{ext} for five center frequencies (13, 14, 15, 16 and 17 GHz) as a function of the post position (pos), depending on the MEMS state. The result for this study is similar to the one obtained for the coupling level as a function of the post position, in this case, the Fig. 3.34(a) follows a growing trend, while Fig. 3.34(b) follows a decreasing trend.

(a) External quality factor as a function of the position of the posts (pos) for five center frequencies (13, 14, 15, 16 and 17 GHz) when MEMS are on short-circuiting the posts to the top wall (ON state).

(b) External quality factor as a function of the position of the posts (pos) for five center frequencies (13, 14, 15, 16 and 17 GHz) when the posts are disconnected to the top metallic wall (MEMS is on OFF state).

Figure 3.34: External quality factor as a function of the position of the posts (pos), depending on the MEMS state, for the structure shown in Fig. 3.31.

To sum up, the coupling tuning mechanism based on MEMS provides a high coupling variation. However, the main issue is that depending on the position of the post, and the coupling variation can be significantly high. Despite this fact, it is an useful resource to add coupling variation in a filter, as will be shown in the next section.

3.3.3 Rectangular waveguide evanescent filter with bandwidth reconfiguration

In this section, by using the method that was studied before, a filter which keeps the bandwidth at two different center frequencies is designed. To realize the bandwidth control, two posts are symmetrically placed for each of the desired couplings (see Figs. 3.31 and 3.33). The bandwidth of the designed filter is 100 MHz at 14 and 17 GHz. Due to the evanescent section, the configuration at 14 GHz will require a higher coupling level than the one at 17 GHz. Thus, the MEMS are in ON state at 17 GHz and OFF state at 14 GHz. The filter can be synthesized by using the information provided in Figs. 3.32 and 3.34 to compute the couplings and the Q_{ext} , respectively. The structure of the filter is shown in Fig. 3.35.

Figure 3.35: Third order filter with two possible states. The waveguide dimensions are $a = 15.79$ mm and $b = 1$ mm. The dimensions of the empty boxes are 7 mm (evanescent rectangular waveguide width) $\times$ $l_{box} = 5.2$ mm (large) $\times$ $h_{box} = 1$ mm (height)). The resonant posts radius is $r = 2$ mm, the slot around them is 1 mm, where the capacitors are considered (red surfaces). The tuning posts radius is $r_i = 0.3$ mm, and the slot around them is 0.1 mm. Finally, the dimensions calculated from Figs. 3.31 and 3.33 are $l_1 = 5.8$ mm, $l_2 = 15.35$ mm, $pos_1 = 0.6$ mm and $pos_2 = 1.01$ mm.

Figs. 3.36 and 3.37 show the scattering parameters for the two states of the filter. It can be observed that the configuration at 17 GHz needs a refinement, but the design is valid as an initial point, that can be used for future optimizations.

Figure 3.36: Scattering parameter magnitude for the configuration with center frequency at 14 GHz. This is the lower center frequency configuration of the filter depicted in Fig. 3.35.

Figure 3.37: Scattering parameter magnitude for the configuration with center frequency at 17 GHz. This is the higher center frequency configuration of the filter depicted in Fig. 3.35

3.4 Tunable SIW evanescent filter prototype

This section is focused on designing a filter at Ku band in SIW technology. To implement this filter, the resonant mode that was used in previous sections is being scaled to Ku band. Two main factors are taken into account for this task. First, as was aforementioned, the resonant mode has a strong dependence on the thickness of the substrate. Thus, the thickness of the substrate employed is an important design parameter. However, on the contrary of the rectangular waveguide, it is possible to choose the relative permittivity of the dielectric substrate, which is an additional degree of freedom. Second, to make this filter realizable, the varactor diodes are modeled with rectangular surfaces, which have the same size as the physical component, and their capacitances are extracted from the specifications of the commercial varactor diode.

Therefore, this section is divided into three subsections: the first one to design a new SIW structure for Ku band and the implementation of the resonant mode with commercial varactor diodes. The second one, a first approximation of the filter without taking into account the parasitic elements of the varactor diode, and the last one, including the parasite inductance of the varactor diode.

3.4.1 Design parameters of a substrate integrated waveguide for Ku band

In order to develop a new filter at Ku band, the first step is to design an adequate SIW structure. This SIW structure will be used for our Ku band prototype alongside its microstrip to SIW transition is illustrated in Fig. 3.38. Table 3.4 contains the parameters of the SIW structure and the input microstrip transmission line, including its transition to the waveguide. The substrate, employed for this design, is ROGER 5880 with $\epsilon_r = 2.2$, and the substrate thickness is 0.127 mm (the smaller one to achieve the highest resonant frequency, however leading to a weak practical structure). This substrate has low losses, which is an important characteristic, to work at Ku band. By using these parameters, the resulting SIW fundamental mode cutoff frequency is 15 GHz. In addition, we should keep in mind that the size of the SIW is really small, whereas the varactor diodes have

to be integrated in the filter structure. Thus, using another substrate with a higher ϵ_r would result in a smaller SIW, even more difficult to integrate varactor diodes with the filter.

	length (mm)
W_{SIW}	7
W_m	0.784
W_t	1.4
L_t	3
p	0.8
d	0.4

Table 3.4: Parameters of a SIW structure and its transition from a microstrip input line. These parameters are graphically represented in Fig. 3.38.

Figure 3.38: Sketch of the substrate integrated waveguide designed for a filter at Ku band, and the microstrip input port and its transition to it. The employed substrate is ROGER 5880 with $\epsilon_r = 2.2$ and the substrate thickness is 0.127 mm.

The resonant mode, employed until now, can be used by controlled using four capacitances aligned to the $\hat{x}$ axis of the structure, as it is shown in Fig. 3.39. By using this distribution of capacitances, we can have full control of the same resonance as that utilized in the S band filter (Fig. 3.9).

Figure 3.39: SIW resonator with four rectangular surfaces (two above the top wall and two below the bottom wall) acting as capacitances (marked as red). The radius of the inner via hole is $r = 1.4$ mm, and the slot is 0.1 mm around the via hole. The capacitor size is $C_l = 1.2$ mm $\times$ $C_w = 0.3$ mm. The SIW parameters are $W_{SIW} = 7$ mm and the lengths of the SIW section is $l_{SIW} = 20$ mm.

Figs. 3.40(a) and 3.40(b) represents the resonant frequency as a function of the thickness of the substrate and the capacitance of the varactor diode, respectively. It is important to remark that the thickness of the substrate is extremely important to scale the resonator up in frequency, as can be observed. From the commercially available panel sizes, we have

chosen the 0.127 mm thickness. This panel size allows us to set the center frequency of the filter at higher frequencies than thicker substrates.

The varactor diodes utilized in the Ku-band filter are SMV1405 from Skyworks [Skyworks, 2019]. This model provides a capacitance range from 0.63 pF to 2.67 pF. Also, this capacitor has a low parasitic resistance (0.8Ω) and one of the lowest parasitic inductance (0.45 nH) than we could find in commercial varactors. Nevertheless, these parasitic elements will not be used for during the first design stage of the filter, although they will be use at later stages.

(a) Resonant frequency as a function of the thickness of the substrate of the resonator depicted in Fig. 3.39. The markers represent computed values, while the lines are interpolated values.

(b) Resonant frequency as a function of the capacitance value of the resonator depicted in Fig. 3.39. The range of analyzed capacitances corresponds to the values of the employed varactor diodes. The markers represent computed values, while the lines are interpolated values.

Figure 3.40: Ku-band SIW resonator as a function of the dielectric substrate thickness and the varactor diode capacitance.

3.4.2 Ku band substrate integrated waveguide filter

Once the substrate integrated waveguide is designed and the resonator is scaled to Ku band, we can design a filter at 15 GHz using all these data. After the design process the filter that is shown in Fig. 3.41 is obtained. This baseline filter has center frequency at 14 GHz and bandwidth $BW = 380$ MHz (2.7% of fractional bandwidth). This bandwidth is measured at constant return loss ripple level, whose value is above 20 dB in this case. As can be observed in the structure of the filter, the bandwidth can be increased if desired, because the input coupling, controlled by the distance l_1 between the microstrip transition and the first/last resonator, can be increased. Besides, the microstrip transition requires a good matching within the range that the filter center frequency will be placed. Otherwise, the degradation when the passband is downshifted in frequency will be considerable.

Three configurations of the filter are obtained by varying the capacitor value (see Fig. 3.42). Table 3.5 contains values of center frequency, bandwidth, return loss level and the capacitance value for each plotted state. Despite the bandwidth decrease, the filter provides a high tuning range, while the return losses do not vary that much. In addition, lower center frequencies can be obtained by taking higher values for the capacitances, because the commercial varactor, that we have considered, can reach up to 2.67 pF.

Figure 3.41: Substrate integrated waveguide filter at Ku band. The length between the microstrip transition and the first resonator is $l_1 = 8$ mm, whereas the distance between the first and second resonator is $l_2 = 16$ mm. On the other hand, the diameter of the resonator via holes is 1.4 mm and the around them is 0.1 mm. The SIW parameters are the ones presented in Fig. 3.38 and listed in Table 3.4.

Figure 3.42: Scattering parameter magnitudes of the SIW filter depicted in Fig. 3.41, for three different states. These states have been obtained by changing the capacitor value as shown in the legend.

C (pF)	f_c (GHz)	BW (MHz)	return losses (dB)
0.7	14	380	20
1.1	12.95	180	16
1.3	12.37	130	15

Table 3.5: Filter center frequency, bandwidth and capacitance values for each state of the filter depicted in Fig. 3.41. The four column corresponds to the return losses level used as reference for measuring the bandwidth.

3.4.3 Substrate integrated waveguide with parasitic elements for manufacturing a prototype

In order to manufacture the filter, some changes have to be introduced to the previous model. First, the parasitic elements of the varactors need to be included. At this point, we only consider the parasitic inductance in the design of the filter. Secondly, we change the commercial varactor to SMV1430, also from Skyworks [Skyworks, 2019], which has the same parasitic elements, while their capacitance range goes from 0.31 pF to 1.24 pF. This change is due to the value of the reactance produced by the parasitic inductance, which is considerable at Ku band. Thus, the total reactance of the varactor diode of this varactor is lower than the reactance of the one employed in the previous design, reaching higher resonant frequencies. On the other hand, other parameters of the SIW remains the same.

In previous simulations, the via holes were considered as metallic post. However, for manufacturing, the via holes are realized by drilling and then metallizing the walls, being impossible to insert a varactor as were done in previous sections. Thus, a small rectangular metallic plate is placed around the via hole, as it is illustrated in Fig. 3.43, allowing to insert the varactor diodes in its extremes. Its relevant geometrical parameters are shown in the drawing. The main difference with respect to Fig. 3.39, is the metallic surface that is in contact with the metallized via hole. Finally, the via holes are not filled with metal. Thus, the resonators require a little adjustment for welding the varactor diodes.

Figure 3.43: Sketch of the resonator with non-filled via holes. The dimension P_c is 1.2 mm, P_w is 0.5 mm and $R = 0.9$ mm. The size of the varactor diode is 0.9 mm $\times$ 0.5 mm.

In Figs. 3.44(a) and 3.44(b), we conduct a study on how the resonant frequency varies if the capacitance and the thickness of the substrate are swept. As was mentioned before, the resonant frequency has a high dependence on the substrate thickness, as can be observed in Fig. 3.44(a). Hence, we take the minimum value that a standard panel of ROGER 5880 can reach, that is 0.127 mm. Furthermore, The tuning range reach up to approximately 6 GHz based on Fig. 3.44(b). Thus, to achieve the highest center frequency, our filter will be designed with the lowest capacitance.

(a) Resonant frequency as a function of the thickness of the substrate. The markers represent computed values, while the lines are interpolated values.

(b) Resonant frequency as a function of the capacitance value. The range of analyzed capacitances corresponds to the values of the employed varactor diodes. The markers represent computed values, while the lines are interpolated values.

Figure 3.44: Ku-band SIW resonator as a function of the dielectric substrate thickness and the varactor diode capacitance.

At last, the baseline filter is designed at the highest desired center frequency. The filter and its relevant dimensions are illustrated in Fig. 3.45. The board has been designed to place the bias network components, that are required by the varactor diodes needed in our design. The bias network consist in a choke inductor connected to the power supply, and the other extreme connected to the metallic plate of one resonator. Therefore, it requires a power supply for each resonator.

As shown in Fig. 3.46, the filter has a baseline configuration with center frequency $f_c = 11.41$ GHz and bandwidth $BW = 180$ MHz. By changing the varactor capacitance to 0.41 pF the filter is shifted to a configuration with cutoff frequency $f_c = 10$ GHz and bandwidth $BW = 160$ MHz measured at return loss level of 13 dB. Although the bandwidth and return loss performances show a degradation, the overall behavior is better than the previous design. This a consequence of the location of the center frequency of the filter, where the SIW is more evanescent than the previous design. Thus, the coupling level is not varying as much as the configuration in which the center frequencies of the filter were placed close to the cutoff of the SIW structure. Moreover, the filter center frequency can be shifted down reaching lower values than those represented in the Fig. 3.46. It should be remarked that this filter achieves a tuning range which is significant when compared to other filters presented in scientific papers.

it is demonstrated that the filter performance is good in term of RF-leakage, since the level of radiated power is almost negligible (less than a 1.5% of the available power). This

Figure 3.45: Layout and dimensions of the designed filter, where $l_1 = 2$ mm and $l_2 = 5.6$ mm. The board dimensions are $b_l = 31.2$ mm $\times$ $b_w = 27$ mm. The board also includes the pads to place the bias network required the polarization of the varactors. The SIW parameters are depicted in Fig. 3.38 and are collected in Table 3.4.

Figure 3.46: Scattering parameter magnitudes of the filter presented in Fig. 3.45. These states are obtained by varying the capacitances from 0.31 pF to 0.41 pF.

is due to the electromagnetic fields are confined in the dielectric substrate. Despite the degradation in bandwidth that presents the last filters, some coupling tuning mechanism can be applied if needed. These coupling tuning mechanisms can be easily integrated in the SIW structure. However, as drawback, they introduce spurious resonances near the passband. In spite of this, the overall performance, considering the tuning range and degradation in bandwidth, is really wide, specially in the last filter (see Figs. 3.45 and 3.46).

3.5 Conclusions

In this chapter three types of tunable filters have been presented. The first one was a tunable lowpass filter, whose results do not show a good performance, due to the lack of tuning capability on the impedance inverters. A future research line derivated from this work, can be followed by using another kind of impedance inverter in order to achieve more options for tuning.

The second type of filter was an evanescent mode SIW filter at S band. This filter presents a 1 GHz of tuning range (from 2 GHz to 3 GHz). For this kind of filter, some coupling tuning mechanism based on NRRs were studied. They provide more than enough coupling tuning to have a constant absolute bandwidth. At last, an SIW filter at Ku band was designed by using materials that can be manufactured and commercial varactor diodes for tuning.

The third type of filter is a reduced height rectangular waveguide filter. This filter is realized by implementing the same resonant mode that is employed in SIW filters. This rectangular waveguide alternative was studied to analyze the advantages and disadvantages in contrast to the SIW implementation. The advantage is that the resonant mode provides a high tuning range (the filters studied in the corresponding section show, approximately 6 GHz of tuning range). The main issue is that these filters require to add some shielding elements to avoid RF-leakage, in contrast to its SIW implementation. Furthermore, this work presents several coupling tuning mechanisms that can be applied to these kinds of filters, similar to the one studied for the SIW implementation. The most relevant coupling tuning mechanism was based on tuning posts connected to the top wall through MEMS. By using this mechanism, a tunable waveguide filter with two states, one with center frequency 14 GHz and other at 17 GHz), were realized, while the constant absolute bandwidth is kept.

Chapter 4

Evanescent waveguide tunable filters

In this chapter, tunable evanescent rectangular waveguide filters are presented. These tunable evanescent rectangular waveguide filter are composed of resonators, that consist of elliptic cylinders which are attached to a side wall in the evanescent section of the rectangular waveguide. The generated resonance has its maximum field concentrated in the air gaps located between the elliptical cylinder and the top and bottom walls. These resonances have strong capacitive behavior that can be employed for tuning purposes. Therefore, by adding a tunable mechanism on the top and bottom walls of the rectangular waveguide, a tunable filter can be developed.

This chapter can be divided into two main blocks. The first one analyses the proposed kind of resonators with a distributed capacitance. After that, some filters that use this kind of resonators are designed. The next iteration includes some additional components to avoid the RF-leakage produced in this filter.

The second main block on the chapter explains how the proposed resonator is adapted to employ commercial components, in order to design a practical prototype filter. Moreover, I will show how the resonator can be adapted for realistic capacitors occupying a small surface (for ANSYS HFSS [Ansoft, 2019] simulations) that represents the encapsulated commercial components, such as varactor diodes. Also, a detailed study of these resonators is performed in order to retrieve some data about trade-offs between resonant frequency, and achieved tuning range. After these analyses, a practical prototype filter is finally designed.

4.1 Evanescent filter with horizontal elliptical posts

This section proposes a novel tunable evanescent resonator. It is shown how the resonance frequency is modified by using two distributed capacitances. In addition, a structure to avoid RF-leakage in this type of filter is proposed, which employs a dielectric substrate that shields the structure. Moreover, the dielectric substrate will be later used as a support to include the physical varactor diodes. Two filters are designed as examples: the first one without any dielectric substrate, and the second one with the dielectric substrate boxes above and under the resonators.

4.1.1 Study of elliptical posts as resonators

The proposed resonator for adding tunability consists of an elliptical cylinder attached to the side wall of an evanescent rectangular waveguide section (see Fig. 4.1). The elliptical cylinder is not touching the top and bottom walls, resulting in an air gap between them. Fig. 4.1 shows the electric field magnitude represented in the top wall of the waveguide structure. It is evident from the figure, that the maximum electric field is located in the areas between the top wall and the elliptical cylinder. The same behavior can also be observed between the bottom wall and the elliptical cylinder. Consequently, in order to introduce tunability, the tuning components can be introduced in these areas, where a capacitive effect is produced.

Figure 4.1: Evanescent rectangular waveguide section with a side post attached to the waveguide as resonator. The elliptical cylinder is not touching the top and bottom walls, existing an air gap in those areas. The electric field magnitude is represented at $f = 15.56$ GHz. The values of the electric field are referred to the normalized power computed in ANSYS HFSS. The dimensions of the structure are: $a = 15.79$ mm, $b = 7.89$ mm, $a_{eva} = 8.21$ mm, $l_{in} = 5$ mm, $l_{eva} = 15$ mm, $w_y = 3$ mm, $w_z = 2.4$ mm and $h = 6$ mm.

The proposed tuning mechanism consists in placing an floating metallic plate in the top and bottom walls, in the areas where the electric field reaches its maximum value. Fig. 4.2 shows the proposed structure; the isolated plates on the top and bottom walls are connected to the main structure through distributed capacitances in the marked surfaces (red areas). These distributed capacitances are modified in order to tune the resonant frequency. Like the previous chapter, the basic idea is based on employing varactor diodes as a tuning mechanism. Although in a first attempt, the isolated plates are floating, in further designs a support structure will be added to produce a more realistic model.

Analyzing the structure of Fig. 4.2 and sweeping the capacitance, the scattering parameter magnitudes presented in Fig. 4.3(a) are obtained. The resonant frequency is shifted from 15.66 GHz to 16.35 GHz, resulting in 690 MHz of tuning range (4.22% of the center frequency). It should be noticed from these results that the resonant frequency varies more with small capacitances. For instance, let us now consider a frequency shift of 420 MHz. This represents the 61% of the total variation, and occurs when the capacitance is changed between 1 pF and 3 pF, while the other 39% is obtained between 3 pF and 50 pF. Consequently, small values of capacitance are better for tuning filters at these frequencies. These results are consistent with SIW filters presented in Chapter 2. However, the produced RF-Leakage is extremely high for the small capacitance values (around 18% of

Figure 4.2: Structure of the proposed evanescent rectangular waveguide tunable resonator. The red surfaces on the top and bottom walls are distributed capacitances, being the plates, in the middle, isolated from the rest of the structure. The slot around the isolated plate is 0.5 mm. Other dimensions are the same as Fig. 4.1.

power radiated for $C = 1$ pF), according to Fig. 4.3(b). Therefore, it is important to find an additional structure that achieves two important functions, supporting for the isolated plates and varactor diodes and, at the same time, reducing the RF-Leakage.

(a) Scattering parameter magnitudes of the resonator depicted in Fig. 4.2. The different configurations have been computed by sweeping the value of the distributed capacitance.

(b) RF-Leakage for each configuration presented in Fig. 4.3(a). This RF-Leakage is computed by using the formula $RFL = (1 - |S_{11}|^2 - |S_{21}|^2) \times 100$.

Figure 4.3: Scattering parameters and RF-leakage for different configurations computed by sweeping the distributed capacitance of the tunable resonator depicted in Fig. 4.2

4.1.2 Filter design

A third order filter has been designed by using the evanescent resonators presented in the previous section. The structure of the filter and its relevant dimensions are illustrated in Fig. 4.4. The marked surfaces are the isolated plates that are connected through distributed capacitors. The distributed capacitors are small slots, as was presented in the previous section. For this filter, a shielding mechanism is not included, in order to measure the amount of power that can be radiated in filtering structure. Thus, taking into account the results from Fig. 4.3(b), a high RF-leakage level is expected in configurations with low capacitance value.

In order to obtain initial dimensions for the filter, it is recommended to carry out the first design without capacitors. After obtaining those values, the filter can be finally synthesized with high values of capacitance, such as 50 pF or 100 pF, which are high enough to represent a "short-circuit" state at these frequencies. This design procedure is recommended in order to obtain the initial dimensions, without the finer mesh required due to the vacuum volume and the radiation boundary condition, needed for the open structures. When placing the capacitors, a significantly dense mesh is needed to run simulations of the whole filter geometry.

Figure 4.4: Structure of a third order filter realized with floating plates (marked surfaces) connected through capacitors. The dimensions are: $a = 15.79$ mm, $b = 7.89$ mm, $a_{eva} = 8.21$ mm, $l_{in} = 5$ mm, $l_{eva} = 15$ mm, $l_{res} = 9$ mm, $l_1 = 12$ mm and $l_2 = 28.12$ mm. The slot around the isolated plates is 0.5 mm. The parameters for each post resonator are given following the same nomenclature of Fig. 4.1. The parameters, common for all post in the filter, are $h = 5$ mm and the semiaxis $w_y = 3.5$ mm. On the other hand, the semiaxis w_z differs, being $w_{z1} = w_{z3} = 6.55$ mm for the first and third resonators and $w_{z2} = 6.17$ mm for the second one.

The scattering parameters, obtained for this filter, are represented in Fig. 4.5. The baseline configuration center frequency is $f_c = 14.05$ GHz and bandwidth $BW = 110$ MHz. The configuration, leading to the highest frequency, leads to $f_c = 15.03$ GHz, achieving almost 1 GHz of center frequency tuning. As can be observed, the scattering parameters of some states of the filter present high insertion losses. As the filter structure was defined without any kind of losses (all the walls are set to perfect electric conductor), all the losses are produced by RF-leakage. The RF-leakage is specially relevant for the tuning state of center frequency ($f_c = 15.1$ GHz), where the $|S_{21}|$ is 5.6 dB at 15.1 GHz. Thus, it is evident that this filter needs a shielding structure to minimize this problem.

4.1.3 Filter shielding to avoid RF-Leakage

A first attempt to solve RF-Leakage problems is to employ dielectric boxes placed surrounding the areas where the capacitors are located. This solution is applied for the previously designed filter, resulting in the structure represented in Fig. 4.6. The filter dimensions are the same as those presented before in Figs. 4.4 and 4.1. Nevertheless, a second refinement of the proposed configuration will be needed in order to introduce the bias voltage for the varactors.

Figure 4.5: Scattering parameter magnitudes of the filter, illustrated in Fig. 4.4, realized with floating plates connected through capacitors. The center frequency is shifted by changing the capacitances for each resonator. The filter also shows high radiation levels when the center frequency is shifted to higher values.

Figure 4.6: Filter with dielectric shielding boxes. The dimensions of this filter are the same that the previous one presented in Fig. 4.4. The only difference is the dielectric boxes located above and under the waveguide where tuning mechanism are placed. The width of the dielectric boxes is the same as that corresponding to the evanescent section, the length is $l_{box} = 12$ mm and height $h_{box} = 2$ mm, and their relative permittivity is $\epsilon_r = 10.2$.

The scattering parameter magnitudes from Fig. 4.7 reveal that the filter does not present RF-Leakage, and the highest frequency state reaches similar values as in the previous design. This leads to 1.3 GHz of tuning range. In addition, the dielectric boxes also introduce resonances around 13.5 GHz, which can be problematic if the resonance reaches the filter bandpass frequencies. However, this is not a major problem, since the resonance is related to the dielectric box dimensions. Thus, the spurious can be moved to proper locations, if the boxes dimensions are chosen carefully.

Figure 4.7: Scattering parameter magnitudes of the filter with dielectric shielding boxes, depicted in Fig. 4.6 for four different tuning states.

4.2 Practical prototype filter

In this section, it is studied the adaptation of the previous resonator with a distributed capacitance to a structure with small surfaces that model realistic varactor diodes. This last resonator is realized to be feasible for practical implementation with commercial components. Thus, the commercial availability of dielectric substrates and varactor diodes is taken into account. After that, a full characterization of this resonator is carried out. At last, a filter is designed using this data to maximize the tuning range.

4.2.1 Prototype resonator analysis

The previous filter needs to be adjusted to obtain a feasible structure for practical implementation. Also, the distributed capacitances must be replaced by more realistic surfaces that have the size of the physical varactor diode. Therefore, the first step is to find a configuration of varactor diodes that has the same behavior as the previously used distributed capacitance.

An equivalent resonator with varactors can be obtained if the distributed capacitances are divided into four rectangular surfaces for each floating plate (see Fig. 4.8). The test consists in sweeping the width of all rectangles from 9.8 mm (the maximum for this structure) to 0.5 mm, in order to study how the resonance varies.

The scattering parameter magnitudes, obtained from sweeping the width of the capacitance surfaces, are presented in Fig. 4.9. These results show how the resonance is shifted

Figure 4.8: Evanescent resonator with discrete capacitances. The red rectangles represent capacitances whose width will be swept. The waveguide dimensions are $a = 15.79$ mm, $b = 7.89$ mm. The evanescent section length is $l_{eva} = 20$ mm, whereas the input length is $l_{in} = 15$ mm. The elliptical cylinder size is: $w_z = 6.5$ mm, $w_y = 3.5$ mm and $h = 5$ mm, following the scheme presented in Fig. 4.1.

down if the capacitor width is reduced. In addition, the RF-leakage is noticeably increased as the size is reduced. Consequently, it is important to find an appropriate shielding structure (based on the studied mechanism of Section 4.1.3) for this kind of resonator, since the width of the varactor is 0.5 mm (SMV1430 from Skyworks [Skyworks, 2019]).

Figure 4.9: Transmission parameters of the evanescent resonator with discrete capacitances depicted in Fig. 4.8 for six different widths of the capacitors.

To introduce the varactor diodes and shielding to the structure at the same time, our solution is to place two SIW structures above and under the resonator, that act as top and bottom walls, as can be observed in Fig. 4.10. These SIW structures have an isolated plate connected through varactor diodes to the rest of the structure. Moreover, this isolated plate contains an attached via hole for feeding the varactor diodes, as detailed in Fig. 4.11. The chosen substrate is ROGER 5880, due to its low dielectric relative permittivity ($\epsilon_r = 2.2$) and its low losses ($\tan \delta = 0.0009$). The substrate thickness for this filter is 3.17 mm (h_{subs} in Fig. 4.10). A high relative permittivity dielectric material

should not be used in this case, since it could lead to the excitation of many higher order resonances.

Figure 4.10: Resonator with discrete capacitances an SIW structure acting as top and bottom walls. The waveguide dimensions are: $a = 15.79$ mm, $b = 7.89$ mm, $a_{eva} = 10.3$ mm, $l_{eva} = 32$ mm and $l_{in} = 12$ mm. The elliptical post height is $h = 5.5$ mm, and their semiaxes are $w_y = 3.5$ mm and $w_z = 6.6$ mm. The thickness of the dielectric substrate is $h_{subs} = 3.17$ mm.

Figure 4.11: Layout of the SIW structure located above/under the resonator. The biasing network will be included in the SIW structure when needed. The SIW structure is implemented by using posts of diameter $d = 0.5$ mm (also employed for the via hole in the center of the floating plates) and periodicity $p = 1$ mm. The capacitor dimensions are: $C_l = 0.9$ mm and $C_w = 0.5$ mm. The isolated plates sizes are: width $P_w = 8.1$ mm and large $P_l = 2.8$ mm, while the slot around them is $slot = 0.1$ mm wide (also employed for the top cover slot between the metallic plate and the via hole).

The resonator is studied to maximize the tuning range and to control the resonant frequency. The relevant dimensions are the semiaxes of the elliptical post w_y and w_z , and the depth of the elliptical post h (see Fig. 4.1). In order to retain some tuning range for correcting manufacturing errors, the test is performed for two capacitances (0.4 pF and 1.1 pF). In Fig. 4.12, it is shown how the resonant frequency is affected by the depth of the elliptical cylinder. Moreover, the tuning range that is achieved by changing this parameter is plotted in the right ordinate axis. It is evident that the depth of the post (h) strongly affects the resonant frequency, whose value is modified from 14.3 GHz to 11 GHz (for $C = 0.4$ pF, which is our baseline configuration). Clearly, this represents

a significant change in the resonant frequency. Thus, this dimension would be the main one to adjust the center frequency of the filter. Besides, it is important to notice that this dimension involves a considerable increase or decrease in the tuning range, obtained by this resonator, as can be observed in the right ordinate axis of Fig. 4.12 (green line). Thus, this parameter leads to a trade-off between center frequency and tuning range.

Figure 4.12: Resonant frequency as a function of the depth of the elliptical post (h). The figure shows two curves of resonant frequency (left axis) that correspond to the limits that were imposed to the varactor. The third line corresponds to the tuning range (right axis); this data is obtained by subtracting the resonant frequency of the $C = 1.1$ pF state to the $C = 0.4$ pF. The markers represent computed values, while the lines are interpolated values.

Once the effect of height h is analyzed, the same test is performed on the semiaxes of the elliptical post. The results are plotted in Figs. 4.13(a) and 4.13(b). Fig. 4.13(a) shows that the tuning range is increased when w_y takes higher values. This result is expected, since the capacitive effect is increased when the air gaps are smaller. Thus, it has a clear impact on the obtained tuning range. On the other hand, w_z reduces the tuning range if it is increased, as it is clear from Fig. 4.13(b) (green line). Moreover, the resonant frequency exhibits low sensitivity when w_z increases from 3 mm to 7 mm.

4.2.2 Filter prototype

By using the data of the previous tests, now we can design a filter prototype. The filter structure is displayed in Fig. 4.14. For this filter, the center frequency is 12.67 GHz and the bandwidth 120 MHz. The distance between resonators (l_2 in Fig. 4.14) controls their coupling level, as is the standard procedure used to couple evanescent resonators [Gonzalez-Iglesias et al., 2012]. Also, the input/output coupling is controlled by adjusting the distance from the first or last resonators (l_1 in Fig. 4.14) to the inductive step at the input/output waveguide, respectively. For designing this filter, the resonator parameters have been chosen as follow. The first adjusted parameter is the depth of the post (h), which is the one that affects the most to the resonant frequency. Therefore, taking $h = 5.5$ mm leads to a trade-off between tuning range and center frequency (see Fig. 4.12). After that, w_y is chosen to maximize the tuning range, so it is adjusted to 3.5 mm (as indicated in Fig. 4.13(a)). Finally, w_z is chosen to match the center frequency of the filter.

For this filter, four different states have been computed by changing the bias voltages

(a) Resonant frequency as a function of the semi-axis w_y (see Fig. 4.10). In addition, the achieved tuning range is represented in the right ordinate axis. The markers represent computed values, while the lines are interpolated values.

(b) Resonant frequency as a function of the semi-axis w_z (see Fig. 4.10). In addition, the tuning range achieved is represented in the right ordinate axis. The markers represent computed values, while the lines are interpolated values.

Figure 4.13: Resonant frequency of the resonator depicted in Fig 4.10 as a function of the semiaxes w_y and w_z .

Figure 4.14: Structure of the electronically tunable third order evanescent rectangular waveguide filter. The waveguide dimensions are: $a = 15.79\text{mm}$, and $b = 7.89\text{mm}$. The lengths are: $l_1 = 8$ mm, $l_2 = 20.5$ mm, $l_{res} = 4$ mm and $a_{SIW} = 12$ mm. The via holes are hollow, and their diameters are $d = 0.5$ mm. The distance between the posts that constitute the SIW walls is $p = 1$ mm. A small metal patch ($1\text{ mm} \times 0.5\text{ mm}$) is placed on the outer surface of the SIW structure, in each resonator, for welding the biasing components. The dimensions of the elliptical post (following the nomenclature of Fig. 4.1) are $w_z = 6.6$ mm and $w_y = 3.5$ mm.

of the varactors (it changes the capacitance in simulation), as shown in Fig. 4.15. The initial design is realized for $C = 0.4$ pF, and from this baseline, the center frequency of the filter is downshifted. The center frequency for the baseline filter is $f_c = 12.67$ GHz, while the bandwidth is $BW = 120$ MHz (measured at -20 dB). However, the tuning causes filter responses that are not longer equiripple.. This effect is produced due to the different coupling level variations between inner couplings and input/output couplings, being the lower center frequency states more similar to a Butterworth filtering response. For the configuration with $C = 1.1$ pF, the filter has center frequency $f_c = 11.895$ GHz with bandwidth $BW = 30$ MHz (again, measured at -20 dB). This is expected, since all tunable filters have a variation in their bandwidth when the filter passband is shifted in frequency.

Figure 4.15: Scattering parameter magnitudes of four configurations of the filter. These configurations are computed by only changing the capacitance associated with each varactor diode. It should be remembered that the filter is continuously tunable; therefore, other configurations with center frequencies in the range shown in the figure can be achieved.

In Fig. 4.16, the scattering parameter magnitudes of the two extreme states are plotted alongside with their spurious resonances. The spurious resonances for this filter are placed near 13.35 GHz. These spurious resonances are due to interactions between posts, since they do not appear in simulations with only one resonator. In addition, it should be pointed out that the spurious bands are moved less than the filter passband. Thus, the SFR is increased as the center frequency of the filter is shifted down.

Figure 4.16: Two extreme configurations of the filter with $C = 0.4$ pF and $C = 1.1$ pF alongside with their out of band responses. There exist two spurious bands near 13.3 GHz that probably are generated by the interaction between posts or high order modes excited in the SIW structure.

4.3 Conclusions

In this chapter, a novel tunable evanescent waveguide filter has been designed and analyzed. This solution is able to reallocate the center frequency of the filter. A noticeable tuning range is observed from the resonators utilized in this filter, reaching 800 MHz of center frequency shift at Ku-band (6.5% with respect 12.25 GHz, which is in the middle of the range). This filter has been studied from the most basic structure, whose main problem was the RF-leakage, until proposed solutions to avoid this problem. The solution is based on using a dielectric substrate acting as top and bottom covers of the base waveguide structure. These dielectric substrates avoid the radiation losses and allow to introduce easily the electronic components required for tuning. Also, this substrate can be employed to place the biasing network on the same board. Further work must be focused on avoiding spurious passbands and enhancing the tuning range of the filter.

Chapter 5

Tunable artificial magnetic conductors applied to rectangular waveguide filters

It is well known that a perfect magnetic conductor PMC [Samani and Safian, 2010, Rajagopal et al., 2017, Foroozesh and Shafai, 2009, Hosseini and Hakkak, 2008] surface reflects the waves with a phase of 0° instead of 180° , corresponding to perfect electric conductors (PEC). However, perfect magnetic conductors do not exist in nature. In order to emulate the 0° reflection phase, artificial magnetic conductors (AMC) are used. In practice, the AMCs are realized by using periodic surfaces. Moreover, their reflection coefficient phase only crosses the zero degree level at only one frequency point [de Cos et al., 2010, Hosseinipanah et al., 2007] (because of the resonant nature of an AMC). The useful bandwidth of an AMC is usually considered to be within the range of $+90^\circ$ to -90° .

In this chapter, we explore AMCs to transform some of the walls of rectangular waveguides from PEC condition to PMC condition or vice versa. This condition change can be performed by varying the capacitance of a varactor diode that is integrated in the AMC. In rectangular waveguide, modifying side walls (one or both) with AMCs, while keeping the upper and lower walls as PEC, can be used for tuning purposes. If, for instance, one side wall is replaced by an AMC, then the rectangular waveguide will be virtually wider, in a frequency range. If we apply this concept to a filter composed of cavities where the TE_{102} mode is taken as the resonant mode, the passband will be shifted to lower frequencies, due to the change in the width of the equivalent rectangular waveguide.

In this chapter, it is analyzed the performance in filters composed of rectangular cavities, when one or two side walls are replaced by AMCs. The chapter will provide the basics concepts of how to design an AMC in ANSYS HFSS [Ansoft, 2019], and the parameters needed for designing AMCs specifically for rectangular waveguides. After that, the designs with rectangular waveguide filters will be studied to establish the different issues that occur and their possible solutions.

5.1 Design and simulation of an artificial magnetic conductor (AMC)

In order to design an artificial magnetic conductor for our rectangular waveguide filter, the first step is to decide the layout of our periodic surface. Each period of this structure is known in the technical literature as *unit cell*. For the sake of simplicity and without loss of generality, the unit cell layout will consist of two capacitive pads joined with a varactor diode (a variable capacitance for simulation purposes), as is illustrated in Fig. 5.1.

As an example to show the design process of practical AMCs, a unit cell for a filter with center frequency at 14.5 GHz and bandwidth 100 MHz is analyzed. This filter will be implemented in a rectangular waveguide with width $a = 15.799$ mm and height $b = 7.899$ mm (WR-62). The proposed layout, shown in Fig. 5.1(a), is customized to a periodicity of $p = 4$ mm. This periodicity is chosen to be equal or less than $\lambda_g/4$ of the rectangular waveguide fundamental mode. There exist two design parameters, namely the height H and width W of the strip. These parameters control the frequency at the reflection coefficient phase crosses the zero degree level, and they will be adjusted using simulations. Fig. 5.1(b) shows how the periodic surface is placed replacing a lateral wall in a waveguide. This view displays other two design parameters, the substrate dielectric permittivity ϵ_r and the thickness of the substrate th , which also adjust the frequency at the PMC is located (zero degree in reflection coefficient phase).

Figure 5.1: a) Layout of the proposed unit cell for a periodic surface. This cell is designed to be used in a rectangular waveguide with the following dimensions: width $a = 15.799$ mm and height $b = 7.899$ mm (WR-62). The dimensions of the surface representing the capacitors are $C_l = 1.2$ mm $\times$ $C_w = 0.3$ mm. b) Front view of the waveguide and the periodic surface acting as side wall.

The proposed unit cell is considered to be part of an infinite periodic surface, and this infinite periodic surface is the one that achieves a PMC at a specific frequency. Therefore, to run simulations of our unit cell as an infinite periodic structure, some boundaries have to be defined in their lateral surfaces. In Fig. 5.2 is displayed the two configurations that ANSYS HFSS [Ansoft, 2019] provides to run this kind of infinite structure simulations. The first one (Fig. 5.2(a)) is utilizing the PEC and PMC boundaries; this configuration only can be employed for a normal incidence wave to the unit cell. The second one (Fig. 5.2(b)) is using Master and Slave boundaries, which allow considering other incidence angles.

Figure 5.2: Simulation schemes for an ANSYS HFSS simulation. The boundaries, given in the figure, are necessary to consider the unit cell as a part of an infinite array of them (periodic surface). a) Configuration for a normal incident wave using PEC and PMC boundaries. b) Configuration with master/slave boundaries, it allows considering different incidence angles. For a normal incidence, the result is the same.

In Fig. 5.3, the simulation structure in HFSS is shown. The setup includes the dielectric substrate marked in yellow, the two strips and the capacitor. The boundaries defined for this simulations are PEC and PMC, as was shown in Fig. 5.2(a). It is important to remark that there is no way to run simulations that only contain the unit cell to accurately measure its reflection phase. Thus, a vacuum volume is placed to separate the waveport¹ from the unit cell. Although the distance between the port and the unit cell (l in Fig. 5.3) is not important, it should not be too short to retrieve an accurate value of the reflection phase. After the simulation is finished, the phase due to the vacuum volume must be discounted by a postprocessing of the data².

Figure 5.3: ANSYS HFSS model for simulation of periodic structures. The periodic structure is composed of two strips connected to a capacitor printed on a dielectric substrate (yellow volume in the picture) of $\epsilon_r = 2.2$ and thickness $th = 1$ mm. The distance between the waveport and the unit cell is $l = 20$ mm. The parameters of the metallic strips are defined in Fig. 5.1, where $h = 2.5$ mm, $w = 1$ mm and $p = 4$ mm. The dimensions of the surface representing the capacitors are $C_l = 1.2$ mm $\times$ $C_w = 0.3$ mm.

¹Waveports are one of the ANSYS HFSS excitations. ANSYS HFSS runs simulations considering that the waveport surface is connected to an infinite structure, which is excited by the modes in the cross-section. Finally, ANSYS HFSS uses those fields associated with the port as a boundary for the 3D problem.

²This postprocessing can be performed directly in ANSYS HFSS in the waveport postprocessing label

A parametric sweep is conducted with three parameters, W , H and the capacitance, whose values take 0.1 pF, 1 pF, and 50 pF. The criteria to optimize the structure is to find the combination of W and H (using the highest capacitance) that places the zero phase condition in the reflection coefficient close to 14.5 GHz (our filter center frequency). Following these criteria, the chosen values are $W = 1$ mm and $H = 2.5$ mm. The phase of the unit cell is plotted in Fig. 5.4 for three states. The 1 pF is the configuration for zero phase condition at 14.5 GHz, 0.1 pF is another state that we will work in the filter and also 50 pF. It should be mentioned that, from all the possible AMCs that can be obtained, the best one is the configuration that achieves the highest tuning range between the 1 pF state and the 0.1 pF state. The last configuration, with $C = 50$ pF, is employed to show how, at these frequencies, the high capacitances are not suitable and only moves 500 MHz the zero degrees reflection phase with respect to the 1 pF state.

In a waveguide structure, the AMC will virtually increase the rectangular waveguide width, and then the resonances will be shifted down in frequency. Therefore, the procedure is based on setting the zero reflection phase condition (PMC) at the center frequency of the highest frequency configuration of the filter. For developing the filter, the highest frequency configuration of the filter is designed with the capacitances of the unit cells set to 0.1 pF. Other states of the filter will be achieved by using capacitances in the range of 0.1 pF to 1 pF.

Figure 5.4: Reflection coefficient phase of a unit cell of the periodic structure for three states 0.1 pF, 1 pF and 50 pF. These results corresponds to the unit cell presented in Fig. 5.1 and using the setup displayed in Fig. 5.3.

5.2 Proof of concept test

A proof of concept test is carried out demonstrate our concept on how the PMC can virtually increase the width of a rectangular waveguide. The test consists in an evanescent rectangular waveguide where one of its side walls is replaced by the periodic surface, as presented in Fig. 5.5. This periodic surface is composed of an array of unit cells that were previously studied and designed (see Fig. 5.3). The evanescent waveguide section has a cutoff frequency of 15 GHz without the periodic surface.

Figure 5.5: Evanescent rectangular waveguide structure with an artificial magnetic conductor acting as a side wall. The strips of the periodic structure and the capacitor are marked in black. The width of the evanescent rectangular waveguide is 10 mm, while the input/output port has standard size ($a = 15.799$ mm and $b = 7.899$ mm, WR-62). The evanescent waveguide section width is $a_{eva} = 10$ mm and its length is $l_{eva} = 48$ mm. The periodic surface is composed of 12 unit cells and its parameters are presented in Fig. 5.3.

The scattering parameter magnitudes obtained for two configurations of the periodic surface, with $C_{AMC} = 0.1$ pF and $C_{AMC} = 1$ pF, are plotted in Fig. 5.6. These responses do not show a clear transmission due to the width step introduced in the input/output ports. As can be observed, the introduction of the periodic surface in the structure changes the cutoff frequency (15 GHz) from the evanescent section, resulting in a practical cutoff frequency of 13.1 GHz in the configuration with capacitances $C_{AMC} = 0.1$ pF, which has a PEC response at these frequencies, based on the reflection phase coefficient of Fig. 5.4. However, using the other state of the periodic surface, with the capacitors set at $C_{AMC} = 1$ pF (PMC at 14.5 GHz), the cutoff of this structure is downshifted to 12.5 GHz. This fact confirms our theory because the lower cutoff frequency of the structure can only be produced by a "virtually" larger width of the evanescent section. Thus, we can conclude that the artificial magnetic conductors are suitable to virtually widen the width of a rectangular waveguide. These results also shows a bandgap that can be observed for the scattering parameters of the configuration with $C_{AMC} = 1$ near to 16 GHz. Although this bandgap is not relevant for this test, this confirms that our periodic surface is working as expected.

To confirm visually the effect of the AMC in a rectangular waveguide, two simulations are performed to evaluate the electric field at 14.5 GHz, one with all capacitors set to $C_{AMC} = 0.1$ pF and the second one with all capacitors set to $C_{AMC} = 1$ pF. The electric fields for both simulations are displayed in Fig. 5.7 and Fig. 5.8. Since the two configurations are propagative at this frequency, as was shown in Fig. 5.6, the electric fields, corresponding to the fundamental mode of the structure, propagate in both cases. Nevertheless, there are some differences. In Fig. 5.7, the maximum amplitude of electric field can be located at the center of the evanescent section (wall set to PEC state with $C_{AMC} = 0.1$ pF), while in Fig. 5.8 is just on the AMC wall (wall set to PMC state with $C_{AMC} = 1$ pF). It is evident that the last configuration shows an electric field distribution corresponding to a rectangular waveguide twice the width of the original one, because the maximum field is located in the side wall of the periodic surface.

Figure 5.6: Scattering parameter magnitudes of the evanescent rectangular waveguide with an artificial magnetic conductor in one of its sides represented in Fig. 5.5. The dimensions of the employed unit cell are given in Fig. 5.3. There are two states of the AMC; the first one ($C_{AMC} = 0.1$ pF) is that synthesizing a PEC boundary condition in the represented frequencies, whereas second one ($C_{AMC} = 1$ pF) models a PMC boundary condition between 13.5 GHz and 15.3 GHz.

Figure 5.7: Electric field in a evanescent rectangular waveguide with an AMC acting as lateral wall represented in Fig. 5.5. These fields are evaluated at 14.5 GHz with the configuration determined by $C_{AMC} = 0.1$ pF. The fields corresponding to the fundamental mode at this particular frequency are propagative and at this frequency the artificial surfaces behaves close to PEC.

Figure 5.8: Electric field in a evanescent rectangular waveguide with an AMC acting as lateral wall represented in Fig. 5.5. These fields are evaluated at 14.5 GHz with the configuration determined by $C_{AMC} = 1$ pF. The fundamental mode fields corresponding to this configuration are propagative at this particular frequency, being the maximum amplitude located on the side wall, where the AMC is placed.

5.3 Rectangular waveguide filter with one AMC acting as side wall

At this point, an inline rectangular waveguide filter is realized in order to demonstrate how to tune a bandpass filter using tunable AMCs. The filter is designed in standard rectangular waveguide WR-62 ($a = 15.79$ mm and $b = 7.89$ mm). The baseline state center frequency is $f_c = 14.5$ GHz and its bandwidth is $BW = 100$ MHz. The AMC employed for this filter is the one employed in Fig. 5.1, whose parameters are: $h = 2.5$ mm, $w = 1$ mm and its periodicity is $p = 4$ mm. The employed substrate thickness is 1 mm, and its relative permittivity is $\epsilon_r = 2.2$ mm. Once the artificial magnetic conductor is modeled, a rectangular waveguide filter could be designed with the desired specifications. The selected resonant mode is chosen TE_{102} , which together with the periodicity of the unit cell means that there are, at least, four unit cells for each resonator. In addition, we want to emphasize that having a relatively high number of cells in each provides a better control of the tunable filter than that obtained for a low number of cells. There is not any problem to employ, for instance, the TE_{101} resonating mode. However, this is not an optimum option, since the low number of unit cells is insufficient to accurately control the center frequency by the tuning techniques that will be discussed later.

The baseline state is designed with all capacitors set to 0.1 pF, which is our closest equivalent to a PEC wall at the frequencies where the filter passbands are located. Therefore, lower center frequency states are obtained by increasing the capacitances (resulting in a virtually wider rectangular waveguide and, consequently a wider rectangular waveguide cavity). Designing the filter by starting from the lower center frequency state (capacitors set to 1 pF) is not a good option, since it is not possible to know exactly where will be located the highest possible center frequency. This fact could lead to problems with the spurious passband in the configuration achieved with all capacitors set to 0.1 pF.

In order to design the filter, there exist two different approaches: the first option is to synthesize a rectangular waveguide filter (as usual with a PEC boundary in all walls), and later we add the periodic structure to a side wall and reoptimize. The second option is to design the filter with the periodic structure, from the beginning. However, taking into account the simulation time of a filter with the periodic structure, this second option is not recommended. Thus, for this PhD. dissertation, the first option will always be employed.

The first design is realized replacing only one lateral wall with a periodic surface, as shown in Fig. 5.9. The yellow volume is the dielectric material, where the strips and the capacitors are placed. This filter uses elliptical posts as irises to control the coupling level. Fig. 5.10 displays the dimensions of the synthesized filter.

Figure 5.9: Structure of a rectangular waveguide filter with an AMC acting as side wall. The dielectric substrate, which is part of the AMC, is marked in yellow being the strips and the capacitors in the inner face. The dimensions of the filter are detailed in the 2D sketch of Fig. 5.10.

Figure 5.10: 2D structure of the filter of Fig. 5.9. The lengths are $l_1 = 15.6$ mm, $l_2 = 25.67$ mm and $l_3 = 26.63$ mm. The minor semi-axis is set to $r_z = 1$ mm, while the major semi-axes, which control the coupling level, are $r_{x1} = 1.45$ mm, $r_{x2} = 3.92$ mm, and $r_{x3} = 4.24$ mm. The parameters of the periodic surface are given in Fig. 5.3.

In Fig. 5.11, two sets of scattering parameter magnitudes are presented. These correspond to set all capacitors either to 0.1 pF or 1 pF, which are our minimum and maximum possible capacitances. The baseline filter has center frequency at $f_c = 14.5$ GHz and bandwidth $BW = 100$ MHz. The state for 1 pF indicates that not all the resonators suffer from the same frequency shift in their resonant frequency. Thus, the filter response for 1 pF is not well tuned. Also, the couplings seem to be modified, due to the virtually larger waveguide width in the areas near the elliptical posts. Besides, the low frequency state shows a spurious band near 14.6 GHz. This spurious response is undesired, because it is really close to the passband of our baseline filter, being hard to remove in the microwave system. Despite the untuned passband in the low frequency state, it can be observed from Fig. 5.11, that this filter offers a noticeable tuning range from 13.75 GHz to 14.5 GHz (approximately 750 MHz), which represents a 5.17% in terms of center frequency shift.

Figure 5.11: Scattering parameter magnitudes for the two extreme configurations of the filter illustrated in Fig. 5.9. The low frequency response presents a spurious passband close to the frequencies where the baseline filter passband is located.

5.4 Capacitance tapering in filters with AMCs

The main problems of the previous structure are the coupling variation and the passband degradation after tuning the AMC. Both problems are produced by the AMC. There exists the possibility to introduce a capacitance tapering as a function of the position of the capacitors in the filter. This tapering aims to solve the problem of the coupling variation. The key idea is to maintain the PEC boundary condition unchanged, in the areas near to coupling posts, while boundary condition close to the center of the resonant cavities becomes PMC. Thus, the coupling between resonators is maintained and the center frequency of the filter shifted at the same time.

In order to analyze the effect of tapering functions, the filter presented in Section 5.3 (see Figs. 5.9 and 5.10) is modified applying tapering functions in its periodic surface. In our case, half-sine and triangular shaped functions are being compared. The half-sine and triangular functions are presented in Figs. 5.12(a) and 5.12(b), respectively. These functions show the capacitance as a function of its normalized position along the resonator. Thus, the lower capacitance values correspond to the positions where the elliptical posts are located. On the other hand, the 0.5 normalized position (Figs. 5.12(a) and 5.12(b)) value corresponds to the center of the resonator. Then, the capacitance will be maximized in the middle of the cavity. In other words, the structure behaves as PEC near the elliptical posts. These plots are particularized to the range of capacitances that is considered ($C = 0.1$ pF to 1 pF), only for illustrating the patten determined by the selected mathematical function for the amplitude distribution. In addition, the maximum value that is set in this function will determine how much the center frequency is shifted. Knowing the location of the posts and the position of the capacitors in the baseline filter, the normalized distance (z_{norm}) can be calculated as:

$$z_{norm} = \frac{z_C - z_i}{z_{i+1} - z_i}, \quad (5.1)$$

Figure 5.12: Capacitance tapering functions in terms of normalized distance.

being z_i and z_{i+1} the position of the posts numbered from the input port 1 to the output port 2 and z_C the position of the capacitors. After that, the values of the capacitors C_i (numbered from the input port to the output port by their position in the propagative direction $\hat{z}$) are obtained from:

$$C_i = C_{min} + (T_j - C_{min}) \cdot \sin(\pi \cdot z_{norm}) \quad (5.2)$$

$$C_i = C_{min} + (T_j - C_{min}) \cdot \Lambda \left(\frac{z_{norm} - 0.5}{0.5} \right), \quad (5.3)$$

being C_{min} the minimum capacitance of the varactor, and T_j is the maximum value of capacitance that the amplitude sine function will take. This T_j value will allow to control the tuning level of each resonator independently. As an example, Figs. 5.13(a) and 5.13(b) show the capacitance values for the previously presented filter (see Fig. 5.9), following a half-sine tapering $C_j = 0.2$ pF. The blue line describes the function that the capacitance follows and the red circles the value that each capacitor takes. The only difference between them is that in Fig. 5.13(b) the tapering is also applied to the input/output port sections.

Finally, these tapering functions are applied to the periodic surface of the previously presented filter (see Fig. 5.9). Due to the complexity of assigning the high number of capacitances, a MATLAB script is employed to generate the structure of the filter. Fig. 5.14(a) and 5.14(b) displays the scattering parameters for the configurations previously mentioned. As it is clear from the figures, the half-sine function offers a better tuning range than the triangular function. Therefore, the next designs will always use the half-sine function tapering. On the other hand, the tapering on the input and output ports has small effects on the performance of the filter. In fact, the results show small differences between, due to the slightly difference between the two configurations. However, for future filters this idea can be employed as a mechanism to slightly tune the input coupling. Thus, the next filters will be realized applying the tapering function to the input waveguide section.

(a) Capacitor values as a function of their position in the filter using a half-sine tapering function. The capacitors in the input and output ports are set to 0.1 pF to emulate PEC condition.

(b) Capacitor values as a function of their position in the filter using a half-sine tapering function. A tapering function have been applied for the capacitors in the input and output ports.

Figure 5.13: Capacitor values as a function of their position in the filter depicted in Fig. 5.10 using a half-sine tapering function.

(a) Scattering parameter magnitudes of the filter for two tapering configurations (with and without tapering at the ports) and the baseline for a half-sine tapering function.

(b) Scattering parameter magnitudes of the filter for two tapering configurations (with and without tapering at the ports) and the baseline for a triangular tapering function.

Figure 5.14: Scattering parameters magnitudes of the filter depicted in Fig. 5.10, for two configurations with a half-sine tapering function applied to the tunable AMC.

Applying tapering to achieve other configurations, results in the filters that are presented in Fig. 5.15. The scattering parameters of the filters show a good tuning range from 13.83 GHz to 14.5 GHz (4,62% of tuning range). In contrast, the filter without tapering presents a better tuning range (13.75 GHz in the lower frequency state). However, the passband was not properly tuned, as previously discussed. Moreover, the scattering parameters do not show any spurious passband appearing at 14.5 GHz. The T_j values, employed in this configuration, are listed in Table 5.1.

Figure 5.15: Scattering parameter magnitudes of the filter depicted in Fig. 5.9. Four tuning states are plotted: the baseline and three with half-sine tapering, with the maximum T_j (5.2) amplitudes: 0.2 pF, 0.5 pF and 1 pF.

Tapering	T_1, T_6	T_2, T_5	T_3, T_4
0.2 pF	0.2	0.187	0.2
0.5 pF	0.5	0.475	0.5
1 pF	1	0.985	1

Table 5.1: T_j values for the tapering distribution. Each variable T_j controls the tapering in a resonator or the input/output port.

To measure the spurious free range, a simulation is performed from 13.5 GHz to 17 GHz. The obtained scattering parameters are shown in Fig. 5.16. The first spurious band appears at 15 GHz, alongside with some other spurious resonances that can be associated to the AMC. It should be noticed that the scattering parameters also show a bandgap near 15.5 GHz, where all the resonances are attenuated. This behavior agrees with previous studies and other works previously published in scientific literature [Fu and Yuan, 2005, Cucini et al., 2004, Li and Wu, 2010] and it is due to the pseudo-periodicity of the employed artificial surfaces.

Figure 5.16: Scattering parameter magnitude of the configuration with maximum amplitude of the tapering 1 pF(half-sine tapering function) for the filter depicted in Fig. 5.9.

5.5 Coupling variation mechanism

In this section, a coupling variation mechanism based on varactor diodes is proposed. For this tuning mechanisms, there are two main points that have to be taken into account. First, the posts of the filter need to be altered to induce a change into the coupling level. Secondly, the maximum field should remain in the center of the rectangular waveguide to ensure a coupling variation. However, until now, only one AMC was included in tunable filters, and due to the effect of the AMC, the maximum field is displaced toward the AMC sidewall. In order to solve this problem, the only possible solution is to add a second AMC, equal to the first one, that will replace the other sidewall. It should be remarked that this second AMC will be changed synchronously to the first one. In fact, this is equivalent to virtually enlarge or shorten the width of the employed waveguide on both sides.

To implement a change of the coupling, we have disconnected the post from the top metallic wall and have placed a varactor diode between them. The structure of this filter is shown in Fig. 5.17. This filter has two AMC, to replace the two lateral walls, being the varactor positions the areas marked in red. The selected AMC is the same as that used in the first filter of Fig. 5.9. In order to design an initial response of the filter, all the varactor diodes capacitances considered for modifying the coupling were set to $C_c = 30$ pF, which would correspond to the widest bandwidth of our filter response.

For this test, a half-sine capacitance tapering is applied to the AMC, whose T_j values are set to 0.2 pF, from the function (5.2). The scattering parameters, displayed in Fig. 5.18, are obtained by changing the capacitance associated to the varactors connected to the couplings (C_c). As observed in Fig. 5.18, this leads to a tuning range between 27 MHz and 100 MHz for the bandwidth of the passband. Though the test is performed by using a big change in the capacitance of the couplings, it is evident that there is a significant variation between the 1 pF state and 2 pF states (40 MHz of bandwidth change approximately).

Figure 5.17: Fourth order filter with two artificial magnetic conductors acting as side walls. The AMC geometrical parameters are given in Fig. 5.3. The posts are connected to the top wall through varactor diodes C_c (red surfaces), instead of being short-circuited. The periodic surface unit cell is the one depicted in Fig. 5.3. The input/output port standard rectangular waveguide dimensions (WR-62) are width $a = 15.79$ mm, and height $b = 7.899$ mm. The dimensions of the filter are: $l_1 = 17.67$ mm, $l_2 = 24.68$ mm, $l_3 = 25.63$ mm, $r_z = 1$ mm, $r_{x1} = 1.95$ mm, $r_{x2} = 4.36$ mm, $r_{x3} = 4.69$ mm. The slot where the capacitor is located is $slot = 0.5$ mm.

Figure 5.18: Scattering parameters of the filter, depicted in Fig. 5.17, with a half-sine tapering configuration with T_j set to 0.2 pF. For this configuration, the bandwidth is reduced from 100 MHz to 27 MHz by changing the varactor capacitance C_c .

5.6 Relation of spurious resonances and artificial magnetic conductor parameters

In the previous section 5.3, some designs of filters including AMCs were analyzed, which present spurious resonances at frequencies where the AMC behaves as a PMC boundary condition. In this section, five AMCs have been designed without any active element (varactor diodes). These AMCs are introduced, replacing both side walls of the rectangular waveguide, instead of only one as it was done before. After that, the scattering parameters will be evaluated. The objective of this test is to find a relation between the AMCs parameters and the spurious resonances that they generate. In addition, as it was previously done, a third order filter will be implemented in order to understand the tuning capabilities of the AMC. The filter is designed to operate at center frequency $f_c = 14$ GHz without any AMC. The importance of this test is to measure how much is shifted down in frequency when a PMC boundary condition is enforced close to their center frequency.

The AMC structure is shown in Fig. 5.19, the dimensions of the strip and the substrate thickness are adjusted to set the PMC (0° phase) boundary condition at the desired frequency. Table 5.2 contains the values of the dimensions of five different AMCs. AMCs one to four have been designed to impose the PMC boundary condition at 14 GHz, while the fifth one is set to 15 GHz. In addition, the fourth AMC is designed with 4 mm periodicity instead of 3 mm to observe if this parameter affects to the spurious. The rest of configurations are designed to show how the thickness of the substrate affects to the slope of the reflection coefficient of the PMC and its relation to the spurious.

Figure 5.19: Artificial magnetic conductor structure with a metallic strip. The geometry of the strip is defined by its width (W) and its length (L). The dielectric has relative permittivity $\epsilon_r = 2.2$ and its thickness will be varied in different designs.

	Thickness (mm)	L (mm)	W (mm)	p (mm)	Slope ($^\circ$ /GHz)
AMC 1	1	6.05	1.5	3	-78,95
AMC 2	0.5	6.7	1.5	3	-180.79
AMC 3	0.127	7.15	1.5	3	-915.28
AMC 4	0.5	6.86	1.5	4	-223.64
AMC 5	0.127	6.7	1.5	3	-839,79

Table 5.2: Dimensions and phase variation rate of the AMCs, depicted in Fig. 5.19, designed to study their effects in a rectangular waveguide filter.

The reflection coefficients of all five designed AMCs are shown in Fig. 5.20. It is evident from the figure and the Table 5.2 that the thickness of the substrate is the main parameter to control the slope of the AMC reflection coefficient. It should be pointed out that this is also related with the electrical size of the dielectric substrate. Therefore, performing a small variation in the substrate thickness is a significant variation in its electrical size. Moreover, from AMC 2 and 4, it is clear that the periodicity changes the slope, making it slightly sharper when it is increased.

Figure 5.20: Reflection phase coefficient for the five configurations listed in Table 5.2.

For this test, the filter displayed in Fig. 5.21 is employed to observe how its passband is shifted depending on the five different AMCs designed. Their responses are plotted in Figs. 5.22(a) to 5.22(e). In all cases, the spurious bands are slightly shifted down in frequency with respect to their designed PMC boundary condition frequency. The scattering parameters in Fig. 5.22(a) show a wide spurious band. In contrast to other cases, this is the widest spurious band. Therefore, we can conclude that slow phase variation AMCs provide wide spurious bands. Nevertheless, the filter is operating at 11.5 GHz as can be observed in Fig. 5.22(a). Thus, this AMC is the one that shift the most to the filter. In contrast, Fig. 5.22(b) displays a narrower spurious band, which verifies the relationship between phase variation slope and spurious "bandwidth". The filter passband is placed at 12 GHz in this case. The configuration of Fig. 5.22(c), which has the AMC with highest phase variation slope, shows a reduced spurious band as compared with other AMCs. Furthermore, the filter passband is now working at 12.7 GHz. Thus, this indicates that there is a relation between the phase variation slope of the AMC and the frequency shift produced due to the AMC. In conclusion, an AMC with slower phase variation range in frequency leads to higher frequency shift of the filter passband, but the spurious band is wider.

Employing a lower periodicity like the AMC 4, does not show a clear effect on the spurious band, as can be observed in the scattering parameters of Fig. 5.22(d). In fact, the scattering parameters are really comparable to the ones shown in Fig. 5.22(b), as expected due to their similar phase variation slope value. In this particular case, the spurious band is slightly narrower, which can be predicted taking into account that the slope is sharper than the one from AMC 2. If we employ an AMC designed to operate at 15 GHz, the spurious band appears at higher frequencies, as it is shown in Fig. 5.22(e).

Figure 5.21: Third order filter structure with non-tunable AMCs acting as side walls. This filter is implemented in standard rectangular waveguide WR-62 with dimensions: width $a = 15.79$ mm and height $b = 7.899$ mm. The elliptical post dimensions are: $r_z = 1$ mm, $r_{x1} = 2$ mm and $r_{x2} = 3.5$ mm. The input/output port length is $l_1 = 34.4$ mm, while the first and second resonator separations are $l_2 = 32.9$ mm and $l_3 = 33.4$ mm, respectively.

To conclude, based on the test results, the spurious band produced in structures with AMCs and the shift generated in a filter passband are related to the electrical parameters of the employed AMC. On the one hand, the spurious band “bandwidth” is directly associated with the slope of the reflection phase coefficient of the AMC, being the spurious wider if the slope takes a low value and vice-versa. Moreover, this spurious band appears at frequencies near to the zero phase reflection coefficient of the AMC. On the other hand, the shift generated to the filter passband is larger if an AMC with a low slope variation is employed.

Therefore, this result gives a trade-off between the spurious band and the tunability range. Provided that a spurious band is unavoidable, a slow phase variation slope AMC can be desirable to achieve a higher tuning range. Even though the generated spurious will be wide, it can be placed far from the useful passband and removed using, for instance, a lowpass filter.

The test also shows that the passband of the filter is shifted down to frequencies, where the phase of the AMC wall corresponds to a PEC state. Thus, this explains that in Fig. 5.22(a) the passband of the filter is located at the lowest center frequency, while Fig. 5.22(c) the filter passband is at the highest of this test. This can be clearly observed in Fig. 5.22(e), where the filter passband is close to 13 GHz as it is the PEC state (180°) of AMC 5 in Fig. 5.20.

(a) Third order filter with AMC 1 replacing both side walls.

(b) Third order filter with AMC 2 replacing both side walls.

(c) Third order filter with AMC 3 replacing both side walls

(d) Third order filter with AMC 4 replacing both side walls.

(e) Third order filter with AMC 5 replacing both side walls.

Figure 5.22: Test of filters with different configurations of AMCs.

5.7 Filter prototypes with AMC

This section aims to design two practical filters, one at S-band and other at Ku-band, that can be physically implemented. For this task, some considerations have to be taken into account. Firstly, the design needs to be open in the external faces to be able to feed the varactor diodes. Secondly, despite of the intention to design the filter with varactor diodes, we will only consider a practical implementation with fixed capacitors, due to the high number of voltage sources that are required to tune each varactor diode independently. In addition, to have a similar result to a varactor diode, we will employ capacitances ranging from 0.3 pF to 1.2 pF that are comparable to the range of capacitance offered by the varactor diode SMV1430 from SKYWORKS (ranging from 0.31 pF to 1.24 pF). The main idea for a practical implementation is to manufacture substrates for the baseline and substrates for the shifted state. Then, the substrates must be manually replaced for measuring the prototype in the two states.

5.7.1 S-band filter prototype

The first filter is designed at S-band where it is known that the lumped elements work always properly. This filter is designed in S-band with center frequency $f_c = 3.6$ GHz and bandwidth $BW = 100$ MHz. The employed rectangular waveguide for this design is dimensioned with a width $a = 58.166$ mm and height $b = 29.083$ mm (WR-229). In order to design this filter, the first step is to design the artificial magnetic conductor (AMC) that will replace the side walls of the rectangular waveguide. The proposed unit cell layout for this filter is illustrated in Fig. 5.23. Although the layout is similar to the one employed in previous designs, the unit cell is slightly different. In this unit cell the capacitor remains in the outer wall of the waveguide. Even though the AMC is implemented with fixed capacitances, this design allows measuring the RF-leakage that can be produced if the filter would be implemented with varactor diodes, which require an external bias network.

Figure 5.23: Layout of the proposed unit cell for a practical periodic surface with $f_{PMC} = 3.7$ GHz, realized in a dielectric substrate. The thickness of the substrate is 4.75 mm, and the dielectric relative permittivity is $\epsilon_r = 2.2$. a) View of the face inside the waveguide. The periodicity of the structure is $p = 16.2$ mm, the distance between the via holes is $sep_posts = 2$ mm, while their diameter is $d = 0.6$ mm. The plates dimensions are $w = 3$ mm and $L = 11.25$ mm, and the separation between the plates is $sep = 0.2$ mm. b) View of the outer face the waveguide. The red surface is a capacitor, whose dimensions are $C_l = 1$ mm and $C_w = 0.5$ mm. The parameters shown in the figure have values $sep_b = 0.2$ mm, $L_b = 1.3$ mm and $w_b = 0.9$ mm. These plates are disconnected from the rest of the structure using a 0.1 mm wide slot.

The phase of the reflection coefficient $\angle S_{11}$ for this structure is presented in Fig. 5.24. As can be observed in the figure, this AMC is adjusted to match the zero phase of the reflection coefficient at 3.7 GHz by using a capacitance of 1.2 pF. Therefore, the spurious band generated by the AMC is at higher frequencies than the highest frequency passband configuration. Consequently, this spurious passband can be removed by, for instance, a lowpass filter. It should be pointed out that a slower phase variation is desirable in order to achieve a higher tuning range. However, it is a hard task in S-band due to the fact that the electrical length of the substrate thickness controls the slope phase variation of the AMC. And because the thickness of the substrate is limited by the commercially available standard panel sizes. Moreover, this AMC needs a slight adjustment when placed in a rectangular waveguide replacing the side walls; this adjustment can be performed by observing the spurious frequency band in a simulation of scattering parameters with an initial design of the of the filter. After that, L is increased to 12.25 mm to finally match the desired PMC boundary condition at 3.7 GHz.

Figure 5.24: Reflection coefficient phase response of the AMC designed unit cell. The PMC boundary condition is set at 3.7 GHz by using a 1.2 pF capacitance. The phase variation slope of the PMC is $-545^\circ/\text{GHz}$, measured near that frequency point.

The structure of the filter is represented in Fig. 5.25. This baseline filter is synthesized by using the AMC with all capacitances set to 0.3 pF, for which the tunable AMC behaves as a PEC boundary condition in the baseline filter passband (the zero of the phase of the reflection coefficient for this configuration of the AMC is at 4.22 GHz). These capacitances will be changed later to shift down the center frequency of the filter. As it was mentioned before, the tapering now is limited to use values in the range from 0.3 pF to 1.2 pF in 0.1 pF steps. Using this limitation, the scattering parameters of the two extreme configurations of the filter are plotted in Fig. 5.26. The center frequency of the filter is downshifted from 3.6 GHz to approximately 3.43 GHz, resulting in a 4.72% tunability range. As drawback, it is evident that the spurious frequency band, due to the AMC, is almost in the baseline configuration passband. Furthermore, it should be pointed out that the filter has almost no RF-Leakage neither in the base configuration nor in the shifted one. Thus, we can consider that the small slots (see Fig. 5.23) and the relatively small via holes compared to the size of the waveguide are enough in this case to avoid a high radiation level.

Figure 5.25: Structure of the filter in S-band. The side walls have been replaced by the previously calculated AMC (see Fig. 5.23). The standard rectangular waveguide for the input/output port is WR-229 ($a = 58.166$ mm and $b = 29.083$ mm). The dimensions are $l_1 = 20$ mm, $l_2 = 92.6$ mm, $l_3 = 98$ mm, $w_z = 4$ mm (this is common to all posts), $w_{x1} = 5$ mm, $w_{x2} = 19.7$ mm and $w_{x3} = 98$ mm.

Figure 5.26: Scattering parameters of the filter presented in Fig. 5.25. The baseline configuration is designed with all the capacitors set to 0.3 pF. On the other hand, the tapering applied to the tapering state is carried out with T_j (5.2) maximum values of 1.2 pF for the first and last cavity, and a maximum capacitance of 1.1 pF for the second and third cavity.

5.7.2 Ku band filter prototype

Following the same principles that were explicated before, another filter operating in Ku-band is designed. The filter has been designed with center frequency $f_c = 14$ GHz and bandwidth 100 MHz (0.71%). The layout of the AMC that has been used for this filter is the same as the one presented in Fig. 5.23. For this filter, the AMC is designed to fulfill a PMC boundary condition at 14.1 GHz to avoid that the spurious frequency band, produced by the AMC, would be located at the frequencies within tuning range of the filter. In Fig. 5.27, the phase of the reflection coefficient corresponding to the designed unit cell analysis is presented. This configuration is obtained, again, with the capacitor value set to 1.2 pF. For this unit cell, the phase variation slope is $-155^\circ/\text{GHz}$, which based on previous results is better to achieve higher tuning range. However, the spurious band will be wider in frequency. It should be mentioned that because of the fundamental mode guided wavelength, the considered substrate thickness (between 0.1 and 3.5 mm) will show a small impact on the desired phase variation slope in S-band, because this physical distance is translated into a small electrical length. On the other hand, the range, employed for this substrate thickness, can become considerably large for a Ku band design, where the guided wavelength is considerably shorter, comparable to the electrical length of the employed dielectric substrate. At last, the structure of the designed filter and its dimensions are presented in Fig. 5.28.

Figure 5.27: Reflection coefficient phase response of the designed AMC unit cell. The PMC boundary condition is set at 14.1 GHz by using a 1.2 pF capacitance. The phase variation slope of the PMC is $-155^\circ/\text{GHz}$, measured near that frequency point. The geometrical parameters corresponding to the plates are $w = 0.5$ mm and $L = 3.3$ mm (3.24 mm after adjusting in the waveguide side), other parameters remains the same values as those used for S-band (see Fig. 5.23).

The scattering parameters obtained for the two extreme configurations, of this filter are presented in Fig. 5.29. As in the previous designs, the baseline filter is designed without tapering, setting all the capacitors to 0.3 pF, while the tapering state is obtained applying a tapering with maximum $T_j = 1.2$ pF (5.2). As can be observed, the tapering state is not well tuned. This is produced due to the limitation of using fixed capacitances, thus the capacitance values, obtained by using the tapering function (5.2), are rounded to the first decimal digit of pF. As is evident from the figure, the RF-Leakage produces a slight miss-adjustment in the filter, leading to complex poles in the baseline configuration. Also,

Figure 5.28: Structure of the filter in Ku-band. The standard rectangular waveguide for the input/output port is WR-62 ($a = 15.79$ mm and $b = 7.89$ mm). The side walls have been replaced by the designed AMC with the response of Fig. 5.27. The dimensions are $l_1 = 21.225$ mm, $l_2 = 22.75$ mm, $l_3 = 20.025$ mm, $w_z = 2$ mm (this is common to all posts), $w_{x1} = 1.9$ mm, $w_{x2} = 6.05$ mm and $w_{x3} = 7$ mm. The thickness of the substrate is 1.575 mm, and the relative permittivity is $\epsilon_r = 2.2$.

the downshifted passband is not equalized due to the same reasons as were exposed before. In addition, the spurious frequency band, in this case, is not well formed. Thus, almost all the spurious resonances are below -20 dB. Fig. 5.29 shows how the center frequency is downshifted from 14 GHz to 13.6 GHz, providing a 2.86% of tuning range.

Figure 5.29: Scattering parameter magnitudes of the filter presented in Fig. 5.28. The tapering applied to the second state is carried out with maximum values of 1.2 pF for the first and last cavity, and a maximum capacitance of 1.1 pF for the second and third cavity.

5.8 Conclusions

In this chapter, the theory on how to design a unit cell of an artificial magnetic conductor and the procedure to design tunable filters using these AMCs has been presented. In addition, a study of the influence of the AMC when it is placed in a rectangular waveguide filter has been performed. This study concludes that there is a trade-off between the undesired spurious frequency band and the tunability range. Also, a way of tuning by

creating a capacitance tapering along the filter is proposed to maintain the bandwidth as much as possible within a certain frequency range. However, as a drawback, this implementation requires a high number of voltage sources to be able to independently tune each capacitor of the structure. Moreover, a tuning mechanism for couplings was presented, achieving a great bandwidth control. To the author knowledge, this is the first time that artificial magnetic conductors have been applied to introduce tunability in rectangular waveguide filters.

Finally, two prototypes were implemented to measure the performance of these filters. These prototypes were realized in S and Ku bands to demonstrate the tuning range that can be achieved by using commercial components. For the S-band filter, the center frequency is displaced 170 MHz (4.72%), whereas for the Ku band filter the center frequency is displaced 400 MHz (2.86%). Further work should focus on a detailed study on how to avoid the undesired spurious frequency bands. However, it should be mentioned that this is not an easy task, due to the extremely high computational cost that requires to run simulations of this kind of structures.

Chapter 6

Mechanically tunable rectangular waveguide filters

This chapter presents new tunable rectangular waveguide filters based on mechanically changing the resonators and/or couplings. These filters usually present a better performance than electronically tunable ones [Pelliccia et al., 2012, Pelliccia et al., 2015, Mansour, 2015]. This is due to the considered resonant modes. The modes, used in electronically tunable filters, have usually a low quality factor, because the electromagnetic fields are typically confined in a small cavity volumes and the losses introduced by the tuning components (varactor diodes and bias network). On the other hand, if a mechanical modification is performed to a resonator, its resonant frequency will change according to the deformation [Basavarajappa and Mansour, 2018]. This geometry modification allows to utilize high quality factor modes to implement the filter and detune it [Nam et al., 2018]. Besides, it is easy to scale the filter to upper frequencies, since mechanical actuators can always be used [Arnold et al., 2014a]. This is an important advantage with respect to electronically tunable filters which depends on the commercial varactors, exhibiting parasitic effects that are not suitable for high frequencies in most of the cases. However, mechanical actuators increase the size of the filter and its volume. This is a disadvantage when compared to the electronically tunable filters.

This chapter will be divided in three sections:

- **Tunable resonators with movable arms.** These resonators are reported to be more precisely tuned by some authors [Iskander et al., 2014]. Thus, we explore several configurations for this kind of resonators. This section aims to show that these options are feasible for realizing tunable filters.
- **Evanescent waveguide tunable filters based on ridge resonators.** These filters are easily tuned by adjusting the depth of the ridges in the evanescent waveguide section. Also, an evanescent rectangular waveguide filter with NRRs is presented, resulting in an option to maintain the absolute bandwidth in these filters.
- **Tunable filters based on movable dielectric cylinders.** The key idea is to employ a movable circular dielectric cylinder to tune the resonant frequency. For these filters, if a dielectric material with low losses is employed, a high quality factor filter can be achieved.

6.1 Tunable resonators with movable arms

Resonators with movable arms are an option to implement tunability in filters. These resonators are tuned by rotating a metallic arm or handle. Thus, the resonant frequencies are going to be related to a given rotation angle. To illustrate this concept, the most simple cavity is shown in Fig. 6.1. This cavity is realized with standard waveguide WR-62 ($a = 15.789$ mm and $b = 7.89$ mm), being its length $l = 15$ mm. In this cavity, the first resonant mode TE_{101} is excited, whereas the arm varies the resonant frequency like a tuning screw. For this configuration, it can be turned from 0° (arm is in XZ plane) to 90° (arm is in XY plane), modifying the TE_{101} resonant frequency from 13.61 GHz to 12.56 GHz, respectively. Also, the unloaded quality factor Q_U varies between configurations, ranging from 6546 to 5461, respectively; this Q_U is obtained considering silver material covering the inner part of the cavity. From the rest of the chapter, all unloaded quality factors will be calculated considering that the metal elements are silver plated in order to reduce ohmic losses.

To improve the tuning range, a highly asymmetrical cavity is one of the best options. This will maximize the angles and configurations that could be useful for producing a larger variation in resonant frequency.

Figure 6.1: Rectangular waveguide cavity with a moving arm placed in the middle. The employed standard rectangular waveguide is WR-62 ($a = 15.789$ mm and $b = 7.89$ mm), the cavity length is $l = 15$ mm. The arm length is $l_{arm} = 8.39$ mm and its height is $h_{arm} = 4$ mm, while the thickness of the arm is $s = 1$ mm.

6.1.1 Rectangular cavity with metal posts

The main drawback of the cavity presented in Section 6.1 (see Fig. 6.1) is that it keeps a high degree of symmetry. An alternative is to employ metal posts in the structure to generate asymmetry in the cavity. In Fig. 6.2, it is shown a resonant cavity with a movable arm. The rectangular prism shape of the resonator improves the effect exerted by the movable arm in the structure. Furthermore, each angle used by the arm produces a different configuration of the resonator along a complete turn of 360° .

The resonant modes of this structure and their unloaded quality factors are presented in Figs. 6.3(a) and 6.3(b). The second, third and fifth modes almost do not vary with the angle of the arm. On the contrary, the first and fourth modes present a wide range of frequency variation. This is specially true for the fourth mode that presents more than 1 GHz of variation from angle 180° to 230° , with a quality factor ranging from 6000 to 6800.

Figure 6.2: Rectangular waveguide cavity with a moving arm placed in the middle with dimensions $l_{arm} = 9.52$ mm and $h_{arm} = 7$ mm. The cavity size is $19.5 \times 19.5 \times 19.5$ mm. The diameter of the all cylinders is 2 mm, being inserted $h_{posts} 7$ mm. The posts are placed symmetrically from the center of the wall. The post from the bottom wall are separated $sep_1 = 8$ mm, while the side posts separation is $sep_2 = 6$ mm. The figure depicts the zero degrees state configuration.

(a) Resonant modes of the cavity depicted in Fig. 6.2 as a function of the angle of the movable arm.

(b) Unloaded quality factor of the cavity depicted in Fig. 6.2 as a function of the angle of the movable arm.

Figure 6.3: Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.2.

The scattering parameters of this resonator, placing the input/output ports in the front and side walls, are represented in Fig. 6.4. The mode under study shifts from 10.77 GHz to 11.85 GHz, which is equivalent to a 1.08 GHz displacement. However, in some configurations another resonance appears at 12.12 GHz. Thus, this resonator generates a spurious resonance close to the passband if the highest frequency state (11.85 GHz) is employed.

Figure 6.4: Scattering parameter magnitudes of the resonator represented in Fig. 6.2, when the ports are placed in the front and side walls.

6.1.2 Resonators combining movable arms and dielectric materials

As it is demonstrated in Sections 6.1 and 6.1.1, movable arms can produce wide tuning ranges. Another alternative, employing movable arms, is to include dielectric pieces into the cavity to generate some kind of asymmetry in the structure. In order to compute the unloaded quality factor of structures with dielectric materials, a dielectric material with relative permittivity $\epsilon_r = 24$ and quality factor $3 \cdot 10^5$ (this data correspond to a commercial dielectric material) is employed for these simulations. The proposed resonant cavity is shown in Fig. 6.5. This cavity includes two blocks of dielectric material placed on top and bottom walls.

A symmetry plane can be defined in this structure; thus, the arm only needs to move along 180° . Therefore, the results, obtained within the range from 180° to 360° , are the same to those obtained in the range from 0° to 180° , as can be observed in Figs. 6.6(a) and 6.6(b). It is important to remark that the resonant frequencies of this structure are really high in frequency. Still, it is interesting to note that the first resonating mode shifts from 18.24 GHz at 0° to 19.45 GHz at 80° , achieving a 1.21 GHz tuning range. In addition, its quality factor ranges from 5270 to 5846.

Comparing with the resonator shown in Fig. 6.2, which has exactly the same size, this resonant cavity produces resonating modes that are excited at higher frequencies, while its quality factor is not improved. Nevertheless, the idea of using dielectric materials for tuning is a feasible option to increase the tuning range associated to movable arms in resonant cavities.

Figure 6.5: Resonant cavity with two dielectric bricks placed touching the center of the top and bottom wall, together with a movable arm centered in the structure. The cavity sides are $l = 8.6$ mm. The side of the dielectric brick below the upper cavity wall is $s_2 = 2$ mm, whereas the dimensions of the dielectric brick on the bottom wall are $s_1 = 3$ mm and $h_1 = 1$ mm. The arm length is $l_{arm} = 4.32$ mm and height is $h_{arm} = 2$ mm, and the diameter of the cylinders is $d = 1$ mm.

(a) Resonant modes of the cavity depicted in Fig. 6.5 as a function of the angle of the movable arm.

(b) Unloaded quality factor of the cavity depicted in Fig. 6.5 as a function of the angle of the movable arm.

Figure 6.6: Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.5.

6.1.3 Rectangular cavities including elements that present angular variations

Instead of using movable arms, another possibility is to employ a part of the filter, which can be rotated to modify the field distribution of the cavity. For instance, the tuning element of the filter could be a movable elliptical cylinder, as shown in Fig. 6.7. This post, inside the cavity, can be rotated 90° . It can be built up either of metal or dielectric material. To illustrate the concept, the option of using a dielectric material is studied.

Figure 6.7: Resonant cavity with an elliptical cylinder in the middle. The elliptic post rotates from 0° to 90° (rotation axis $\hat{y}$). The 0° state is the one shown in this figure. The elliptical post dimensions are $r_x = 7.5$ mm, $r_z = 2$ mm and $h = 5$ mm. The dielectric material relative permittivity is $\epsilon_r = 24$ and its quality factor is $3 \cdot 10^5$. The standard waveguide width is $a = 15.79$ mm and $b = 7.89$ mm (WR-62), while the cavity length is $l = 50$ mm.

The resonant modes computed for this resonator and their unloaded quality factor are plotted in Fig. 6.8(a) and Fig. 6.8(b). This simple geometry allows to tune their resonant frequency more than 1 GHz, for instance, either with the first or the fourth mode. The quality factor of all these modes is higher than $Q = 6 \cdot 10^3$ in all cases.

(a) Resonant modes of the cavity depicted in Fig. 6.7 as a function of the rotation angle.

(b) Unloaded quality factor of the cavity depicted in Fig. 6.7 as a function of the rotation angle.

Figure 6.8: Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.7.

6.2 Evanescent rectangular waveguide tunable filter based on ridge resonators

Ridge resonators in evanescent rectangular waveguide are a good option to make tunable filters. The height of the ridges controls the resonant frequency. Thus, if a screw is used as ridge, a motor can easily change its penetration into the rectangular waveguide. Therefore, it is an interesting type of filter for tunability purposes. In Fig. 6.9, the top and side view of the filter are presented. This filter is designed as baseline at $f_c = 13$ GHz with bandwidth $BW = 100$ MHz. The rectangular waveguide standard (WR-75) has width $a = 19.05$ mm and height $b = 9.525$ mm. The resonant frequency is controlled with the post heights h_1 and h_2 .

Figure 6.9: a) Top view of the evanescent rectangular waveguide filter. The relevant dimensions are shown. b) Side view of the filter. The standard waveguide input/output ports dimensions are width $a = 19.05$ mm and $b = 9.525$ (WR-75). The dimensions are $d = 2$ mm, $l_1 = 9.8$ mm and $l_2 = 23.9$ mm, while the width of the evanescent section is $a_{eva} = 10$ mm. The height of the posts varies depending on the tuning state. The penetration of the posts for each configuration are collected in Table 6.1.

The scattering parameters of three tuning states of this filter are displayed in Fig. 6.10. In the figure, three states are shown. These states are achieved by varying only the height of the inserted posts h . It should be noted that this filter presents a continuous tuning. Therefore, any filter with center frequency between these configurations is also a possible state. Moreover, the range of tuning is not limited to the states shown in Table 6.1, where the relevant dimensions of the three different configurations are presented. The height of the posts h can vary from almost zero to the height b of the rectangular waveguide used in our design. However, the center frequency of the filter should not reach the cutoff of the evanescent waveguide section. It should be noticed that the bandwidth of the filters is reduced when their center frequency is downshifted, leading to a 20% of bandwidth reduction.

Figure 6.10: Scattering parameters for the filter presented in Fig 6.9. The filter center frequency tuning range goes from $f_c = 12.5$ GHz to 13 GHz. The lowest frequency configuration has a reduced bandwidth of 80 MHz.

	f_c (GHz)	BW (MHz)	h_1 (mm)	h_2 (mm)
State 1	13	100	3.49	3.4
State 2	12.75	90	3.62	3.54
State 3	12.5	80	3.7	3.77

Table 6.1: Specifications of the states of the filter shown in Fig. 6.9 and their corresponding ridge post heights.

6.2.1 Coupling modification

The filter depicted in Fig. 6.9 shows a good performance in terms of tuning range at the expense of a bandwidth reduction. An option is to employ NRRs to modify the couplings, at the same time that the resonant frequency of the ridge resonators is changed. To illustrate this concept, an horizontal metallic post is used as NRR in the middle of two resonators as shown in Fig. 6.11. For this test, the coupling between the two main resonators is measured when the depth (t) and diameter (d_c) of the horizontal circular cylinder are changed.

Figure 6.11: a) Top view of a structure of two ridge resonators coupled in an evanescent rectangular waveguide. Between these posts it is placed a third circular cylinder post which constitutes a third resonator, employed as NRR for controlling the coupling between the other. b) Side view of the test structure. The standard waveguide input/output ports dimensions are width $a = 19.05$ mm and $b = 9.525$ mm (WR-75). Other dimensions are $d = 2$ mm, $l_1 = 9.8$ mm and $l_2 = 23.9$ mm, while the width of the evanescent section is $a_{eva} = 10$ mm. The diameter of the third post d_c and its penetration into the waveguide section t are modified to evaluate the coupling level.

The result of this test is plotted in Fig. 6.12. The result indicates that the coupling varies due to a non-resonating resonator. Otherwise, the coupling level would have been decreased, because of the additional blockage effect introduced by the horizontal post on the direct coupling.

Even though the configuration with diameter 6 mm presents a better performance for tuning the coupling, the level of coupling decreases for depths longer than 7 mm. This effect can be explained taking into account that at 7 mm the resonance begins to scale up in frequency, as can be observed in the scattering parameters shown in Fig. 6.13. This effect could be produced due to the capacitance effect between the side post and the opposite side wall. Due to this behavior, it is not possible to reach higher coupling levels, but the change is enough to maintain the bandwidth of the baseline filter in a reasonable range of frequencies.

Figure 6.12: Coupling level between the two main resonators as a function of the penetration of the post for two different values of the radius of the horizontal post (see Fig 6.11). The computed values are marked with circles, while other values are obtained by interpolation.

Figure 6.13: Scattering parameter magnitudes of the structure depicted in Fig. 6.11 for different depths of the horizontal post, namely, $t = 7$ mm, 8 mm and 9 mm. It should be noticed the behavior of the highest resonance, produced by the NRR, emerging between 14.5 GHz and 15 GHz.

6.2.2 Tunable filter with reconfigurable bandwidth

Based on the first filter (see Fig. 6.9) and the study of the coupling provided by the horizontal post acting as NRR, a baseline filter is designed with the upper center frequency at $f_c = 13$ GHz and bandwidth $BW = 100$ MHz. The filter is depicted in Figs. 6.14 and 6.15. The drawings show the relevant dimensions of the designs. The filter has the same kind of relevant dimensions as the previous one. The only difference is the presence of the NRRs. The initial depth of these coupling-resonators is $t_1 = t_2 = 5$ mm for practical considerations. On the other hand, the radius is set to 3 mm to maximize the coupling variation. The dimensions that change for the different tuning states of the filter are listed in Table 6.2. It should be remarked that even for the presented lowest state, the bandwidth initially set for the baseline response is maintained. In addition, based on the previous studies, the filter can achieve even lower frequencies maintaining the bandwidth.

	f_c (GHz)	h_1 (mm)	h_2 (mm)	t_1 (mm)	t_2 (mm)
State 1	13	3.39	3.47	5	5
State 2	12.75	3.48	3.6	5.45	5.6
State 3	12.5	3.57	3.72	5.9	6.2

Table 6.2: Reconfigured dimensions and center frequency for each tuning state of the filter.

Figure 6.14: 3D view of the evanescent rectangular waveguide filter with non-resonating resonators.

The scattering parameters computed for the three tuning states of the filter are displayed in Fig. 6.16. These states have the same bandwidth, showing a wide spurious free range, taking into account that non-resonating resonators were used as couplings. It should be remarked again that these horizontal posts are acting as non-resonating resonators.

Figure 6.15: a) Top view of the filter with NRRs to modify the coupling. b) Side view of the filter. The circular cylindrical posts that act as non-resonating resonators are placed in the middle of the distance between the main resonators, and between the step discontinuity and the first and last resonator for the input/output coupling. The standard waveguide input/output ports dimensions are width $a = 19.05$ mm and $b = 9.525$ mm (WR-75). The dimensions of the filter for its baseline response are: $l_1 = 11.5$ mm, $l_2 = 25.3$ mm, $d = 2$ mm and $d_c = 3$ mm. The penetration of the posts into the waveguide are collected in Table 6.2.

Figure 6.16: Scattering parameters of the structure depicted in Fig. 6.14. The bandwidth is maintained for the three tuning states of the filter. Also, the resonances produced by the non-resonating resonators can be seen in the graph.

6.3 Tunable filters based on movable dielectric cylinders

For space communications systems, it is a common practice to employ filters consisting of cavities loaded with high performance (low loss and high ϵ_r) dielectric materials. There exist two main reasons for this: on one hand, the high unloaded quality factors (Q_U) that these resonators can achieve, in order to minimize the passband insertion losses. On the other hand, the electromagnetic fields are confined close to the dielectric material. Consequently, it usually leads to a significant volume reduction without producing a negative impact on the unloaded quality factor (Q_U) of the resonator. However, the main drawback of this kind of filter is the high number of spurious resonances that can appear due to the high dielectric permittivity of the employed material. Therefore, one of the main task in the design process, is to find a configuration for the resonators that enhance the spurious free range while the unloaded quality factor is maintained.

In this section, two filters using resonant modes with high unloaded quality factors are presented. To add tuning capabilities, the dielectric material of circular shape is displaced inside or outside a rectangular cavity. The design process starts by analyzing the structure in order to evaluate its resonant modes and its tuning ranges. After that, the geometry of the resonator is modified to remove other resonances near the selected mode. The first filter, implemented with a dielectric ring shaped cylinder aligned to the $\hat{x}$ axis, is designed at Ka-band, achieving tuning ranges up to 2 GHz, while the quality factor for all configurations is always above $31 \cdot 10^3$. The second filter is composed of cavities loaded with dielectric circular cylinders aligned to the $\hat{y}$ axis. The employed resonant mode achieves tuning ranges up to 4 GHz, while its quality factor ranges from $5 \cdot 10^3$ to $41 \cdot 10^3$ depending on the configuration.

6.3.1 Tunable filters based on movable lateral dielectric circular cylinders

As explained in the previous section, the proposed geometry for the dielectric resonator is presented in Fig. 6.17. The figure also represents the relevant dimensions that are employed to design the resonator. This resonator consists of a dielectric circular cylinder oriented along the $\hat{x}$ axis which has a concentric hole in it. The ring shaped cylinder can be moved into or out of the cavity in order to tune the resonant frequency. It should be mentioned that the offset (0.2 mm) of the ring cylinder is small with respect to the walls. In practical implementations, this offset needs to be filled with a spacer material (low permittivity dielectric material). Due to this separation from the walls, the quality factor achieves higher values. The depth of the ring shaped cylinder is controlled by the parameter Offset_x , shown in Fig. 6.17. The relative permittivity of the employed dielectric material is $\epsilon_r = 24$, and its typical quality factor of $3 \cdot 10^5$.

After analyzing the structure, the resonant modes chart plotted in Fig. 6.18(a) is calculated. This chart presents six modes which have a good tuning range with respect to the depth of the post inside the cavity. The quality factors, displayed in Fig. 6.18(b), present significant high values when compared with other filters presented in this chapter. We find specially interesting the third mode whose unloaded quality factor is above $31 \cdot 10^3$ in the range from 0.5 mm to 6 mm of depth.

Figure 6.17: 3D view of the resonator with a dielectric circular ring shaped cylinder. The frequency variation is achieved by modifying the dimension Offset_x . For our implementation the dimensions are: standard rectangular waveguide WR-34 ($a = 8.636$ mm and $b = 4.318$ mm), $l_{res} = 10$ mm, $d = 2.88$ mm, $d_i = 1.6$ mm and $h = 8.5$ mm.

(a) Resonant modes of the cavity represented in Fig. 6.17 as a function of the depth of the dielectric ring shaped cylinder inside the cavity.

(b) Unloaded quality factor of the cavity represented in Fig. 6.17 as a function of the depth of the dielectric ring shaped cylinder inside the cavity.

Figure 6.18: Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.17.

The electric field distribution of the fourth resonant mode is represented in Fig. 6.19. It is clear from this electric field pattern that the standard rectangular waveguide inductive iris (WR-34) can excite this resonant mode, since the resonant mode electric field have components in the $\hat{y}$ direction.

Figure 6.19: Electric field distribution of the third resonant mode for the structure of Fig. 6.17 at 19 GHz.

For this kind of resonators, it is interesting to perform a test of the shape of the iris coupling window. Even though the coupling level is going to change if the center frequency of the filter is tuned, some filters have shown that the coupling level varies less than in others. Therefore, for a movable dielectric ring cylinder, our test consists in measuring how the coupling level changes when the depth is modified in the two configurations illustrated in Fig. 6.20. In Fig. 6.20(a), the iris is attached to the same side wall as the dielectric ring cylinders, whereas in Fig. 6.20(b) the iris is centered with respect to the rectangular cavities. For the test, the input and output couplings are small to avoid affecting the test. The baselines for this test is when the offset_x is 1 mm for the two structures, being their interresonator coupling level almost the same value. This will allow for an easy comparison between them.

(a) Test structure for the iris attached to a side wall. The input/output iris width is 4.5 mm, while the second iris width is 5 mm.

(b) Test structure for the centered iris. The input/output iris width is 4 mm, while the second iris width is 4.75 mm.

Figure 6.20: Test coupling structures consisting in two resonators coupled by an iris (centered or attached to a side wall). The dimensions for the two dielectric ring shaped resonators and its cavities are the same as the one presented in Fig. 6.17.

The test leads to some interesting results, as can be observed in Fig. 6.21. In this figure, the coupling levels are obtained by varying the depth of the horizontal post into the cavity. The markers in Fig. 6.21 represent the computed values, while the lines are

interpolated ones. Each marker is associated to the center frequency at which the coupling was measured. On one hand, the lateral iris displays a lower coupling variation than the centered iris. Also, the tuning range obtained by the side iris configuration is slightly lower (220 MHz) if compared to the centered iris configuration. On the other hand, for the lateral iris, the coupling level is increased. However, for the last value, the coupling level changes its trend and starts to decrease. This effect is related to the iris width, which is 5 mm in that case. Thus, it is evident that at some point, depending on the iris width, both cases will behave in the same way. At last, we can conclude that the lateral iris is better to achieve a lower variation of the bandwidth in this kind of filter.

Figure 6.21: Coupling level as a function of the depth of the posts in the cavity (Offset_x) represented in Figs. 6.20(a) and 6.20(b). The computed values are cross and circular markers, while the interpolated values are lines. The center frequencies that have been used to evaluate the couplings, are indicated for each marker.

Finally, a baseline filter with center frequency $f_c = 19$ GHz and bandwidth $\text{BW} = 120$ MHz is synthesized in standard rectangular waveguide WR-34 ($a = 8.636$ mm and $b = 4.318$ mm). The 3D structure of the filter and its dimensions are given in Fig. 6.22. The filter was designed by using lateral irises, since they produce less variation in the couplings than centered ones, as it was studied before. It should be remarked that the input/output iris is almost completely open, giving the maximum possible input/output coupling. Hence, a filter with wider bandwidth cannot be synthesized unless another kind of enhanced input coupling is employed. Thus, this filter achieves bandwidths near to the 0.64% of their center frequency.

Fig. 6.23 displays the scattering parameters computed for the filter depicted in Fig. 6.22. Also, Table 6.3 contains center frequencies and bandwidth of each state. The filter presents a significantly high tuning range (1.5 GHz), which is 7.9% of tuning range in percent. Even though lateral irises were used, the couplings are modified if the center frequency is shifted up. Moreover, the bandwidth (measured approximately at constant return loss level) is increased as the filter passband is shifted up in frequency. However, due to the modifications in the couplings, some reflection poles become complex and the return losses level cannot be maintained. This behavior can clearly be seen in configurations with $\text{offset}_x = 4$ mm and 5 mm (see Fig. 6.23). In spite of these last configurations, the ones for $\text{offset}_x = 1$ mm, 2 mm and 3 mm show a good performance with a reasonable increased

Figure 6.22: Tunable filter built using lateral dielectric ring shaped cylinders. The employed standard rectangular waveguide is WR-34 ($a = 8.636$ mm and $b = 4.318$ mm), irises length $l_i = 1$ mm. The cavity length and the inner diameter of dielectric ring cylinders are the same as those studied before in Fig. 6.17 ($l_{res} = 10$ mm and $d_i = 1.6$ mm). The external diameter of each cylinder is adjusted to match the center frequency of the filter being $d_1 = 1.4305$ mm, $d_2 = 1.4445$ mm, $d_3 = 1.4455$ mm for the first, second and third resonator, respectively. Other resonators are keeping symmetry from input to output.

of the bandwidth. Furthermore, the filter states can be more equalized if each resonator is adjusted with a different value of $offset_x$. However, this will increase the number of actuator drives employed to reconfigure the filter.

Offset _x (mm)	f_c (GHz)	BW (MHz)
1	19	120
2	19.13	130
3	19.4	140
4	19.8	150
5	20.46	-

Table 6.3: Tuning states of the tunable filter, depicted in Fig. 6.22, by changing the depth of the lateral dielectric cylinders.

To analyze the spurious free range obtained with this filter, a test with the configurations with $offset_x = 1$ mm, 2 mm and 5 mm is performed. The computed scattering parameter magnitudes are plotted in Fig. 6.24. In addition, the figure has an to show details of the spurious band. These scattering parameter magnitudes show that the configuration with $offset_x = 1$ mm has a spurious band at 19.62 GHz, which is extremely narrow, and it is very weakly excited. This spurious appears in this analysis because a discrete sweep (provided as a simulation option in the commercial software ANSYS HFSS) was performed instead of the fast sweep (also from ANSYS HFSS) that was used in Fig. 6.23. Therefore, if this spurious is a problem, the best option is to employ the state with $offset_x = 2$ mm as baseline filter. This configuration does not show any problem related to this spurious in the considered frequency range. Still, the filter has a high tuning range (recall that the $offset_x$ can reach 8 mm if needed).

Figure 6.23: Scattering parameter magnitudes of the filter depicted in Fig. 6.22. The baseline filter is designed for $\text{offset}_x = 1$ mm.

Figure 6.24: Scattering parameter magnitudes of the filter designed with lateral dielectric ring cylinders. These scattering parameters show the frequency range from the rectangular waveguide fundamental mode cutoff frequency to 25 GHz. An inset was included to show details of the excited spurious generated spurious band. This inset has been depicted after calculating 100 frequency points equispaced by 50 KHz.

6.3.2 Tunable filters based on movable vertical dielectric circular cylinders

An alternative to the cavities used in Section 6.3.1 is to change direction of the dielectric cylinder to the y -axis. Fig. 6.25 displays the resonator under study and its parameters. The dielectric material employed for this design is the same as was previously used in section 6.3.1 with relative permittivity $\epsilon_r = 24$ and typical quality factor $3 \cdot 10^5$.

Figure 6.25: 3D view of the proposed resonator with a dielectric circular cylinder that can vary its depth into the rectangular cavity. The implementation is performed for a standard rectangular waveguide WR-34 ($a = 8.636$ mm and $b = 4.318$ mm), $l_{res} = 8$ mm, $h = 3.5$ mm and $d = 2$ mm. The parameter $offset_y$ controls how much the dielectric cylinder is inserted into the rectangular cavity.

In Figs. 6.26(a) and 6.26(b), the resonant mode chart and unloaded quality factors obtained for each mode are plotted. The second and third resonating modes are degenerated and present a good tuning range. Moreover, their unloaded quality factor are significantly high (more than 10^4 when the $offset_y$ is less than 2.3 mm). It should be mentioned that the fourth and fifth modes do not represent any problem if the structure is excited by a standard input waveguide with an U shaped iris windows, because their electric field components are orthogonal to the TE_{10} . However, the first mode is the fundamental one TE_{101} , which is perturbed by the dielectric circular cylinder. Therefore, the fundamental mode TE_{101} resonant frequency is lower than expected, due to the high permittivity dielectric cylinder used in this structure. Thus, the use second and third modes that are going to be used to synthesize a filter, is limited by the aforementioned first mode. The SFR between these two resonant modes and the first mode vary between two and eight depending of the configuration. Thus, this is an acceptable SFR for implementing a filter with this resonant modes.

The electric and magnetic field distributions of the selected modes are illustrated in Figs. 6.27(a) and 6.27(b), respectively. The magnetic field pattern indicates that the mode can be excited by a standard input waveguide and a U shape iris window, due to the H_x component. Although the option of employing the two degenerate resonant modes, as dual model cavities, could be appropriate to reduce the footprint and mass, for the sake of a better understanding of this filter, we are going to design it, using only one of the modes.

With these considerations in mind, a sixth order baseline filter is synthesized at center frequency $f_c = 23.875$ GHz and bandwidth $BW = 250$ MHz (the filter passband ranges from 23.75 GHz to 24 GHz, which is a fractional bandwidth of 1.05%). The filter and

(a) Frequency of the resonant modes of the cavity depicted in Fig. 6.25 as a function of the depth (Offset_y) of the dielectric circular cylinder inside of it.

(b) Unloaded quality factor of the cavity depicted in Fig. 6.25 as a function of the depth (Offset_y) of the dielectric cylinder inside of it.

Figure 6.26: Resonant modes and unloaded quality factors of them computed for the cavity illustrated in Fig. 6.25.

(a) Electric field distribution of the third resonant mode of the cavity depicted in Fig. 6.25 at 24.5 GHz.

(b) Magnetic field distribution of the third resonant mode of the cavity depicted in Fig. 6.25 at 24.5 GHz.

Figure 6.27: Electromagnetic fields of the third resonant mode of the cavity depicted in Fig. 6.25.

its dimensions are illustrated in Fig. 6.28. For this filter, an U shape iris windows is employed to implement the couplings between resonators, because this shape is better for keeping the coupling level as constant as possible when the tuning is applied. It should be mentioned that this filter can achieve high input coupling levels. Therefore, it is possible to design filters with wider bandwidths as compared to the previous option.

Figure 6.28: 3D view of a sixth order filter with dielectric circular cylinders that can vary its depth into the cavity. The implementation is performed for a standard rectangular waveguide WR-34 ($a = 8.636$ mm and $b = 4.318$ mm), $l_i = 1$ mm, $l_{res} = 8$ mm and $h_y = 3.5$ mm. The diameters of the cylinders (see Fig. 6.25) are $d_1 = 1.9868$ mm, $d_2 = 2.0065$ mm and $d_3 = 2.0098$ mm for the first, second and third resonator, respectively. The rest of the resonators are symmetrical with respect to the center plane from input to output. The coupling irises have a constant height $h_y = 2$ mm, whereas their widths are $w_{x1} = 5.62$ mm, $w_{x2} = 4.69$ mm, $w_{x3} = 4.37$ mm and $w_{x4} = 4.34$ mm.

Four tuning states of the filter are plotted in Fig. 6.29. Table 6.4 contains the center frequencies and bandwidths for each tuning state of the filter. These states show a degradation in the bandwidth of the filter. The state with $\text{offset}_y = 3$ mm leads to only 140 MHz bandwidth, which is a significant reduction if compared with the baseline filter used as reference (250 MHz). Nevertheless, for this state the filter center frequency has been shifted more than 2 GHz. It should be pointed out that the tuning range is considerable (almost 2 GHz) for this kind of filter.

Offset _y (mm)	f_c (GHz)	BW (MHz)
1	23.875	250
2	24.385	190
2.5	25.115	170
3	26.24	140

Table 6.4: Tuning states of the tunable filter, depicted in Fig. 6.28, by changing the depth of the dielectric circular cylinders.

A wide frequency range was evaluated to analyze the spurious free range of this filter. The scattering parameters for the two extreme configurations, for the states with

Figure 6.29: Scattering parameter magnitudes of the filter depicted in Fig. 6.28 for four different tuning states. Each configuration is obtained by only changing the offset_y parameter, according to Table 6.4.

$\text{offset}_y = 1$ mm and 3 mm, are plotted in Fig. 6.30. The frequency range analyzed range from the cutoff frequency of this standard waveguide to 28 GHz. As it was mentioned before, the filter produces an spurious mode at frequencies below its baseline passband. This spurious band can be moved really close to the passband when the filter is tuned to upper frequencies, as can be observed in the figure. Another option would be to employ the aforementioned spurious mode as the main mode for synthesizing tunable filters if a high quality factor is not a major requirement, being this one the fundamental mode for the filter.

Figure 6.30: Scattering parameters magnitudes of the filter depicted in Fig. 6.28 for the two extreme configurations ($\text{offset}_y = 1$ mm and $\text{offset}_y = 3$ mm). This figure also shows the spurious band that appears below the frequencies of the filter main passbands.

6.4 Conclusions

In this chapter, several mechanically tunable rectangular waveguide filters have been presented. Cavities with movable arms present interesting resonant modes that allow changing the resonant frequency within a wide tuning range. Together with the fact of the high precision that stepper motors can achieve, these resonators are an excellent option to implement tunable filters at high frequencies.

On the other hand, tunable ridge evanescent rectangle waveguide filters also present a wide tuning range. These filters provide a extremely good response in terms of spurious free range. Although a degradation in the bandwidth can be observed when the center frequency is downshifted, this kind of filters maintains a suitable level of return losses in the passband in a wide tuning range. Furthermore, an alternative to adjust the couplings with non-resonating resonators was developed. This filter can keep the constant absolute bandwidth within a 500 MHz of center frequency shift. In the worst case of the presented configuration, the filter will have a spurious free range over 1 GHz. In addition, it should be pointed out that the selectivity of the filter is still good, even considering the spurious mode excited by the non-resonating resonators.

Two filters were realized by using movable dielectric circular cylinders. These two filters present a considerably high tuning range. Besides, these filters can achieve a really high quality factor (can reach up to $36 \cdot 10^3$ for the selected mode of the horizontal dielectric cylinder resonator and $40 \cdot 10^3$ for the selected mode of the vertical dielectric cylinder resonator). Nevertheless, when the center frequency is shifted to higher frequencies the bandwidth decreases, due to the modification of the couplings. We have found that the shape of the employed coupling geometry is extremely important to obtain a lower coupling level variation with frequency, at least in some configurations. Despite this fact, the overall performance of these filters has a extremely good trade-off between quality factor, number of employed actuator drives (all cylinders in these examples are equally displaced), spurious free range and tuning range. This last structure studied can be employed for dual-mode resonators, which is interesting to reduce the total size of the filter. The employment of this dual-mode resonators is proposed for future works.

Chapter 7

Novel lowpass rectangular waveguide filters using dielectric and magnetic materials

In the first stage of the doctoral thesis, we investigated novel lowpass filter designs that improve the multipactor breakdown power threshold. The multipactor effect is a phenomenon that appears under vacuum conditions (commonly found in space communications systems) and could damage or even destroy the RF component [Vaughan, 1988]. This effect is produced by an electron avalanche that occurs when the free electrons, that are already present inside the RF-component, are accelerated due to the intense RF electric fields and impacts on a metallic plate. There exists a probability of a new electron released into the vacuum determined by the secondary emission yield (SEY) parameter (depends on the material) [Pomerantz and Marshall, 1951, Talcott, 1962, Simon and Williams, 1968]. These electrons can be accelerated again and impact on other metallic plate, releasing new electrons. If this provokes a self-sustained multiplication of free electrons, it can finally damage the RF component.

There are many scientific documents that analyze the multipactor effect in RF components with dielectric materials or ferrite materials [Keneshloo et al., 2012, Gonzalez-Iglesias et al., 2012, Gonzalez-Iglesias et al., 2014, Gonzalez-Iglesias et al., 2017, Cardona et al., 2018, Semenov et al., 2009, Hubble et al., 2017, Sounas et al., 2014]. The use of dielectric materials can modify the multipactor breakdown power threshold depending on the relative permittivity and the SEY coefficient of the material, as is documented in [Torregrosa-Penalva et al., 2010]. Based on these data, our proposal is to employ dielectric materials in certain critical areas of the filter in order to avoid the movement of free electrons in that areas. In this way we expect to increase the multipactor power threshold that is usually produced in electrically small gaps when the free electrons are accelerated by the fringing fields. The filters presented in this chapter were published in Quesada-Pereira et al., 2015.

On the other hand, it is interesting to analyze how magnetic materials interact when they are introduced in a lowpass filter. Here, we will show that these materials can be employed to make lowpass filters with transmission zeros.

7.1 Filter with dielectric coated metallic posts

Improving the multipactor breakdown power threshold in filters can be performed by using materials whose SEY parameter is lower than the SEY of the metal wall. An option to improve the multipactor threshold is to design a filter made with circular shaped metallic posts. The rounded surfaces scatters the impact of the electrons in many different directions, instead of on the same area, as it occurs in the filters with rectangular iris windows. Furthermore, if the posts are coated by a dielectric material whose SEY is low, then the multipactor breakdown power threshold could significantly be increased. A sixth order lowpass filter, illustrated in Fig. 7.1, is proposed as a design example to show the aforementioned concepts. The filter is synthesized with cutoff frequency $f_c = 11$ GHz, and return loss level $RL = 20$ dB by using the design technique presented in [Vera-Castejón et al., 2014]. For this design the employed dielectric material has a relative permittivity of $\epsilon_r = 2.8$, because it is the value of a material obtained from a commercial 3D printer that could be used for practical implementation of the idea.

Figure 7.1: Structure of a sixth order lowpass filter implemented by using posts coated by dielectric material acting as impedance inverters. The filter has been design in WR-75 standard rectangular waveguide ($a = 19.05$ mm and $b = 9.525$ mm). The thickness of the coating is $t_c = 1$ mm, whereas the minor semi-axes of the elliptic posts is $c = 1$ mm. Other dimensions are the major semi-axes of the elliptical posts $d_1 = 1.14$ mm, $d_2 = 2.45$ mm, $d_3 = 2.99$ mm, $d_4 = 3.11$ mm and the distances between the posts $l_1 = 10.48$ mm, $l_2 = 8.95$ mm and $l_3 = 8.28$ mm. The relative dielectric permittivity is $\epsilon_r = 2.8$.

The scattering parameters obtained for this filter are represented in Fig. 7.2, showing that the design process has been effective. The power handling capabilities of this filter will depend on the material employed in the coating of the posts. Thus, the rounded shape together with a dielectric material with a better SEY than commonly used metals will significantly improve the multipactor breakdown power threshold.

Figure 7.2: Scattering parameters of the filter presented in Fig. 7.1. The lowpass filter is implemented by using metallic posts coated by a dielectric material with constant thickness, as shown in Fig. 7.1.

7.2 Filter with metallic posts covered with dielectric material

To improve the multipactor breakdown power threshold with respect to the previous filter, the next step is to completely remove the gaps in the structure. This could be performed by using metallic posts that are fully covered by dielectric material, as shown in Fig. 7.3. This leads to a filter where free electron movement is not possible in critical areas. The sixth order lowpass filter cutoff frequency is $f_c = 5.5$ GHz, and its return loss level is $RL = 20$ dB, using this concept, is presented in Fig. 7.3. The dielectric material employed in this design is the same as the previous one. Furthermore, a good SEY is not strictly necessary for this filter as the electrons cannot move in critical gaps.

Figure 7.3: Structure of a sixth order lowpass filter implemented by using posts covered by dielectric material as impedance inverters. The filter has been designed in WR-187 standard rectangular waveguide ($a = 47.54$ mm and $b = 22.14$ mm). The dimensions of the impedance inverters are $w_d = 2.5$ mm, $w_c = 2$ mm, $h_1 = 4.06$ mm, $h_2 = 13.14$ mm, $h_3 = 15.79$ mm, $h_4 = 16.21$ mm. The distances between impedance inverters are $l_1 = 11.24$ mm, $l_2 = 8.03$ mm and $l_3 = 7.08$ mm. The relative dielectric permittivity is $\epsilon_r = 2.8$.

The scattering parameters, obtained from this filter, are plotted in Fig. 7.4. It should be noticed that even if it is not possible to trigger multipactor in the inverters as usual in critical gaps, it can occur in the rectangular waveguide sections that connect the inverters.

However, it is expected that the risk in these areas would be small, since they do not have strong electric fields.

Figure 7.4: Scattering parameter magnitude of the lowpass filter presented in Fig. 7.3. The filter is composed of rectangular posts covered with dielectric material from the bottom wall to the top wall of the rectangular waveguide.

7.3 Filter with inductive dielectric posts as inverters

Another option to avoid free electron movements in critical areas is to employ dielectric material cylinders oriented along the y axis (full inductive), as shown in the filter represented in Fig. 7.5. The filter cutoff frequency is $f_c = 5.5$ GHz and return loss level $RL = 20$ dB. Even though, the multipactor phenomenon can occur in sections out of the dielectrics, it is not probable if a high dielectric permittivity is employed for the design. This is due to the fact that the fields would be very confined inside the high permittivity dielectric material with low values in the vacuum regions.

Figure 7.5: Structure of a sixth order lowpass filter implemented by using dielectric material post as impedance inverters. The filter has been design in WR-187 standard rectangular waveguide ($a = 47.54$ mm and $b = 22.14$ mm). The lengths between posts are $l_1 = 19.85$ mm, $l_2 = 17.13$ mm, $l_3 = 16.16$ mm. The semiaxis of the elliptical post oriented in the propagation direction is $c = 2$ mm, while others are $d_1 = 0.61$ mm, $d_2 = 1.24$ mm, $d_3 = 1.55$ mm and $d_4 = 1.62$ mm. The dielectric relative permittivity employed for the post is $\epsilon_r = 20$.

The scattering parameters calculated for this filter are presented in Fig. 7.6. Despite the good in-band performance of the filter, the filter has a spurious band near to its passband. In fact, this spurious band is due to resonances that are excited inside the

dielectric elliptical cylinders, as it is evident from the figure where six resonances can be observed within the spurious band. It should be pointed out that even changing the permittivity, the resonances will be approximately at the same frequencies to achieve the impedance inverter value that the Chebyshev synthesis requires [Cameron et al., 2007].

Figure 7.6: Scattering parameter magnitudes of the filter presented in Fig. 7.5. The scattering parameters show spurious resonances produced by the dielectric posts between 7.5 GHz and 9.5 GHz.

7.4 Filter with magnetic materials

Another goal in lowpass filters is to achieve a configuration that allows to place transmission zeros near the passband, to improve the selectivity of the filter. However, in lowpass filters, an ideal response is difficult to be implemented with transmission zeros, as is typical in bandpass filters. Thus, one way to achieve transmission zeros is to introduce some resonant behavior in the filter. The proposed impedance inverter to implement this filter consist of one capacitive iris and two blocks of a material with relative permeability μ_r , as is shown in Fig. 7.7, where the structure of a six order filter, with cutoff frequency $f_c = 5.5$ GHz and return loss level $RL = 20$ dB, is represented.

The scattering parameter magnitudes, illustrated in Fig. 7.8, present two transmission zeros between 7 GHz and 8 GHz. These zeros are produced because the iris behaves as a resonator. Thanks to these zeros, the response achieves a better selectivity. In addition, this configuration makes the filter very compact when compared with traditional lowpass filters.

Figure 7.7: Structure of a sixth order lowpass filter implemented by using magnetic material slabs as part of the impedance inverter. The filter has been design in WR-187 standard waveguide ($a = 47.54$ mm and $b = 22.14$ mm). The lengths between posts are $l_1 = 6.78$ mm, $l_2 = 3.92$ mm, $l_3 = 3.2$ mm. The heights of the irises are: $h_1 = 11.99$ mm, $h_2 = 7.37$ mm, $h_3 = 5.27$ mm, $h_4 = 4.78$ mm, and the iris thickness is $w_c = 2$ mm. The magnetic materials posts dimensions are: $h_m = 16.61$ mm, $w_m = 1$ mm, whereas their separation from the capacitive irises is $w_g = 0.5$ mm. The relative permeability of the employed material is $\mu_r = 3$.

Figure 7.8: Scattering parameter magnitudes of the filter presented in Fig. 7.7. The filter is composed of impedance inverters realized with capacitive irises and magnetic materials near them.

7.5 Conclusions

In this chapter, four kinds of lowpass filters are presented. The filters with dielectric material could improve the multipactor breakdown power threshold if a good selection of the materials is done. In addition, by using dielectric materials from the bottom wall to the top wall of the rectangular waveguide, could also improve the multipactor breakdown power threshold, since free electron movement is not possible in areas where the electromagnetic fields are intense. In spite of this, the multipactor phenomenon can still occur in the rectangular waveguide sections between the impedance inverters. Moreover, this chapter presents for the first time fully inductive rectangular waveguide lowpass filters by using dielectric materials. As drawback, these filters present a spurious band produced by resonances excited inside the dielectric posts. However, these resonances do not affect to the response of the lowpass filter, while at the same time the multipactor effect could only be produced in sections with non-intense electromagnetic fields. On the other hand, lowpass filters with magnetic materials are studied for first time. These filters achieve a lowpass filtering function where the irises become resonant. These resonant irises produce transmission zeros that can be placed near the passband, as it was demonstrated using simulations, improving the selectivity of filters.

Chapter 8

Conclusions

In this PhD. thesis, tunable filters with both electrical and mechanical tuning are presented. All these filters show a good performance in terms of the tuning range, which is suitable for many commercial applications. Furthermore, some bandwidth tuning mechanisms based on non-resonating resonators were proposed.

In Chapter 2, the relevant design techniques for reconfigurable filters were presented. The first section shows a design technique that starts from the coupling matrix and allows designing a filter by using scattering parameters simulations. This design technique is extremely useful for electronically tunable filters, where the typical eigensolver solutions to compute the couplings are not a good option due to the need to connect capacitors through external ports. The presented second design technique is based on the scaled coupling matrix, which achieves exceptional initial dimensions for the filters, even for complex topologies (quartets, trisections or folded canonical topologies in general). This design technique also provides a way to design filters with non-resonating resonators (NRR) as couplings in a natural way. Thus, some filters were designed to study the influence of NRRs in their responses. The main conclusion is that the NRRs can be used as couplings, but they generate some spurious at the frequencies where they are resonating.

In Chapter 3, the design of a novel SIW filter based on varactor diodes with a high tuning range was presented. This filter allows the introduction of non-resonating resonators as couplings. As it was demonstrated, this filter can operate in Ku-band with capacitances in the range suitable for commercial varactor diodes. Moreover, it is demonstrated that this resonating mode can be employed for the design of a new rectangular waveguide filter. The particularity of this rectangular waveguide filter is that the height of the waveguide is considerably reduced in order to excite the relevant mode at high frequencies. Moreover, this filter included a bandwidth tuning mechanism based on MEMS, which does not generate additional spurious resonances. However, each MEMS only provides two different coupling states. The importance of this coupling tuning mechanism is that it can be included in many filter configurations.

Chapter 4 shows the design of tunable evanescent rectangular waveguide filters with varactor diodes. These evanescent filter provides a wide tunability range using capacitances within the range of commercial varactors. In this chapter, we have used dielectric

substrates as additional elements, which provide support for the electrical components needed for tunability purposes. At the same time these dielectric substrates provide shielding to avoid RF-Leakage. Therefore, the introduction of SIW structures makes a huge difference in the RF-Leakage levels as compared to empty regular waveguide technology. Overall, the filter displays a suitable tuning range in Ku-band, with a moderate bandwidth reduction.

Chapter 5 introduces a new concept in tunable filters, which is based on tunable artificial magnetic conductors (AMCs). In the chapter, it is described the design process for tunable artificial magnetic conductors, whose purpose is to be used as tunable elements in rectangular waveguide filters. An in-depth study of the parameters involved in the design of tunable filters using these AMCs is carried out, resulting in a design criterion that can be followed to implement this kind of structures. Moreover, this chapter provides a bandwidth tuning mechanism based on floating posts. This implementation can be hard for a practical implementation, but it displays a remarkable tuning control. On the other hand, each unit cell of the AMC can be tuned in a different way following a tapering function. This solution requires a more complex feeding network, but it achieves almost a constant bandwidth tuning in the filter, without the inclusion of additional components. Furthermore, these filters were realized with the resonating mode TE_{102} in standard rectangular waveguide. Thus, it is reasonable to expect a high quality factors in this kind of filters. The tuning range that can be achieved with the capacitances, provided by commercial varactor diodes, is more than enough for many communication systems, being minimum the RF-Leakage.

In Chapter 6, we explore some filters based on mechanical tuning. The first section shows resonators based on angular movable parts, which provide tuning ranges of more than 1.5 GHz in Ka-band in some modes and with unloaded quality factors over 5000. The second section is dedicated to evanescent rectangular waveguide filters. These filters are always attractive, due to the wide spurious free ranges that can be achieved. Furthermore, another evanescent filter with NRR is included in this section, which demonstrates a structure able to maintain the absolute bandwidth of the filter in a 500 MHz range. This filter has a spurious free range of 1 GHz with the NRR included. The last section presents resonators tuned by movable dielectric circular cylinders. These resonators display extremely good performance in terms of quality factors (above $3 \cdot 10^5$ in some configurations) and tuning range. On the other hand, the drawbacks are the moderate coupling variation, which is expected in any tunable filter, and the bandwidth limitations due to the low input coupling that can be obtained. However, due to the achieved unloaded high-quality factors, these filters are really interesting for commercial applications, taking in account that some bandwidth tuning can be introduced. For instance, NRR can be used to provide tunability in the couplings, and the input coupling might be improved with a coaxial probe input, instead of using a the traditional rectangular waveguide input.

Finally, in Chapter 7, some novel lowpass filters are presented. These filters are intended to improve the multipactor breakdown power threshold. The main idea of this chapter is to use dielectric materials in areas where the filter is more sensitive to the multipactor effect. As it was shown in the chapter, there are many configurations of impedance inverters where the dielectric material can be easily introduced. Also, the employment of magnetic materials is studied, resulting in a filter with transmission zeros.

These transmission zeros are generated due to the resonant behavior produced by the magnetic material in close proximity to the iris, making them to be resonant. This is a relevant result, taking into account that lowpass filters are usually implemented without transmission zeros.

Some future lines can be extracted from these results:

- Coupling tuning mechanism based on non-resonating resonators (NRR).

This coupling mechanism allows its inclusion in many filter configurations, so it is relevant. However, their resonances are finally translated into spurious from the filter passband point of view. Thus, a better understanding of them can be extremely useful to improve the ratio between coupling variation and spurious free range. This approximation was utilized in Chapter 3, providing a reasonable spurious free range, while the bandwidth of the filter is maintained along its different tuning states.

- Implementation of parallel-connected two-port networks for evanescent resonators.

The tunable resonators presented in Chapter 4 can be used to implement filters with parallel-connected networks topologies, like the presented in [Cameron et al., 2007, Ch. 9, p. 359]. A fourth order filter coupling diagram is illustrated in Fig. 8.1 as example. By using this kind of topology it is possible to move the resonators asymmetrically, resulting in a bandwidth control.

Figure 8.1: Coupling-routing diagram, of a parallel-connected two-port network. This topology represents a fourth order filter.

- Alternative implementations of tunable artificial magnetic conductors (AMCs).

In Chapter 5, we developed some simple designs for AMCs. Although these AMCs are valid for making tunable filters, it is interesting to explore other designs that can perform better in the filter or might reduce the spurious band that is generated due to this structure.

- Exploring alternative coupling tuning mechanism for filters implemented with one AMC.

In Chapter 5, we explore a coupling tuning mechanism based on connecting the floating post, used as couplings, through a varactor to the top and bottom walls. The main issue is that this configuration requires two AMCs to work properly. This two AMCs leads to the spurious bands as was demonstrated. As a future line, it is proposed a configuration with small floating post using MEMS or varactors for a configuration of the waveguide with only one AMC acting as side wall, as shown in Fig. 8.2.

Figure 8.2: Second order filter with an AMC acting as side-wall. The couplings are implemented with floating posts connected though MEMS to the top and bottom walls.

- Resonators tuned by movable arms.

In Chapter 6, the resonators with movable arms display good performance in terms of tuning range and quality factors, especially those implemented with dielectric blocks. An interesting future line is to implement filters with these resonators and study their parameters in terms of bandwidth reduction and the possibility of obtaining folded canonical topologies with them.

- Coupling geometries for constant bandwidth tunable filters.

In Chapter 6, the filters with movable circular cylinders lead to really good results, but they have their minor problems with the variation of couplings in different tuning states. Two research future lines can be followed to improve these designs. The first one is to introduce NRRs to solve the problems with the coupling reduction for the applications that require it. The second one is to find other coupling geometries that result in a constant tuning bandwidth. In order to improve the geometries, a possible future line is to employ additive manufacturing because the standard coupling windows that can be implemented with standard manufacturing techniques are very limited.

List of publications

Journals

1. Quesada-Pereira, F. D., Romera-Pérez, A., Vera-Castejón, P., Álvarez-Melcón, A. (2015). Integral-equation formulation for the analysis of capacitive waveguide filters containing dielectric and metallic arbitrarily shaped objects and novel applications. *IEEE Transactions on Microwave Theory and Techniques*, Volume 63, Pages: 3862–3873.

In [Quesada-Pereira et al., 2015] are included the most relevant lowpass filter designs that are presented in Chapter 7. This article presents an efficient integral-equation formulation that is employed for analyzing arbitrarily shaped objects in capacitive waveguide filters.

2. Martínez-Martínez, D., Pons-Abenza, A., Romera-Pérez, A., Hinojosa, J., Quesada-Pereira, F., Álvarez-Melcón, A., and Guglielmi, M. (2018). Advanced filter design technique based on equivalent circuits and coupling matrix segmentation. *International Journal of Circuit Theory and Applications*. Volume 46, 2018, Pages: 1055–1071.

This paper presents an application of the step by step design technique presented in 2 to implement dual-mode all inductive rectangular waveguide filters.

3. Romera-Pérez, A., Pons-Abenza, A., Martínez-Martínez, D., Álvarez-Melcón, A., and Quesada-Pereira, F. D. (2019). Filter design for folded canonical topologies based on equivalent circuit segmentation. *AEU - International Journal of Electronics and Communications*. Volume 109, 2019, Pages 157–165, ISSN 1434-8411, <https://doi.org/10.1016/j.aeue.2019.07.010>.

In [Romera-Pérez et al., 2019a] it is proposed a strategy to apply the step by step technique to design folded canonical topologies. This strategy is based on using multiple ports to compute the multiple paths that the signal can follow. This paper corresponds to the section 2.3.

4. Alejandro Pons-Abenza, José-María García-Barceló, Antonio Romera-Pérez, Alejandro Alvarez-Melcon, Fernando D. Quesada-Pereira, Juan Hinojosa-Jiménez, Marco Guglielmi, Vicente E. Boria Esbert, Lara Arche-Andradas. Design and Implementation of Evanescent Mode Waveguide Filters Using Dielectrics and Additive Manufacturing Techniques, *AEU - International Journal of Electronics and Communications*, 2020, 153065, ISSN 1434-8411, <https://doi.org/10.1016/j.aeue.2020.153065>.

Conferences

1. Romera-Pérez, A., Pons-Abenza, A., Quesada-Pereira, F. D., Álvarez-Melcón, A. (2017). Diseño de topologías complejas de filtros de microondas en tecnología guía de onda mediante la técnica de segmentación de la matriz de acoplamientos. In Symposium nacional de la unión científica internacional de radio (URSI), ETSI, Cartagena, 7-9 September 2017.

In this conference, the step by step design technique for folded canonical topologies of section 2.3 was presented. To illustrate the concept, a quartet topology filters is designed to show a practical example of the proposed design technique.

2. Romera-Pérez, A., Pons-Abenza, A., Quesada-Pereira, F. D., Álvarez-Melcón, A., and Arche-Andradas, L. (2019). A continuously tunable Ku-band third order evanescent waveguide filter based on varicap diodes. In 4th Workshop on advanced flexible telecom payloads, The Netherlands, Noordwijk, ESTEC, ESA, 4-6 March 2019.

In this paper, it was presented the tunable evanescent waveguide filter, shown in Chapter 4, implemented using varactor diodes.

3. Romera-Pérez, A., Pons-Abenza, A., Quesada-Pereira, F. D., Álvarez-Melcón, A., and Arche-Andradas, L. (2019). Filtro evanescente en banda Ku de tercer orden resintonizable de forma continua con diodos varicap. In Symposium nacional de la unión científica internacional de radio (URSI), ETSI, Sevilla, 4-6 September 2019.

In this conference, the latest results of the evanescent waveguide filter, shown in Chapter 4, was presented.

Bibliography

- [Acar et al., 2015] Acar, O., Johansen, T. K., and Zhurbenko, V. (2015). A high-power low-loss continuously tunable bandpass filter with transversely biased ferrite-loaded coaxial resonators. *IEEE Transactions on Microwave Theory and Techniques*, 63(10):3425–3432.
- [Adhikari et al., 2013] Adhikari, S., Ghiotto, A., and Wu, K. (2013). Simultaneous electric and magnetic two-dimensionally tuned parameter-agile SIW devices. *IEEE Transactions on Microwave Theory and Techniques*, 61(1):423–435.
- [Aitchison et al., 2001] Aitchison, C., Mbabele, M., Moazzam, M., Budimir, D., and Ali, F. (2001). Improvement of third-order intermodulation product of RF and microwave amplifiers by injection. *IEEE Transactions on Microwave Theory and Techniques*, 49(6):1148–1154.
- [Anand and Liu, 2015] Anand, A. and Liu, X. (2015). Capacitively tuned electrical coupling for reconfigurable coaxial cavity bandstop filters. In *2015 IEEE MTT-S International Microwave Symposium*. IEEE.
- [Anand et al., 2013] Anand, A., Small, J., Peroulis, D., and Liu, X. (2013). Theory and design of octave tunable filters with lumped tuning elements. *IEEE Transactions on Microwave Theory and Techniques*, 61(12):4353–4364.
- [Ansoft, 2019] Ansoft (2019). Ansys electronic desktop, HFSS. www.ansoft.com.
- [Aouidad et al., 2015] Aouidad, H., Favennec, J.-F., Rius, E., Manchec, A., and Clavet, Y. (2015). A tunable filter based on miniature SIR coaxial resonators. In *2015 European Microwave Conference (EuMC)*. IEEE.
- [Arnold et al., 2015] Arnold, C., Parlebas, J., and Zwick, T. (2015). Center frequency and bandwidth tunable waveguide bandpass filter with transmission zeros. In *2015 European Microwave Conference (EuMC)*. IEEE.
- [Arnold et al., 2014a] Arnold, C., Parlebas, J., and Zwick, T. (2014a). Reconfigurable waveguide filter with variable bandwidth and center frequency. *IEEE Transactions on Microwave Theory and Techniques*, 62(8):1663–1670.
- [Arnold et al., 2014b] Arnold, C., Parlebas, J., and Zwick, T. (2014b). Reconfigurable waveguide filter with variable bandwidth and center frequency. *IEEE Transactions on Microwave Theory and Techniques*, 62(8):1663–1670.
- [Arregui et al., 2010] Arregui, I., Arnedo, I., Lujambio, A., Chudzik, M., Benito, D., Jost, R., Gortz, F. J., Lopetegi, T., and Laso, M. A. G. (2010). A compact design of high-power spurious-free low-pass waveguide filter. *IEEE Microwave and Wireless Components Letters*, 20(11):595–597.

- [Atia and Williams, 1972] Atia, A. E. and Williams, A. E. (1972). Narrow-bandpass waveguide filters. *IEEE Transactions on Microwave Theory and Techniques*, 20(4):258–265.
- [Bandler et al., 1994] Bandler, J. W., Biernacki, R. M., Chen, S. H., Grobelny, P. A., and Hemmers, R. H. (1994). Space mapping technique for electromagnetic simulation. 42(12):2536–2544.
- [Barrera and Huff, 2013] Barrera, J. D. and Huff, G. H. (2013). Analysis of a variable SIW resonator enabled by dielectric material perturbations and applications. *IEEE Transactions on Microwave Theory and Techniques*, 61(1):225–233.
- [Basavarajappa and Mansour, 2018] Basavarajappa, G. and Mansour, R. R. (2018). Design methodology of a tunable waveguide filter with a constant absolute bandwidth using a single tuning element. *IEEE Transactions on Microwave Theory and Techniques*, 66(12):5632–5639.
- [Bozzi et al., 2011] Bozzi, M., Georgiadis, A., and Wu, K. (2011). Review of substrate-integrated waveguide circuits and antennas. *IET Microwaves, Antennas & Propagation*, 5(8):909.
- [Cameron et al., 2007] Cameron, R. J., Kudsia, C. M., and Mansour, R. R. (2007). *Microwave Filters for Communication Systems*, pages 432–467. Wiley. ISBN: 978-0-471-45022-1.
- [Cardona et al., 2018] Cardona, V., Joaquin, J., Boria, Vicente, E., Guglielmi, Marco, Melgarejo, and Carlos, J. (2018). Multipactor characterization of ferrite materials for space applications. In *2018 IEEE MTT-S International Conference on Numerical Electromagnetic and Multiphysics Modeling and Optimization (NEMO)*. IEEE.
- [Cho and Rebeiz, 2015] Cho, Y.-H. and Rebeiz, G. M. (2015). Tunable 4-pole noncontiguous 0.7–2.1-GHz bandpass filters based on dual zero-value couplings. *IEEE Transactions on Microwave Theory and Techniques*, 63(5):1579–1586.
- [Cogollos et al., 2012] Cogollos, S., Brumos, M., Boria, V. E., Vicente, C., Gil, J., Gimeno, B., and Guglielmi, M. (2012). A systematic design procedure of classical dual-mode circular waveguide filters using an equivalent distributed model. *IEEE Transactions on Microwave Theory and Techniques*, 60(4):1006–1017.
- [Cucini et al., 2004] Cucini, A., Caiazzo, M., Bennati, P., and Maci, S. (2004). Quasi-TEM waveguides realized by FSS-walls. In *IEEE Antennas and Propagation Society Symposium, 2004*. IEEE.
- [Dang et al., 2015] Dang, J. H., Gough, R. C., Morishita, A. M., Ohta, A. T., and Shiroshima, W. A. (2015). A tunable X-band substrate integrated waveguide cavity filter using reconfigurable liquid-metal perturbing posts. In *2015 IEEE MTT-S International Microwave Symposium*. IEEE.
- [de Cos et al., 2010] de Cos, M. E., Alvarez, Y., Cosmina-Hadarig, R., and Las-Heras, F. (2010). Novel SHF-band uniplanar artificial magnetic conductor. *IEEE Antennas and Wireless Propagation Letters*, 9:44–47.
- [De Los Santos, 2002] De Los Santos, H. J. (2002). *RF MEMS circuit design for wireless communications*. Artech House.

- [Deslandes and Wu, 2003] Deslandes, D. and Wu, K. (2003). Single-substrate integration technique of planar circuits and waveguide filters. *IEEE Transactions on Microwave Theory and Techniques*, 51(2):593–596.
- [Foroozesh and Shafai, 2009] Foroozesh, A. and Shafai, L. (2009). Effects of artificial magnetic conductors in the design of low-profile high-gain planar antennas with high-permittivity dielectric superstrate. *IEEE Antennas and Wireless Propagation Letters*, 8:10–13.
- [Franke et al., 2014] Franke, T., Gaebler, A., Prasetyadi, A. E., and Jakoby, R. (2014). Tunable Ka-band waveguide resonators and a small band band-pass filter based on liquid crystals. In *2014 44th European Microwave Conference*. IEEE.
- [Fu and Yuan, 2005] Fu, Y. and Yuan, N. (2005). Reflection and transmission between rectangular waveguide and EBG based TEM waveguide. In *2005 Asia-Pacific Microwave Conference Proceedings*. IEEE.
- [Goldberg, 1989] Goldberg, D. (1989). *Genetic Algorithm in Search, Optimization and Machine Learning*, chapter 3, pages 60–88. Addison-Wesley.
- [Gonzalez-Iglesias et al., 2014] Gonzalez-Iglesias, D., Gimeno, B., Boria, V. E., Gomez, A., and Vegas, A. (2014). Multipactor effect in a parallel-plate waveguide partially filled with magnetized ferrite. *IEEE Transactions on Electron Devices*, 61(7):2552–2557.
- [Gonzalez-Iglesias et al., 2017] Gonzalez-Iglesias, D., Monerris, O., Gimeno, B., Diaz, M. E., Boria, V. E., Martin, P., Gomez, A., Fernandez, O., Vegas, A., Casas, F., Anza, S., Vicente, C., Gil, J., Mata, R., Montero, I., and Raboso, D. (2017). Novel multipactor studies in RF satellite payloads: Single-carrier digital modulated signals and ferrite materials. In *2017 IEEE MTT-S International Conference on Numerical Electromagnetic and Multiphysics Modeling and Optimization for RF, Microwave, and Terahertz Applications (NEMO)*. IEEE.
- [Gonzalez-Iglesias et al., 2012] Gonzalez-Iglesias, D., Soto, P., Anza, S., Gimeno, B., Boria, V. E., Vicente, C., and Gil, J. (2012). Multipactor susceptibility charts for ridge and multiridge waveguides. *IEEE Transactions on Electron Devices*, 59(12):3601–3607.
- [Goudos and Sahalos, 2010] Goudos, S. K. and Sahalos, J. N. (2010). Pareto optimal microwave filter design using multiobjective differential evolution. *IEEE Transactions on Antennas and Propagation*, 58(1):132–144.
- [Guglielmi, 1994] Guglielmi, M. (1994). Simple CAD procedure for microwave filters and multiplexers. *IEEE Transactions on Microwave Theory and Techniques*, 42(7):1347–1352.
- [Guglielmi and Melcon, 1993] Guglielmi, M. and Melcon, A. A. (1993). Novel design procedure for microwave filters. In *1993 23rd European Microwave Conference*, pages 212–213.
- [Hinojosa et al., 2011] Hinojosa, J., Pereira, F. D. Q., Martinez-Mendoza, M., and Alvarez-Melcon, A. (2011). Optimization-oriented design of RF/microwave circuits using inverse-linear-input neuro-fuzzy-output space mapping with two different dimensionality simulators. *Circuits and Systems I: Regular Papers, IEEE Transactions on*, 58(1):176–185.

- [Hinojosa et al., 2015] Hinojosa, J., Quesada-Pereira, F. D., Bozzi, M., and Alvarez-Melcon, A. (2015). Efficient optimization-oriented design methodology of high-order 3-D filters using 2-D and 3-D electromagnetic simulators. *International Journal of Circuit Theory and Applications*, 43(10):1431–1445.
- [Hong and Lancaster, 2001] Hong, J. and Lancaster, M. (2001). *Microstrip filters for RF/Microwave Applications*. John Wiley & Sons, USA.
- [Hosseini and Hakkak, 2008] Hosseini, M. and Hakkak, M. (2008). Characteristics estimation for jerusalem cross-based artificial magnetic conductors. *IEEE Antennas and Wireless Propagation Letters*, 7:58–61.
- [Hosseinipanah et al., 2007] Hosseinipanah, M., Wu, Q., and Fu, J. (2007). Design and simulation of a novel two-layer artificial magnetic conductors. In *2007 International Symposium on Microwave, Antenna, Propagation and EMC Technologies for Wireless Communications*. IEEE.
- [Huang et al., 2016] Huang, F., Zhou, J., and Hong, W. (2016). Ku band continuously tunable circular cavity SIW filter with one parameter. *IEEE Microwave and Wireless Components Letters*, 26(4):270–272.
- [Hubble et al., 2017] Hubble, A. A., Chaplin, V. H., Clements, K. A., Spektor, R., Partridge, P. T., and Graves, T. P. (2017). Multipactor breakdown threshold reduction due to magnetic confinement in parallel fields. *IEEE Transactions on Plasma Science*, 45(7):1726–1730.
- [Iskander et al., 2014] Iskander, M. A., Nasresfahani, M., and Mansour, R. R. (2014). A constant-Q tunable combline bandpass filter using angular tuning technique. In *2014 44th European Microwave Conference*. IEEE.
- [Jiang et al., 2019] Jiang, D., Liu, Y., Li, X., Wang, G., and Zheng, Z. (2019). Tunable microwave bandpass filters with complementary split ring resonator and liquid crystal materials. *IEEE Access*, pages 1–1.
- [Keneshloo et al., 2012] Keneshloo, R., Dadashzadeh, G., and Okhovvat, M. (2012). Investigation of multipactor effect on return loss degradation on dielectric resonator filters. In *2012 15 International Symposium on Antenna Technology and Applied Electromagnetics*. IEEE.
- [Lai and Jeng, 2006] Lai, M.-I. and Jeng, S.-K. (2006). Compact microscript dual-band bandpass filters design using genetic-algorithm techniques. 54(1):160–168.
- [Lee et al., 2015] Lee, B., Nam, S., Koh, B., Kwak, C., and Lee, J. (2015). K-band frequency tunable substrate-integrated-waveguide resonator filter with enhanced stopband attenuation. *IEEE Transactions on Microwave Theory and Techniques*, 63(11):3632–3640.
- [Levy, 1973] Levy, R. (1973). Tapered corrugated waveguide low-pass filters. *IEEE Transactions on Microwave Theory and Techniques*, 21(8):526–532.
- [Li and Wu, 2010] Li, D. and Wu, K. (2010). Geometric characteristics, physical mechanism, and electrical analysis of quasi-TEM waveguides with dipole-FSS walls. In *2010 IEEE MTT-S International Microwave Symposium*. IEEE.

- [Liu et al., 2012] Liu, X., Katehi, L. P. B., Chappell, W. J., and Peroulis, D. (2012). Power handling of electrostatic MEMS evanescent-mode (EVA) tunable bandpass filters. *IEEE Transactions on Microwave Theory and Techniques*, 60(2):270–283.
- [Lu et al., 2010] Lu, W., Wang, L., Gu, S., Aplin, D. P. R., Estrada, D. M., Yu, P. K. L., and Asbeck, P. M. (2010). InGaN/GaN schottky diodes with enhanced voltage handling capability for varactor applications. *IEEE Electron Device Letters*, 31(10):1119–1121.
- [Mansour, 2015] Mansour, R. R. (2015). High-Q tunable filters for wireless base station applications. In *2015 IEEE International Conference on Microwaves, Communications, Antennas and Electronic Systems (COMCAS)*. IEEE.
- [Mansour et al., 2014] Mansour, R. R., Huang, F., Fouladi, S., Yan, W. D., and Nasr, M. (2014). High-Q tunable filters: Challenges and potential. *IEEE Microwave Magazine*, 15(5):70–82.
- [Martínez-Martínez et al., 2018] Martínez-Martínez, D., Pons-Abenza, A., Romera-Pérez, A., Hinojosa, J., Quesada-Pereira, F., Álvarez-Melcón, A., and Guglielmi, M. (2018). Advanced filter design technique based on equivalent circuits and coupling matrix segmentation. *International Journal of Circuit Theory and Applications*, 46(5):1055–1071.
- [Matthaei and Yones, 1980] Matthaei, G. and Yones, E. (1980). *Microwave Filters, Impedance Matching Networks, and Coupling Structures*. Artech House, Boston, Massachusetts, USA.
- [Máximo-Gutiérrez, 2015] Máximo-Gutiérrez, C. (2015). Diseño de filtros paso-bajo en tecnología de guía de onda integrada en sustrato (siw), para aplicaciones espaciales. <http://repositorio.upct.es/handle/10317/5010>.
- [Máximo-Gutiérrez et al., 2019] Máximo-Gutiérrez, C., Hinojosa, J., and Álvarez-Melcón, A. (2019). Design of wide band-pass substrate integrated waveguide (SIW) filters based on stepped impedances. *AEU - International Journal of Electronics and Communications*, 100:1–8.
- [Nam et al., 2018] Nam, S., Lee, B., Kwak, C., and Lee, J. (2018). A new class of K-band high-Q frequency-tunable circular cavity filter. *IEEE Transactions on Microwave Theory and Techniques*, 66(3):1228–1237.
- [Pelliccia et al., 2015] Pelliccia, L., Cacciamani, F., Farinelli, P., and Sorrentino, R. (2015). High-Q tunable waveguide filters using ohmic RF MEMS switches. *IEEE Transactions on Microwave Theory and Techniques*, 63(10):3381–3390.
- [Pelliccia et al., 2012] Pelliccia, L., Cacciamani, F., Sorrentino, R., Farinelli, P., Ligander, P., and Persson, O. (2012). High-Q MEMS-tunable waveguide filters in k-band. In *2012 42nd European Microwave Conference*. IEEE.
- [Peng et al., 2015] Peng, Y., Wang, T., Jiang, W., Rahman, B. M. F., Xia, T., and Wang, G. (2015). Electrically tunable bandpass filter with patterned permalloy thin-film-enabled engineered substrate. *IEEE Transactions on Magnetics*, 51(11):1–4.
- [Polat et al., 2019] Polat, E., Reese, R., Jost, M., Schuster, C., Nickel, M., Jakoby, R., and Maune, H. (2019). Tunable liquid crystal filter in nonradiative dielectric waveguide technology at 60 GHz. *IEEE Microwave and Wireless Components Letters*, 29(1):44–46.

- [Pomerantz and Marshall, 1951] Pomerantz, M. and Marshall, J. (1951). Fundamentals of secondary electron emission. *Proceedings of the IRE*, 39(11):1367–1373.
- [Quesada-Pereira et al., 2015] Quesada-Pereira, F. D., Romera-Pérez, A., Vera-Castejón, P., and Álvarez-Melcón, A. (2015). Integral-equation formulation for the analysis of capacitive waveguide filters containing dielectric and metallic arbitrarily shaped objects and novel applications. *IEEE Transactions on Microwave Theory and Techniques*, 63(12):3862–3873.
- [Rajagopal et al., 2017] Rajagopal, S., Chennakesavan, G., Subburaj, D. R. P., Srinivasan, R., and Varadhan, A. (2017). A dual polarized antenna on a novel broadband multilayer artificial magnetic conductor backed surface for LTE/CDMA/GSM base station applications. *AEU - International Journal of Electronics and Communications*, 80:73–79.
- [Rayas-Sanchez and Gutierrez-Ayala, 2008] Rayas-Sanchez, J. E. and Gutierrez-Ayala, V. (2008). A general EM-based design procedure for single-layer substrate integrated waveguide interconnects with microstrip transitions. In *2008 IEEE MTT-S International Microwave Symposium Digest*. IEEE.
- [Romera-Pérez et al., 2019a] Romera-Pérez, A., Pons-Abenza, A., Martínez-Martínez, D., Álvarez-Melcón, A., and Quesada-Pereira, F. D. (2019a). Filter design for folded canonical topologies based on equivalent circuit segmentation. *AEU - International Journal of Electronics and Communications*.
- [Romera-Pérez et al., 2019b] Romera-Pérez, A., Pons-Abenza, A., Quesada-Pereira, F. D., Álvarez-Melcón, A., and Arche-Andradas, L. (2019b). Filtro evanescente en banda Ku de tercer orden resintonizable de forma continua con diodos varicap. In *Simposium nacional de la unión científica internacional de radio (URSI)*, Sevilla.
- [Romera-Pérez et al., 2017] Romera-Pérez, A., Quesada-Pereira, F. D., Álvarez-Melcón, A., and Pons-Abenza, A. (2017). Diseño de topologías complejas de filtros de microondas en tecnología guía de onda mediante la técnica de segmentación de la matriz de acoplamientos. In *Simposium nacional de la unión científica internacional de radio (URSI)*, ESTEC. ESA.
- [Romera-Pérez et al., 2019c] Romera-Pérez, A., Quesada-Pereira, F. D., Álvarez-Melcón, A., and Pons-Abenza, A. (2019c). A continuously tunable Ku-band third order evanescent waveguide filter based on varicap diodes. In *4th Workshop on advanced flexible telecom payloads*.
- [Rosenberg et al., 2016] Rosenberg, U., Beyer, R., Kraus, P., Sieverding, T., Papanastasiou, A., Pueyo-Tolosa, M., Iglesias, P. M., and Ernst, C. (2016). Reconfigurable doublet dual-mode cavity filter designs providing remote controlled center frequency and bandwidth re-allocation. In *2016 46th European Microwave Conference (EuMC)*. IEEE.
- [Saeedi et al., 2014] Saeedi, S., Lee, J., and Sigmarsson, H. (2014). Broadband implementation of tunable, substrate-integrated, evanescent-mode, cavity bandpass filters. In *2014 44th European Microwave Conference*. IEEE.
- [Samani and Safian, 2010] Samani, M. F. and Safian, R. (2010). On bandwidth limitation and operating frequency in artificial magnetic conductors. *IEEE Antennas and Wireless Propagation Letters*, 9:228–231.

- [Sans et al., 2014] Sans, M., Selga, J., Rodriguez, A., Bonache, J., Boria, V. E., and Martin, F. (2014). Design of planar wideband bandpass filters from specifications using a two-step aggressive space mapping (ASM) optimization algorithm. 62(12):3341–3350.
- [Sekar et al., 2011] Sekar, V., Armendariz, M., and Entesari, K. (2011). A 1.2–1.6-GHz substrate-integrated-waveguide RF MEMS tunable filter. *IEEE Transactions on Microwave Theory and Techniques*, 59(4):866–876.
- [Sekar and Entesari, 2012] Sekar, V. and Entesari, K. (2012). A half-mode substrate-integrated-waveguide tunable filter using packaged RF MEMS switches. *IEEE Microwave and Wireless Components Letters*, 22(7):336–338.
- [Semenov et al., 2009] Semenov, V., Rakova, E., Nefedov, I. M., Shereshevski, I. A., Vdovicheva, N. K., Udiljak, R., Anderson, D., Lisak, M., and Puech, J. (2009). Multipactor breakdown in waveguide irises. In *2009 IEEE International Vacuum Electronics Conference*. IEEE.
- [Shen et al., 2015] Shen, M., Huang, Y., and Shao, Z. (2015). A tunable microstrip dual-mode bandpass filter based on liquid crystal technology. In *2015 IEEE International Conference on Communication Problem-Solving (ICCP)*. IEEE.
- [Simon and Williams, 1968] Simon, R. E. and Williams, B. F. (1968). Secondary-electron emission. *IEEE Transactions on Nuclear Science*, 15(3):167–170.
- [Sirici et al., 2012] Sirici, S., Martinez, J., Taroncher, M., and Boria, V. (2012). Analog tuning of compact varactor-loaded combline filters in substrate integrated waveguide. In *2012 42nd European Microwave Conference*. IEEE.
- [Skyworks, 2019] Skyworks (2019). <http://www.skyworksinc.com/>.
- [Sounas et al., 2014] Sounas, A. L., Sorolla, E., and Mattes, M. (2014). Multipactor analysis in parallel-plate waveguides loaded by a single dielectric slab. In *The 8th European Conference on Antennas and Propagation (EuCAP 2014)*. IEEE.
- [Talcott, 1962] Talcott, R. (1962). The effects of titanium films on secondary electron emission phenomena in resonant cavities and at dielectric surfaces. *IRE Transactions on Electron Devices*, 9(5):405–410.
- [Torregrosa-Penalva et al., 2010] Torregrosa-Penalva, G., Coves, Á., Gimeno-Martinez, B., Montero, I., Vicente, C., and Boria, V. E. (2010). Multipactor susceptibility charts of a parallel-plate dielectric-loaded waveguide. *IEEE Transactions on Electron Devices*, 57(5):1160–1166.
- [Vaughan, 1988] Vaughan, J. (1988). Multipactor. *IEEE Transactions on Electron Devices*, 35(7):1172–1180.
- [Vera-Castejón et al., 2014] Vera-Castejón, P., Correas-Serrano, D., Quesada-Pereira, F. D., Hinojosa, J., and Álvarez-Melcón, A. (2014). A novel low-pass filter based on rounded posts designed by an alternative full-wave analysis technique. *IEEE Transactions on Microwave Theory and Techniques*, 62(10):2300–2307.
- [Wang et al., 2005] Wang, W., Lu, Y., Fu, J. S., and Xiong, Y. Z. (2005). Particle swarm optimization and finite-element based approach for microwave filter design. *IEEE Transactions on Magnetics*, 41(5):1800–1803.

- [Wu et al., 2003] Wu, K., Deslandes, D., and Cassivi, Y. (2003). The substrate integrated circuits - a new concept for high-frequency electronics and optoelectronics. In *6th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Service, 2003. TELSIKS 2003*. IEEE.
- [Yang et al., 2014] Yang, T., Ho, K., and Rebeiz, G. M. (2014). Compact self-shielded 2–3 GHz high-Q coaxial fixed and tunable filters. *IEEE Transactions on Microwave Theory and Techniques*, 62(12):3370–3379.
- [Yang et al., 2013] Yang, X., Gao, Y., Wu, J., Beguhn, S., Nan, T., Zhou, Z., Liu, M., and Sun, N. X. (2013). Dual H- and E-field tunable multiferroic bandpass filter at Ku-band using partially magnetized spinel ferrites. *IEEE Transactions on Magnetics*, 49(11):5485–5488.
- [Yang et al., 2018] Yang, Z., Psychogiou, D., and Peroulis, D. (2018). Design and optimization of tunable silicon-integrated evanescent-mode bandpass filters. *IEEE Transactions on Microwave Theory and Techniques*, 66(4):1790–1803.
- [Yassini et al., 2012] Yassini, B., Yu, M., and Keats, B. (2012). A Ka-band fully tunable cavity filter. *IEEE Transactions on Microwave Theory and Techniques*, 60(12):4002–4012.
- [You et al., 2016] You, B., Lu, S., Chen, L., and Gu, Q. J. (2016). A half-mode substrate-integrated filter with tunable center frequency and reconfigurable bandwidth. *IEEE Microwave and Wireless Components Letters*, 26(3):189–191.
- [Yuceer, 2016] Yuceer, M. (2016). A reconfigurable microwave combline filter. *IEEE Transactions on Circuits and Systems II: Express Briefs*, 63(1):84–88.