Synthesis and Design of Asymmetrical Dual-band Bandpass Filters Based on Equivalent Network Simplification

P. Lenoir [1], S. Bila [1], F. Seyfert [2], D. Baillargeat [1], S. Verdeyme [1]

(1) IRCOM, 123 Avenue Albert Thomas, 87060 Limoges, France
(2) INRIA, 2004 Route des Lucioles, 06902 Sophia Antipolis, France

Abstract – **Although the synthesis of symmetrical dual-band bandpass filters has been studied in the literature, very little seems to be known about the general asymmetrical case. In this paper, a procedure for the synthesis of asymmetrical dual-band bandpass filters implemented with in-line dual-mode cavities is proposed. The latter starts with the determination of a filtering characteristic fulfilling the electrical specifications in the spectral domain by means of an optimization scheme. As to the coupling topology, the in-line architecture that is intended for the hardware realisation, as well as the asymmetrical nature of the response, lead naturally to the choice of an extended box topology or its variants. It was recently shown that these topologies possess the property of multiple solutions, meaning that the related coupling matrix synthesis problem admits several solutions. On one hand, this multiplicity offers some additional flexibility to the designer, but on the other hand, working with multiple solutions may lead to ambiguities during the tuning process. Our procedure takes the best of both worlds : after an exhaustive synthesis step yielding a list of equivalent coupling matrices, simplifications of the original topology are obtained by cancelling some particular couplings. The locations of cancelled couplings are chosen so as to preserve the electrical response and to provide some important hardware simplifications. It is also shown that the resulting simplified coupling topology no longer has the multiple solution property, solving ambiguity problems that might occur during the tuning step. The procedure is demonstrated on two examples of 11-pole asymmetrical dual-band bandpass filters. A numerical model driven by a coupling matrix extraction process is used to determine the geometrical dimensions. Finally an experimental prototype is built in order to validate our design approach.**

Index Terms – **microwave filters, circuit synthesis, coupling matrix, dual band filters**

I. INTRODUCTION

The demand for advanced filtering functions has considerably increased with the development of space telecommunications. For example, in satellite communication systems, highly selective transfer functions with self-equalised group delays are required for IMUX channels. Another emerging application in this domain is the design of dual-band bandpass filters used to transmit non-contiguous channels to the same geographical region through one beam [1]. In this case, a single high power amplifier (HPA) can be used together with the dual-band bandpass filter, dramatically simplifying the system architecture.

An approach for implementing such a circuit consists of designing two classical single-band bandpass filters, one for each passband. Their input/output ports are then connected together through waveguide junctions. However, this approach leads to a complex design procedure since waveguide junctions and filters have to be optimised together to comply with the mechanical constraints. Indeed, each channel must have the same length and input/output waveguide ports must have the same orientation. Another approach consists of designing a single circuit realising the dual-band characteristic. This straightforward approach requires the synthesis of an advanced filtering function but makes the hardware implementation easier since a classical filter architecture can be used.

Narrow-band filters, dedicated to space applications, are generally implemented using cavities or resonators since they offer better performances in terms of losses and power handling. For reducing mass and volume, these resonant elements are often excited on dual-modes. Furthermore, non-adjacent couplings between resonant elements are generally required in order to add transmission zeroes to the transfer function for improving the selectivity and/or flattening the group delay. A practical way to implement a filter with dual-mode cavities, while permitting non-adjacent couplings, is the in-line architecture which consists of connecting dual-mode cavities in a row. The latter architecture is used in this work for implementing asymmetrical dual-band bandpass filters.

The synthesis of a microwave filter starts with the selection of a transfer function that fulfils the electrical specifications. For single-band filtering characteristics, quasi-elliptic polynomial functions given by explicit formulas are widely employed [2]. For symmetrical dual-band characteristics of even degree and with an even number of transmission zeroes, the latter formulas may be adapted by means of frequency transformations [3].

This is no longer the case for more general situations and gets designers to use direct optimisation methods [1], [4]-[6]. In this work, a local optimisation method is applied where the starting point is computed from the quasi-elliptic synthesis of each individual channel.

In a second step, an equivalent lumped element network is synthesized in order to realize the selected transfer function. The equivalent network is characterised by its coupling topology specifying the distribution of zero and non-zero couplings between resonators. The latter has obviously to be consistent with the filter architecture, as well as with the transfer function it is supposed to realise. The main difficulty comes here from the conflicting requirements of the design, on one hand, a topology able to realise several asymmetric transmission zeroes, while on the other hand, the latter should remain simple enough to admit a classical hardware implementation. In our case, the hardware implementation is an in-line dual-mode cavity architecture ideally with no cross couplings. This will lead to considering the use of extended box coupling topologies [7], as well as their extensions in order to synthesise some extra transmission zeroes. For these topologies, it was shown recently that the number of solutions to the coupling matrix synthesis problem is high [8]. The current work demonstrates how to use this extra flexibility by providing rules to be applied in order to obtain significant simplifications of the original topology while keeping the electrical response nearly unchanged. The latter translates into hardware simplifications like transformation of cross-irises into single-arm irises or realignment of irises with respect to the cavities. Such a simplification approach has been employed in [9] for implementing a symmetrical filter architecture while realising an asymmetrical single-band transfer function.

Finally, the proposed approach is also consistent with numerical modelling techniques which are often used by designers. These methods are used along with a coupling matrix extraction algorithm [10]-[13]: this allows driving the tuning process, so as to converge towards a device implementing the ideal coupling matrix. Nevertheless, when working with topologies which admit multiple solutions, the latter coupling matrix extraction step returns a list of several equivalent coupling matrices. This leaves the difficult task of choosing the right one to the designer and thus represents the main drawback of topologies with multiple solutions. The current work shows how hardware simplifications obtained in the preceding step of our procedure solve the addressed ambiguity and allow to use a tuning process based on a well-posed coupling matrix extraction problem.

The approach is illustrated by the design of two asymmetrical dual-band filters at Ka-band. In the first part, an 11-pole dual-band bandpass filter with 4 transmission zeroes is designed. The synthesis procedure is presented from the determination of the characteristic function, up to the simplified network construction. The latter network allows the simplification of cross irises into single-arm irises without any notable effect on the circuit behaviour. A numerical model and an experimental model are also investigated in order to demonstrate the efficiency of the proposed approach. In the second part, the approach is repeated, synthesising an 11-pole dual-band bandpass filter with 5 zeroes. Here the simplified network allows re-aligning of all the distributed elements for an easier hardware implementation of the in-line dual-mode cavity filter.

II. DESIGN OF AN 11 POLE-4 ZERO DUAL-BAND BANDPASS FILTER

The electrical specifications of the dual-band bandpass filter to be designed are :

- a first passband centred at 18.362-GHz, with a 39-MHz bandwidth,
- a second passband centred at 18.508-GHz, with a 78.5-MHz bandwidth.

A 20-dB return loss in each passband and a 10-dB insertion loss in the intermediate stopband are required. An insertion loss greater than 25-dB is also desired in the lower and upper stopbands.

Starting from these electrical specifications, the transfer and reflection functions of the dual-band filter are calculated.

A. Characteristic function selection

The characteristic function $D(s)$ can be written as a polynomial rational function :

$$D(s) = \frac{\prod_{i=1}^{N}(s - s_{ri})}{\prod_{i=1}^{Nz}(s - s_{pi})} = \frac{R(s)}{P(s)}, \tag{1}$$

where s_{ri} and s_{pi} are respectively the normalised reflection and transmission zeroes,
N is the number of reflection zeroes, i.e. the order of the filtering function and
Nz is the number of transmission zeroes.

The modulus of the transfer function admits the following simple expression in terms of its characteristic function :

$$\left|S_{21}(s)\right|^2 = \frac{1}{1+\varepsilon^2 \left|D(s)\right|^2} \tag{2}$$

where ε is an adjustable real parameter.

Applying the procedure described in [6], the initial dual-band characteristic function is constructed from two single-band functions. In our case, the lower passband is realised with a 5th order quasi-elliptic function and the upper passband is realised with a 6th order quasi-elliptic function. Each single-band characteristic function presents 2 transmission zeroes for improving the selectivity in the stopbands.

As a result, the dual-band characteristic function is initialised with 4 transmission zeroes and 11 reflection zeroes. The initial reflection and transmission zeroes are then slightly retuned in order to improve the transmission feature (2) within the two passbands, leading to the following normalised values :

$$\begin{aligned}
&s_{p1} = -j\ 1.063,\quad s_{p2} = -j\ 0.565,\quad s_{p3} = j\ 0.205,\quad s_{p4} = j\ 1.063,\\
&s_{r1} = -j\ 0.996,\quad s_{r2} = -j\ 0.943,\quad s_{r3} = -j\ 0.823,\quad s_{r4} = -j\ 0.682,\quad s_{r5} = -j\ 0.610,\\
&s_{r6} = j\ 0.261,\quad s_{r7} = j\ 0.354,\quad s_{r8} = j\ 0.557,\quad s_{r9} = j\ 0.774,\quad s_{r10} = j\ 0.922,\quad s_{r11} = j\ 0.985
\end{aligned} \tag{3}$$

with $\varepsilon = 47$.

The transfer and reflection functions corresponding to these values are presented in figure 1.

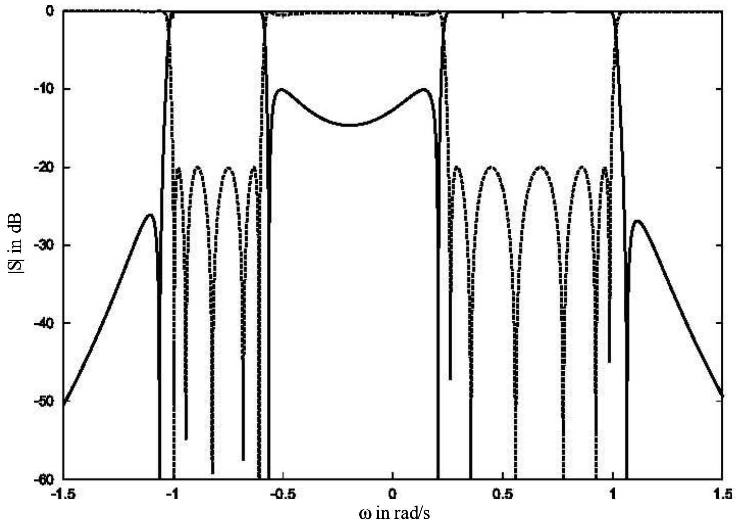

Fig.1 – Ideal transfer (—) and reflection (---) functions

B. Exact synthesis

An equivalent lumped element network that realises the previous transfer function is now synthesised. The related coupling topology has to be adapted to the exact synthesis of the desired asymmetrical characteristic, as well as compatible with the in-line dual-mode cavity architecture. The extended box topology presented in Fig. 2 meets our requirements since this coupling topology allows to realise any asymmetrical transfer function of order 11 with 4 transmission zeroes.

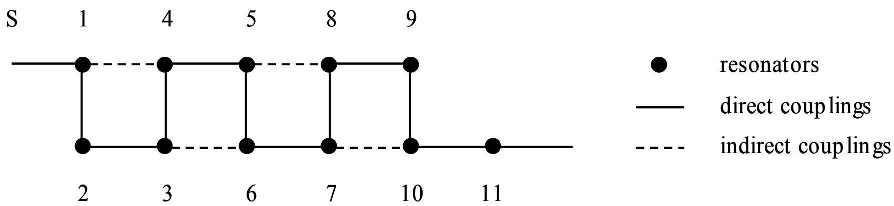

Fig.2 – Coupling topology realising 11 pole-4 zero transfer functions

The exhaustive synthesis method presented in [8] is applied in order to determine all the coupling matrices that correspond to the previous coupling topology. The method is based on computations that solve exhaustively an algebraic system of equations related to the synthesis problem.

Generically, an 11-pole 4-zero transfer function can be realised in 384 manners with the above coupling topology. This theoretical number, called the reduced order, is the number of complex solutions to the synthesis problem and does not depend on the considered filtering characteristic (only on the coupling topology).

Nevertheless, the number of real solutions, *i.e.* the only ones of physical interest, depends on the numerical values of the characteristic polynomials. For our particular filtering characteristic, 66 real solutions are found.
Theoretically, any solution among these 66 ones could be chosen to design the filter. But our goal is now to simplify the initial coupling topology by cancelling one or several couplings without severely affecting the electrical response. To this end, some rules have to be observed when seeking to simplify the coupling topology:

- the number of couplings in the shortest coupling path, between source and load, has to be preserved in order to keep the number of transmission zeroes constant,
- couplings corresponding to irises are cancelled in priority, since couplings realised with screws can hardly be completely set to zero in practice because of remaining residual couplings.

The latter rule indicates that starting from the coupling diagram in Fig. 3, our simplification will apply only to cancel one or several horizontal couplings. The shortest path rule imposes some conditions on cancellable horizontal couplings. For example, if coupling M_{14} (between resonators 1 and 4) is cancelled, all the couplings in the inferior path (M_{23}, M_{36}, M_{67}, M_{710}) needs to remain non zero.

C. Approximate synthesis with a simplified network

Following latter rules, solutions with low cross couplings M_{14} and M_{58} are explored. A good candidate, out of all 66 matrices is the following matrix :

$R_{in} = R_{out} = 0.563$

$$\begin{pmatrix}
-0.115 & 0.773 & 0 & -0.079 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0.773 & 0.142 & 0.521 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0.521 & -0.221 & 0.514 & 0 & -0.264 & 0 & 0 & 0 & 0 & 0 \\
-0.079 & 0 & 0.514 & 0.296 & 0.697 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0.697 & -0.305 & 0.425 & 0 & 0.094 & 0 & 0 & 0 \\
0 & 0 & -0.264 & 0 & 0.425 & 0.318 & 0.507 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0.507 & -0.323 & 0.189 & 0 & 0.390 & 0 \\
0 & 0 & 0 & 0 & 0.094 & 0 & 0.189 & 0.482 & 0.314 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.314 & -0.101 & 0.267 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0.390 & 0 & 0.267 & 0.144 & 0.777 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.777 & -0.115
\end{pmatrix} \qquad (4)$$

The cancellation of M_{14} and M_{58} modifies the resulting transfer and reflection functions as shown in Fig. 3. But compensating this effect with the remaining couplings, the original transfer function is almost recovered as shown in Fig. 4. The coupling topology is the one presented in Fig. 5 and the final coupling matrix is then :

$R_{in} = R_{out} = 0.\ 571$

$$\begin{pmatrix}
-0.106 & 0.781 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0.781 & 0.147 & 0.476 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0.476 & -0.243 & 0.478 & 0 & -0.267 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0.478 & 0.315 & 0.725 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0.725 & -0.313 & 0.422 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & -0.267 & 0 & 0.422 & 0.320 & 0.555 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0.555 & -0.297 & 0.149 & 0 & 0.399 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0.149 & 0.472 & 0.300 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.300 & -0.113 & 0.245 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0.399 & 0 & 0.245 & 0.130 & 0.787 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.787 & -0.106
\end{pmatrix} \qquad (5)$$

Our approach leads to the simplification of two cross coupling irises into single-arm irises. Moreover, the tuning process through coupling matrix extraction will be simplified since the reduced order of the simplified topology is one as computed with methods detailed in [8]. More precisely, the previous simplified topology does not allow to realise all the transfer functions of 11^{th} order with 4 transmission zeroes but, when the latter is a realisable one, there corresponds only one coupling matrix. In other words, the original transfer function in Fig. 1 can not be realised exactly with the simplified coupling topology, but the approximate transfer function in Fig. 4 can only be realised with the coupling matrix in (5).

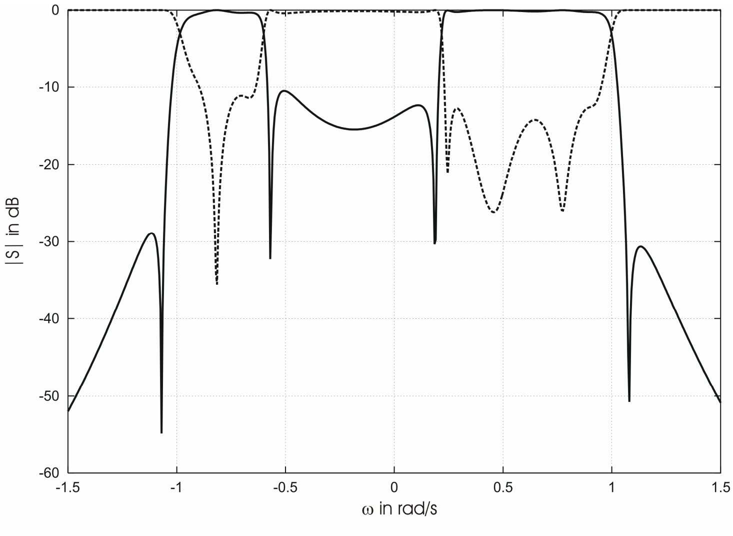

Fig.3. – Transfer (—) and reflection (---) functions when couplings M_{14} and M_{58} are neglected (no compensation)

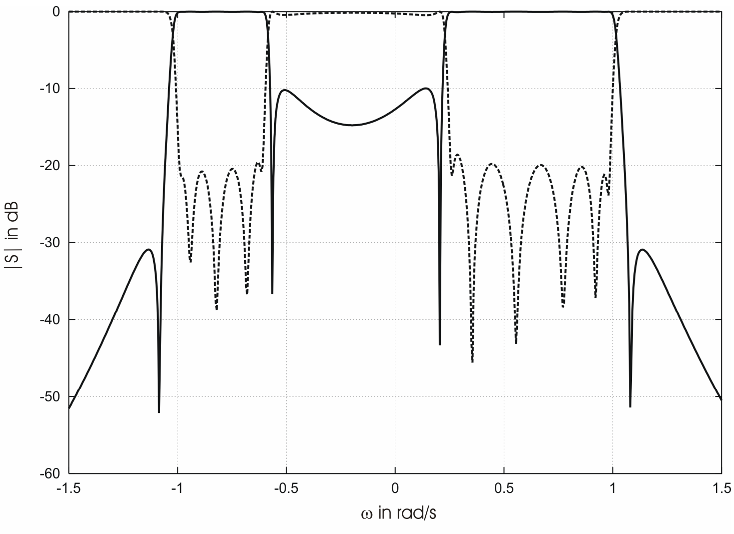

Fig.4. – Approximate transfer (—) and reflection (---) functions (neglected couplings M_{14} and M_{58} are compensated)

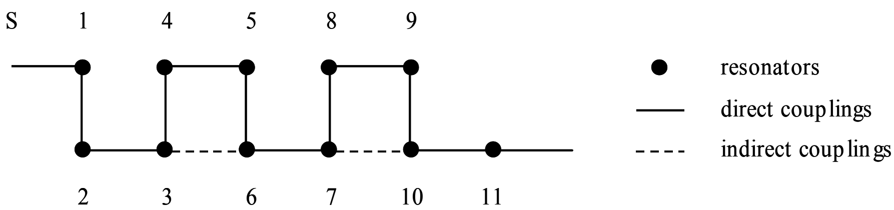

Fig.5 – Simplified coupling topology leading to the 11 pole-4 zero approximate transfer function in figure 4

D. Electromagnetic optimisation

The electromagnetic model of the in-line dual-mode cavity filter is presented in Fig. 6.

The filter is designed for the TE_{113} mode, applying the electromagnetic optimisation procedure presented in [10]. Each electromagnetic analysis is followed by a coupling matrix extraction step yielding some corrections on the modelled geometrical dimensions.

The transfer and reflection functions obtained from the electromagnetic model are given in Fig. 7. The numerical model behaviour is slightly different from the ideal one since parasitic couplings between resonant elements have been compensated [14].

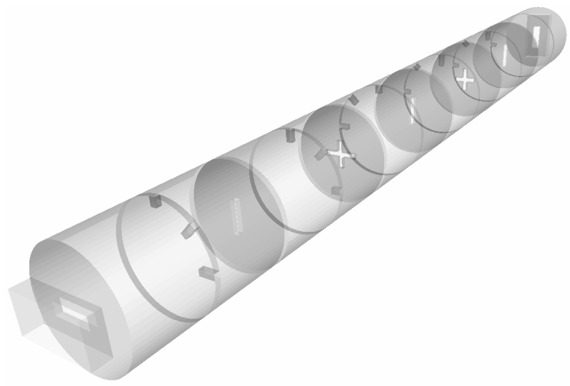

Fig.6 – Electromagnetic model of the 11-pole 4-zero asymmetrical dual-band bandpass filter

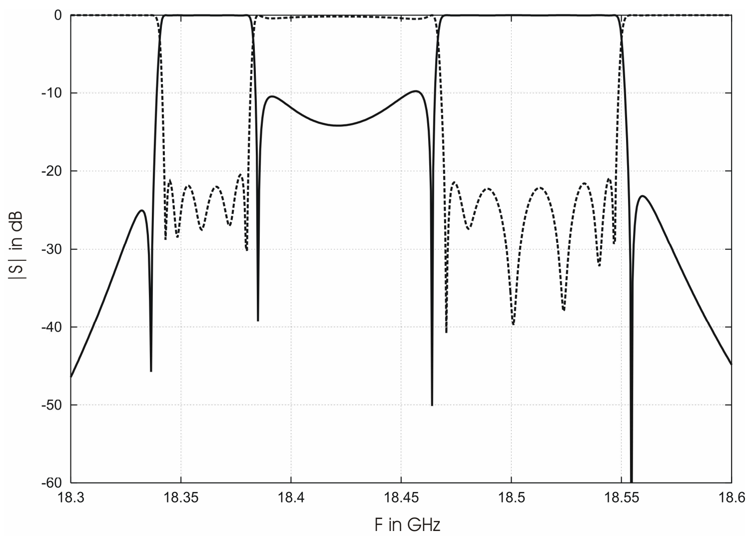

Fig.7 – Transfer (—) and reflection (---) functions obtained with the electromagnetic model

E. Measurements

The filter has been built and tested. A picture of the realised prototype is presented in Fig. 8.

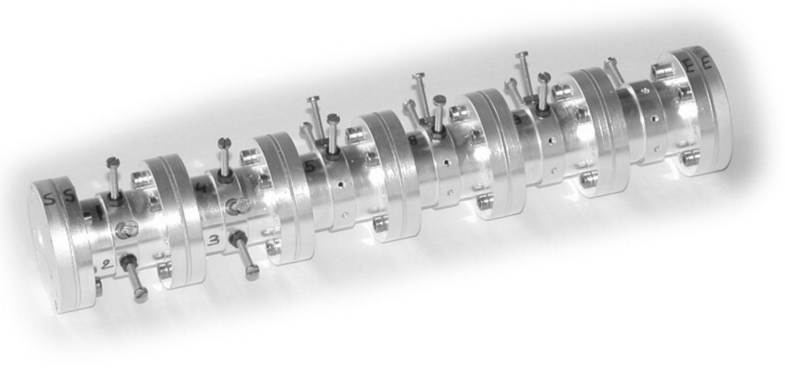

Fig.8 – Picture of the realised dual-band bandpass filter

The measured transfer and reflection functions are presented in Fig. 9. A good agreement is achieved between theory and measurement, validating our design approach with simplified coupling topologies.

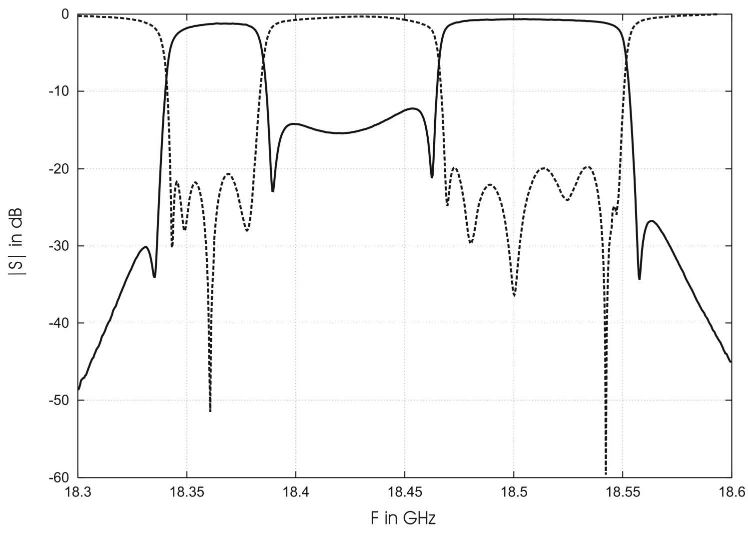

Fig.9 – Experimental transfer (—) and reflection (---) functions

III. SYNTHESIS OF AN 11 POLE-5 ZERO DUAL-BAND BANDPASS FILTER

In order to deepen the intermediate stopband, a 5th transmission zero is added to the transfer function. The insertion loss in the intermediate stopband is then specified to be 25-dB. The previous synthesis procedure is repeated.

A. Characteristic function selection

Applying the same method, the following transmission and reflection zeroes are computed :

$$\begin{array}{llllll} s_{p1} = -j\ 1.065, & s_{p2} = -j\ 0.515, & s_{p3} = -j\ 0.153, & s_{p4} = j\ 0.161, & s_{p5} = j\ 1.067, & \\ s_{r1} = -j\ 0.996, & s_{r2} = -j\ 0.940, & s_{r3} = -j\ 0.818, & s_{r4} = -j\ 0.683, & s_{r5} = -j\ 0.610, & \\ s_{r6} = j\ 0.261, & s_{r7} = j\ 0.346, & s_{r8} = j\ 0.532, & s_{r9} = j\ 0.753, & s_{r10} = j\ 0.915, & s_{r11} = j\ 0.985 \end{array} \quad (6)$$

with $\varepsilon = 40$.

The transfer and reflection functions corresponding to these values are presented in Fig. 10.

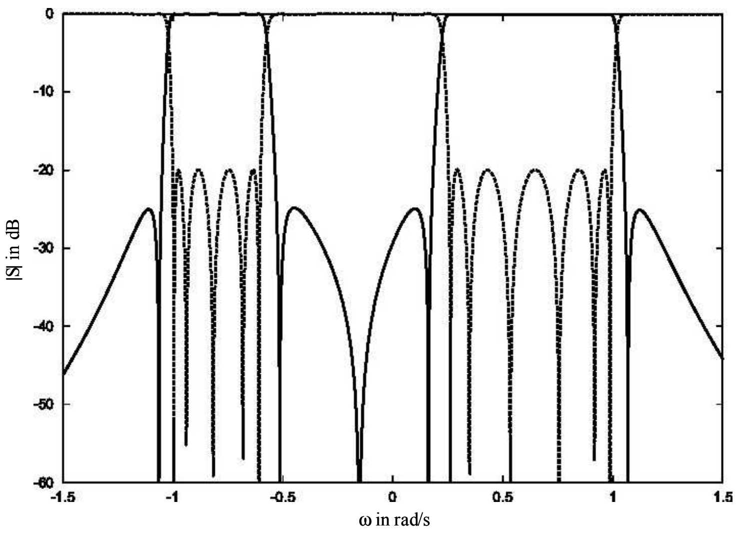

Fig.10. – Ideal transfer (—) and reflection (---) functions

B. Exact synthesis

In order to realise and adjust an additional transmission zero, the extended-box topology needs to be modified while keeping in mind the following facts :

- one degree of freedom, *i.e.* one extra coupling, must be added to the actual extended box topology in order to enable to adjust the position of the new transmission zero,
- the shortest path between resonators 1 and 11 in the new coupling topology must be of length 5 to satisfy the minimum path rule.

The latter requirements are met by the coupling topology in Fig. 11 by adding cross-coupling M_{911} to the original extended box (Fig. 2). This coupling topology allows to realise any transfer function of order 11 with 5 transmission zeroes.

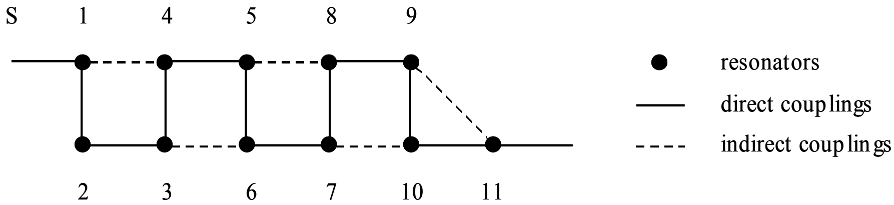

Fig.11 – Coupling topology realising 11 pole-5 zero transfer functions

The reduced order of this topology is found to be 963, and applying an exhaustive synthesis from the selected transfer function leads to 81 real coupling matrices.
The cross-coupling M_{911} is the main problem of the above coupling topology as an angle is necessary between the last coupling iris and the last cavity in order to realise it. Our approach will therefore focus on simplifications of the coupling topology that allow a realisation with aligned irises and cavities.
Obviously, the rules given for the first example still hold valid. However, one can note that now the shortest coupling path is unique. Consequently, none of the following couplings M_{14}, M_{45}, M_{58}, M_{89} and M_{911} can be cancelled.
In order to recover an aligned architecture, a possible way is to suppress coupling M_{1011}. Indeed the latter cancellation will lead to the simplified coupling topology presented in Fig. 12,

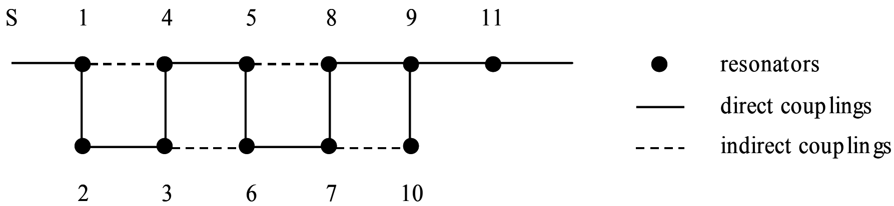

Fig.12 – Simplified coupling topology proposed for realising the 11 pole-5 zero approximate transfer function

C. Approximate synthesis with a simplified network

Following our approach based on approximate synthesis, solutions with low cross coupling M_{1011} are investigated. The following coupling matrix is then a good candidate :

$R_{in} = R_{out} = 0.563$

$$\begin{pmatrix}
-0.116 & 0.352 & 0 & -0.694 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0.352 & -0.588 & 0.309 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0.309 & -0.612 & 0.046 & 0 & 0.262 & 0 & 0 & 0 & 0 & 0 \\
-0.694 & 0 & 0.046 & 0.332 & 0.308 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0.308 & -0.038 & 0.653 & 0 & -0.311 & 0 & 0 & 0 \\
0 & 0 & 0.262 & 0 & 0.653 & 0.166 & 0.174 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0.174 & 0.813 & 0.074 & 0 & 0.152 & 0 \\
0 & 0 & 0 & 0 & -0.311 & 0 & 0.074 & -0.461 & 0.430 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.430 & 0.146 & 0.198 & 0.778 \\
0 & 0 & 0 & 0 & 0 & 0 & 0.152 & 0 & 0.198 & 0.726 & -0.004 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.778 & -0.004 & -0.116
\end{pmatrix} \qquad (7)$$

Considering this late matrix, couplings M_{34} and M_{78} have also weak values and should also be cancelled applying the approximate synthesis; but since these couplings are implemented with coupling screws, they are preserved in the simplified coupling topology.
Neglecting the coupling M_{1011}, the resulting transfer and reflection functions are only slightly modified, and by compensating with the remaining couplings, the original transfer function is recovered as shown in Fig. 13. The final coupling matrix, which is consistent with the coupling topology presented in figure 12, is then :

$R_{in} = R_{out} = 0.563$

$$\begin{pmatrix} -0.116 & 0.352 & 0 & -0.694 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0.352 & -0.588 & 0.309 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0.309 & -0.612 & 0.046 & 0 & 0.262 & 0 & 0 & 0 & 0 & 0 \\ -0.694 & 0 & 0.046 & 0.331 & 0.307 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0.307 & -0.038 & 0.653 & 0 & -0.311 & 0 & 0 & 0 \\ 0 & 0 & 0.262 & 0 & 0.653 & 0.166 & 0.174 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0.174 & 0.813 & 0.075 & 0 & 0.152 & 0 \\ 0 & 0 & 0 & 0 & -0.311 & 0 & 0.075 & -0.460 & 0.430 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.430 & 0.145 & 0.196 & 0.778 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0.152 & 0 & 0.196 & 0.727 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0.778 & 0 & -0.116 \end{pmatrix} \quad (8)$$

Applying this approach, the hardware implementation is highly simplified since all the irises and cavities are aligned. Moreover, the tuning process through coupling matrix extraction will be also simplified since here again the reduced order of our simplified topology is found to be one.

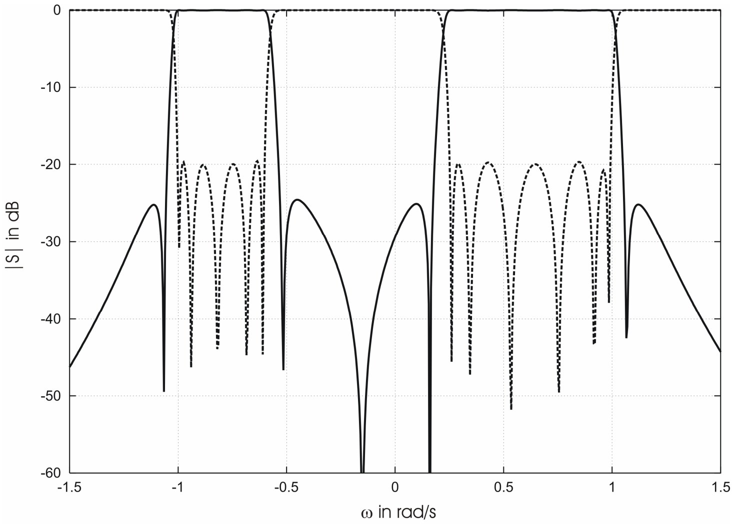

Fig.13. – Approximate transfer (—) and reflection (---) functions (neglected coupling M_{1011} is compensated)

IV. Conclusion

This paper presents an approach based on equivalent network simplification for the synthesis and the design of asymmetrical dual-band bandpass filters implemented with in-line dual-mode cavities. The first step involves an exact and exhaustive synthesis yielding a list of equivalent coupling matrices which are consistent with the extended box coupling topology or variations of it. In a second step, the proposed approach takes advantage of the multiple solution property of these coupling topologies by providing some rules for selecting a coupling matrix to be used as the starting point for an approximate synthesis procedure. The approximate synthesis allows then some substantial simplifications of the initial coupling topology by cancelling one or several weak couplings between resonators. The simplified coupling topology makes the hardware implementation easier and also solves ambiguity problems that may occur during the tuning phase by restoring the well-posedness of the coupling matrix extraction step.

The proposed approach is applied to synthesise and to design two asymmetrical dual-bandpass filters implemented with in-line dual-mode cavities. When applied to an 11-pole 4-zero microwave filter, the proposed approach allows to replace two cross irises by single-arm irises when compared with an exact synthesis. A numerical model and an experimental prototype of this filter have been fabricated in order to validate the theoretical results. The approach is repeated with an 11-pole –5-zero microwave filter and the approximate synthesis allows to realign all the distributed elements with respect to an exact synthesis.

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