Analysis and Realization of *L*-Band Dielectric Resonator Microwave Filters

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Abstract—**The development of space communications necessitates that microwave devices used in satellite systems have good temperature and vibration characteristics, low weight and size. High power *L*-band dielectric resonator (DR) microwave filters which can solve these problems have been developed and are reported herein. The electromagnetic and electrical parameters of different microwave dielectric resonator structures have been computed by means of the two dimensional and three dimensional finite element method (FEM), which can be applied both for free and forced oscillation systems. In this paper, we propose, design and evaluate the response of a new type of filter using rectangular dielectric resonators excited in their TM_{110} mode.**

Introduction

High power band pass filters are required in mobile communications systems operating in L frequency band. At these frequencies, the thin-invar empty cavities filters excited in their TE_{111} modes have very large dimensions ($\phi = 150$ mm, $L = 200$ mm) and cannot realize the low mass required in space specifications.

The use of dielectric materials which combine high Q, good thermal stability and high dielectric constant permit reducing the size and the weight of the microwave devices.

Some solutions have been already proposed to realize L-band dielectric resonator filter: in particular dual mode dielectric resonator longitudinal (Fig. 1) [1], [2] or planar (Fig. 2) [3], [4] structures. Another solution consists of using half (Fig. 3) [5] or quarter (Fig. 4) [5], [6] dielectric resonators filters excited on their $TE_{01\delta}$ modes. Unfortunately dual mode dielectric resonator filters are not capable of handling higher power levels. The last two structures allow obtaining a good dissipation of temperature but the unloaded Q_u factor of each resonator is low.

To solve the problem of both filter size and temperature dissipation, we propose to use the TM_{010} cylindrical mode [7] or TM_{110} rectangular mode DR inserted into a metallic cylindrical enclosure, in which the DR axis and that of the metallic waveguide are perpendicular (Fig. 5).

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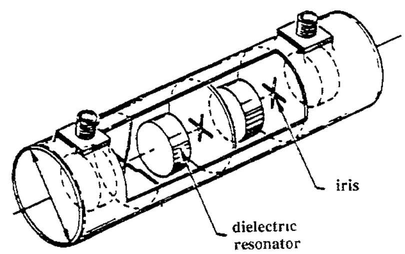

Fig. 1. Dual mode dielectric resonator filter (longitudinal)

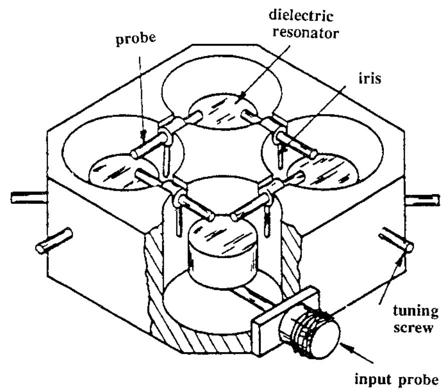

Fig. 2. Dual mode dielectric resonator filter (planar).

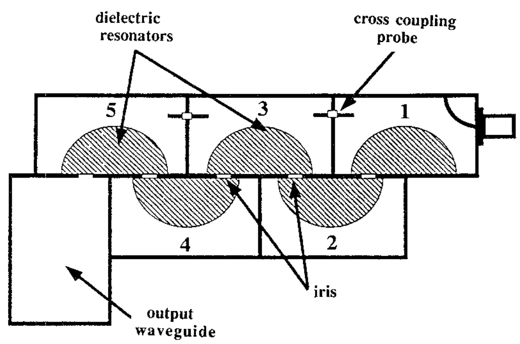

Fig. 3. 5th degree elliptic filter with $TE_{01\delta}$ image resonator half disc.

The analysis of these types of filters necessitates use of a 3-D free or forced oscillations finite element method [8], [9] having first order mixed elements. In effect, this method permits obtaining:

resonant frequency F_0;
magnetic field lines;

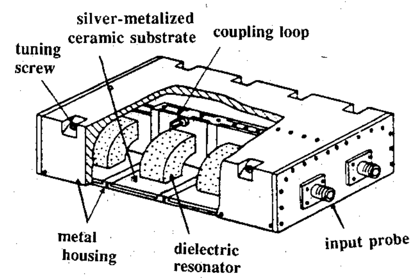

Fig. 4. Quarter-cut $TE_{01\delta}$ resonator filter.

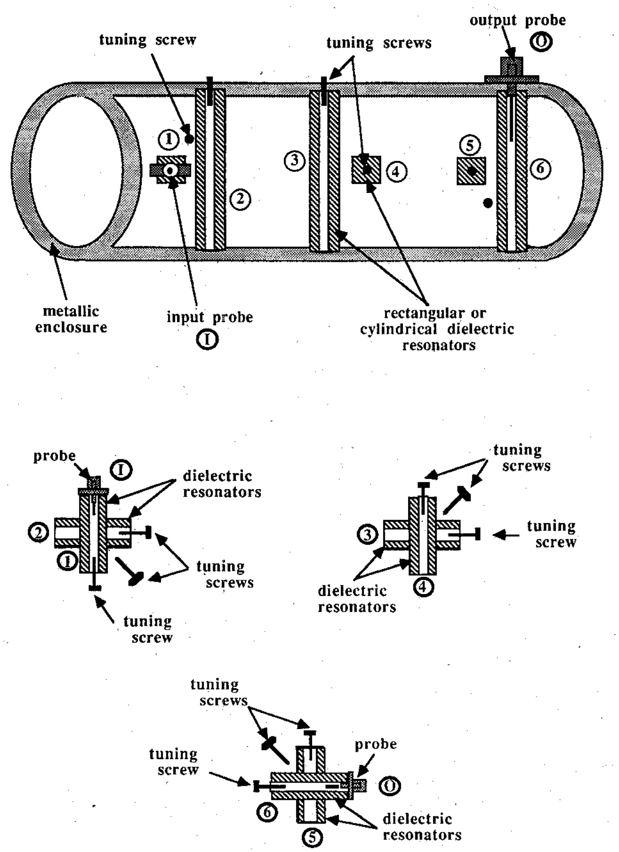

Fig. 5. Configuration of a 6 poles short circuited DR filter.

electric field lines;
stored energies;
unloaded Q_u factor;
coupling coefficients between two adjacent DR;
coupling of dielectric resonators with the external ports.

Mixed Elements

The rigorous evaluation of electromagnetic and electrical parameters of a shielded dielectric resonator cannot be done by a straightforward analytic method, but only by a numerical solution implemented on a computer, usually a large mainframe computer.

Considering electric field $\vec{E}$ and magnetic field $\vec{H}$ as distribution vectors in Maxwell's equations [10], and solving these latter, we obtain the general propagation equation (1) available to compute any free and forced oscillation microwave problems:

$$\iiint_V \left(\frac{1}{p}\{\vec{\nabla}x\}\vec{\psi}\right)\cdot(\{\vec{\nabla}x\}\vec{\phi})\,dV - k_0^2\iiint_V q\vec{\psi}\cdot\vec{\phi}\,dV = -j\omega u\sum_{k=1}^{n}\iint_{S_k}\vec{J_{S_k}}\cdot\vec{\phi}\,dS_k, \tag{1}$$

with

$$k_0^2 = \omega^2\epsilon_0\mu_0,$$

where

V: volume of the structure
S_k: surface of the access plane ($k = 1, \cdots, n$)
ω: angular frequency of the fields.

We denote $\{\vec{\nabla}x\}$ as the rotational operator applied to a function.

*For magnetic $\vec{H}$ field formulation:

$\vec{\psi} = \vec{H}$
$\vec{\phi} = \vec{\phi_m}$: test function normal to magnetic surfaces
$p = \epsilon_r$ (permittivity)
$q = \mu_r$ (permeability)
$u = \epsilon_0$
$\vec{J_{S_k}} = \vec{J_{mS_k}}$: magnetic surface currents at the access planes S_k.

*For electric $\vec{E}$ field formulation:

$\vec{\psi} = \vec{E}$
$\vec{\phi} = \vec{\phi_e}$: test function normal to electric surfaces
$p = \mu_r$
$q = \epsilon_r$
$u = \mu_0$
$\vec{J_{S_k}} = \vec{J_{eS_k}}$: electric surface currents at the access planes S_k.

Equation (1), which describes the electromagnetic behavior of the structure, is discretized and solved using the finite element program library Modulef [11].

The method consists of dividing the studied structure into triangular (2-D) or tetrahedral (3-D) subdomains. The unknown function $\vec{\psi}$ and the test function $\vec{\phi}$ are approximated by first order mixed elements [12], [13]:

$$\vec{\psi} = \sum_i \psi_i\vec{\omega}_i \qquad 1 \le i \le N \tag{2}$$

where

N is the number of modes
$\vec{\omega}_i$: basis vectorial functions

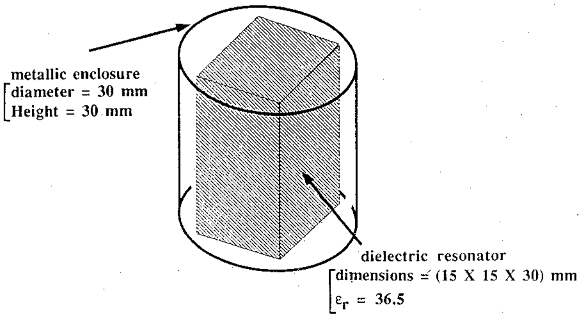

Fig. 6. Rectangular DR into metallic cylindrical enclosures.

TABLE I
RESONANT FREQUENCIES OF THE FIRST MODES FOR THE RESONATOR OF FIG. 6 AND THE COMPUTING TIME

	Lagrange Elements				Mixed Elements		
	H Formulation				H Formulation		
Order of the polynomial	2				1		
Number of the points	75				75		
Number of the nodes	405				660		
	Frequency (GHz)	Mode	Computing Time		Frequency (GHz)	Mode	Computing Time
In the 1.4 GHz to 2 GHz band	1.4444	spurious	38 nm	In the 1.4 GHz to 3 GHz band	1.4577	TM_{110}	11 mn
	1.4642	TM_{110}			2.6493	hybrid	
	1.6542	spurious			2.6589	hybrid	
	1.7424	spurious			2.7264	hybrid	
	1.9206	spurious			2.7334	hybrid	
	1.9419	spurious					
	1.9573	spurious					

ψ_i: complex numbers which define the vectorial function $\vec{\psi}$. These numbers are not function values but weighted circulations of the function along the edges.

In the case where the polynomial functions are of degree $k = 1$, the degrees of freedom ψ_i associated with the ith edge defined by Nedelec [13] are given by

$$\psi_i = \int_a (\vec{\psi} \cdot \vec{\tau}) N_i \, dl \tag{3}$$

where

$\vec{\tau}$: unit tangential vector at the edge

N_i: basis functions, for each node i the value of N_i is 1 at the node i, and 0 at the other ones

$\vec{\psi}$: unknown function which is approximated by Lagrange elements in the expression (3):

$$\vec{\psi} = \sum_i \vec{\psi}_i N_i.$$

To prove the advantages or this new formulation over the old one (which used Lagrange polynomials), we have evaluated the resonant frequencies of the first modes for a rectangular DR inserted into a metallic cylindrical enclosure (Fig. 6).

In Table I we can show that the mixed elements present many advantages; in particular they permit:

elimination of spurious responses;
reduction of computing time.

Electromagnetic Parameters

The finite element method, using mixed elements, has been applied to the evaluation of the resonant frequency F_0, Q_u factor and electric and magnetic field of parallelepiped and cylindrical short-circuited DR. As an example of the results obtained with 3-D FEM, we present respectively in Figs. 7 and 8, the magnetic field of TM_{010} cylindrical DR mode and TM_{110} rectangular DR mode. In the Fig. 9, we give the variation of the Q_u factor as a function of height of the DR; for the filter design the height is taken equal to 45 mm which is the best compromise to obtain a small size and a large unloaded Q_u factor.

We conclude that a good realization of a 1.6 GHz microwave DR filter, will result from using the following dimensions:

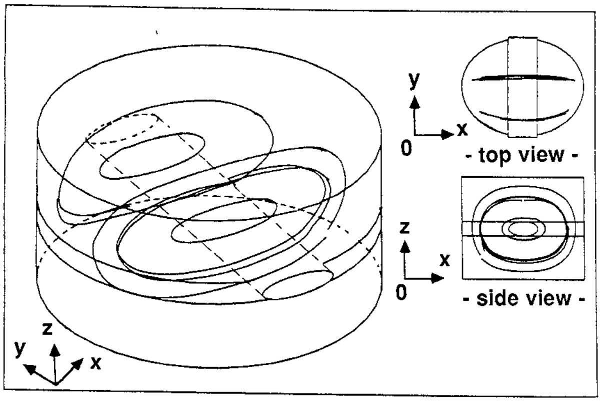

Fig. 7. Magnetic field lines of TM_{010} cylindrical DR mode.

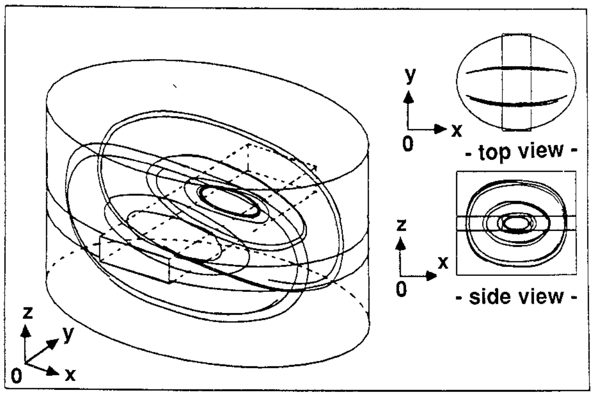

Fig. 8. Magnetic field lines of TM_{110} rectangular DR mode.

Cylindrical DR

diameter of DR
= 12 mm
height of DR
= 45 mm
permittivity of DR ϵ_r
= 36.5
diameter of cavity
= 44 mm
height of cavity
= 75 mm

Rectangular DR

dimensions of DR
= (10.5 × 10.5 × 45) mm
permittivity of DR ϵ_r
= 36.5
diameter of cavity
= 44 mm
height of cavity
= 75 mm

Electrical Parameters

To reduce size and weight of microwave filters it is necessary to study new coupling structures. The theoretical and experimental coupling coefficients between two rectangular DR as a function of the inter-stages distance are shown in Fig. 10. Using the symmetry of the structure, the coupling coefficient k [14] is accurately calculated from the following relation:

$$k = \frac{f_{oe}^2 - f_{om}^2}{f_{oe}^2 + f_{om}^2} \qquad (4)$$

in which f_{oe} and f_{om} are the resonant frequencies corresponding respectively to even and odd modes.

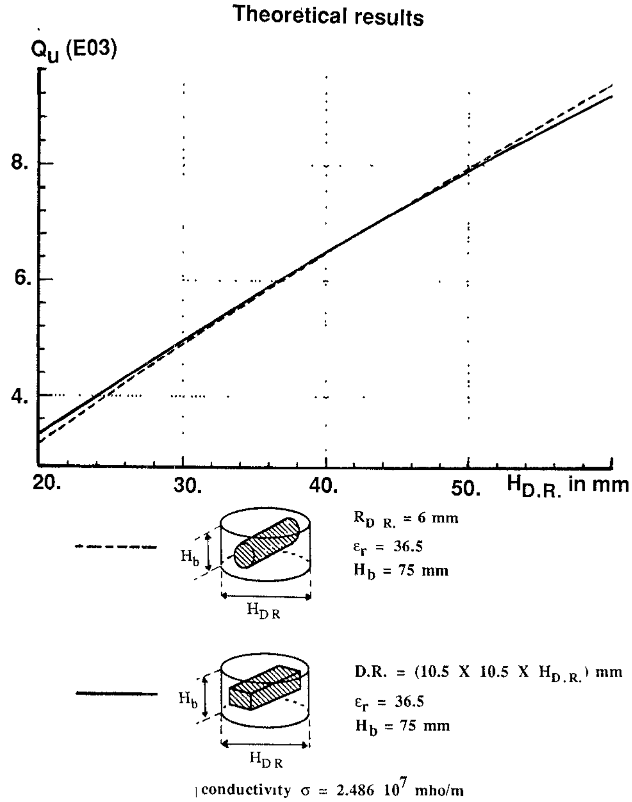

Fig. 9. Migration path of unloaded Q_u factor as a function of the height of DR.

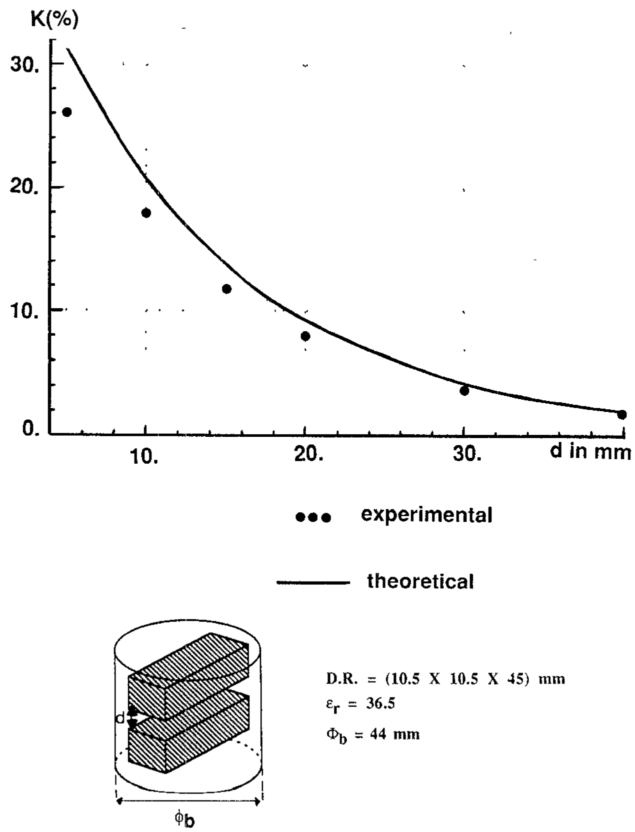

Fig. 10. Variation of the inter-resonators coupling as a function of the inter-stages distance (experimental and theoretical).

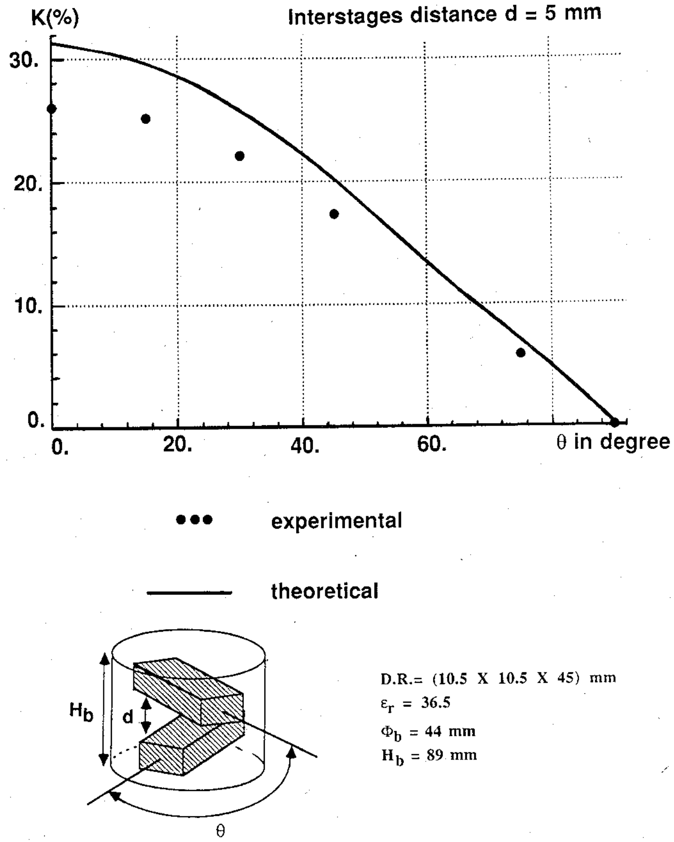

Fig. 11. Variation of the inter-resonators coupling as a function as the angle θ between two DR.

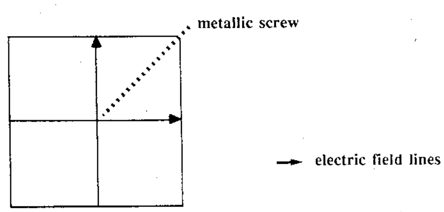

Fig. 12. Structure with its metallic screw configuration of the right angle.

But this configuration does not permit obtaining small coupling coefficient together with small size of microwave filters. So a new structure in which DR and waveguide axis are perpendicular is considered. This permits obtaining small coupling coefficients for small distances between the DR. This configuration can be exploited to reduce size and weight of DR structures, in particular by considering:

- the inter-resonator coupling as a function as the angle θ between two DR;
- the inter-resonator coupling between two orthogonal DR with metallic screw;
- the inter-resonator coupling with a rectangular iris between them.

Let it be noted that in these cases we cannot use the symmetry of the structure for the coupling coefficient computation and so it is necessary to take into account the whole structure.

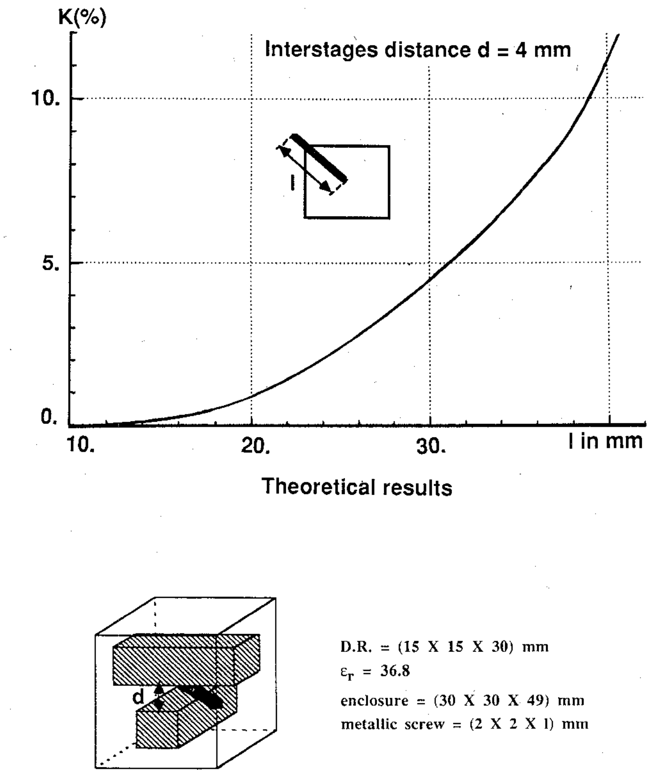

Fig. 13. Inter-resonators coupling between two orthogonal DR with metallic screw.

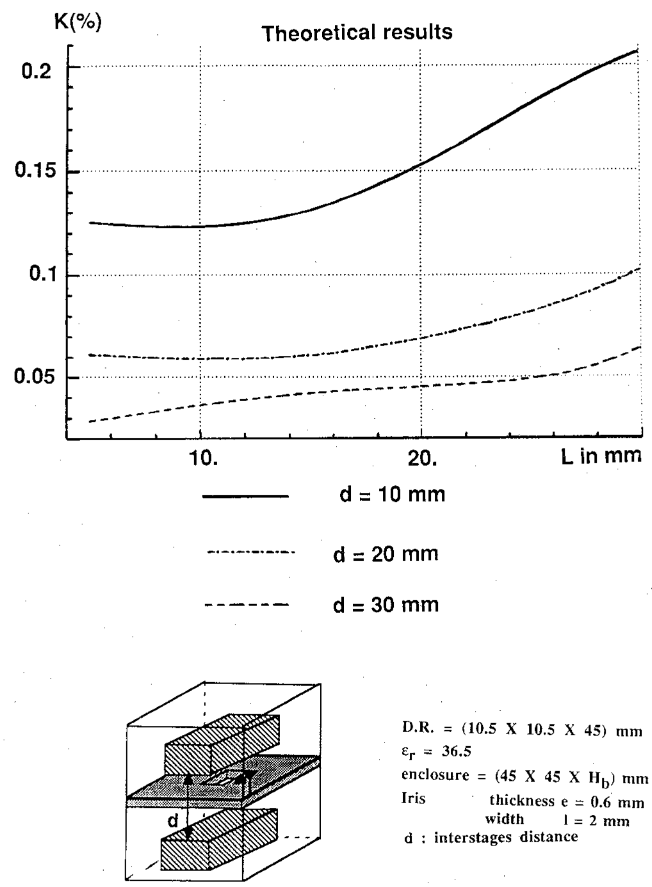

Fig. 14. Inter-resonators coupling with a rectangular iris between them.

TABLE II
PARAMETERS OF THE BAND PASS ELLIPTIC FILTER

Center frequency			1.50 GHz			
Bandwidth			25 MHz			
Normalized input and output impedance			1.142			
Center frequency:						
transmission losses			0.3 dB			
reflexion losses			>21 dB			
Normalized coupling matrix	0	0.9026	0	−0.03	0	0
	0.9026	0	0.6506	0	0	0
	0	0.6506	0	0.5918	0	−0.03
	−0.03	0	0.5918	0	0.6506	0
	0	0	0	0.6506	0	0.9026
	0	0	−0.03	0	0.9026	0

The coupling coefficient k is also given by the following relation:

$$k = \frac{F_{02} - F_{01}}{F_0}.$$

For a mounted pair of DR the transmission coefficient as a function of frequency will show two maximum, respectively at F_{01} and F_{02} and one minimum at F_0 of the resonator overcoupled.

In Fig. 11, we give comparison between calculated and measured coupling coefficients versus the angle θ for an interstage distance d equal to 5 mm. With such a configuration we can have a small coupling coefficient value for a small distance between two DR according to angle θ.

When $\theta = 90°$, the coupling coefficient is equal to zero. However, two orthogonal DR can be coupled by using a metallic tuning screw (Fig. 12) which destroys the orthogonality between the TM_{110} mode excited in the two resonators. For example presented here we can note that when the length of the metallic screw in the enclosure is larger than 10 mm (Fig. 13), an inter-resonator-coupling appears.

We can also note that this configuration permits obtaining negative coupling coefficient value and this result will be very useful in realizing elliptic band-pass filter functions [15], [3].

Another solution to reduce size of DR structures consists of using an iris (Fig. 14), but added metallic components increase losses in microwave filter responses.

The theoretical results obtained present a good agreement with the experiments ones, and can lead to the realization of compact high power band-pass elliptic filters which use TM_{110} parallelepiped DR modes.

ELLIPTIC FILTER

Filter Performances

An experimental 6 poles elliptic prototype has been designed and constructed with rectangular DR. Parameters of the filter are given in Table II. A schematic representation of the coupling matrix coefficients is shown in Fig. 15. The coupling coefficients have positive values.

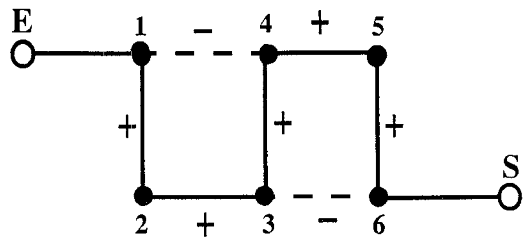

Fig. 15. Scheme representation of the coupling matrix coefficients.

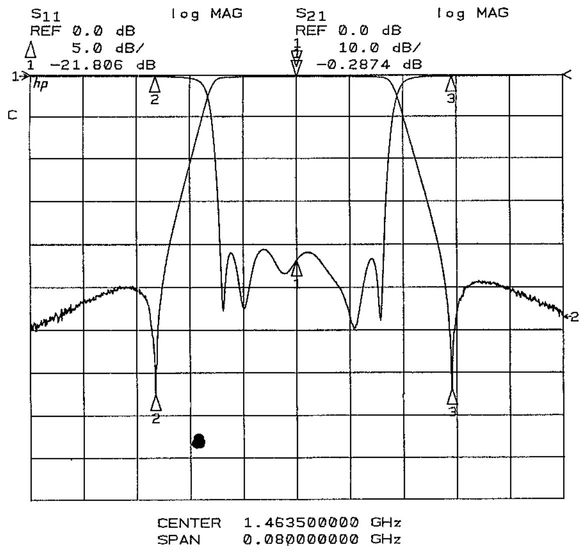

Fig. 16. Frequency responses of the experimental filter.

The mounting of the dielectric resonators inside the cavities is based up on the differential expansion phenomena between the aluminum housing and the dielectric resonators. The metallic cavity is heated to a high temperature to provide a sufficient increase in diameter of the cavity to put the dielectric resonator inside it by means of a particular mounting tool.

Fig. 16 shows the response of the 25 MHz bandwidth experimental filter. The in-band insertion losses are approximately equal to 0.3 dB and the return loss measured is 25 dB. The unloaded quality factor is equal to 6500. Fig. 17 shows the position of the nearest spurious modes.

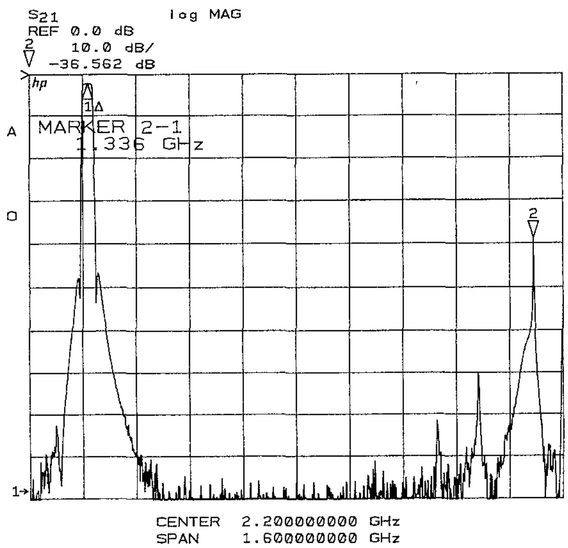

Fig. 17. Spurious responses.

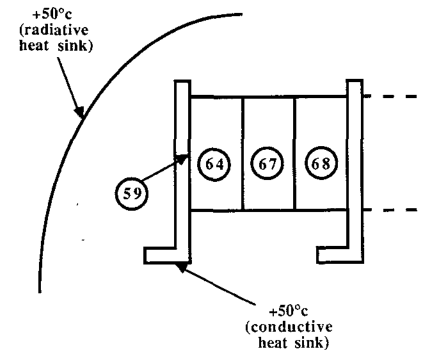

Fig. 18. Cavity temperatures map on half a filter.

Thermal Design

Thermal interfaces:

dissipated power: P_d
input power: 60 W
electrical losses: 0.4 dB.

So $P_d = 5.3$ W; for this kind of filter, the finite element method has permit to know the following dissipated power distribution:

60% in the dielectric resonator
40% in the internal cavity.

Reference temperature is taken on the equipment mounting feet. Thermal analysis is performed considering conductive and radiative coupling with the payload environment.

Main objectives for the thermal design are

- to minimize temperature excursion of filters in the designing cases (cold case low RF level/hot case high RF level);
- to minimize thermal gradients along filters;
- to minimize hot spot temperatures.

The thermal analysis performed with specific software on electrical enclosure, gives the results of the Fig. 18, for a reference temperature of +50°C.

Conclusion

In this paper, we have applied a new 3-D finite element formulation to compute electromagnetic and electrical parameters of microwave DR devices.

The comparison carried out between theoretical and experimental results shows the use fulness of this formulation.

More accurate results should be obtained by using second order mixed elements.

Theoretical and experimental results obtained are applied to realize an L-band dielectric resonator elliptic filter, which can be integrated in future satellite microwave devices.

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