

Coupling Matrix Designing of a Cross-Coupled Resonator Waveguide Filter Based on FT and SSOA Techniques Using ANSYS-HFSS

Abdelhakim Boudkhil^{1,2}, Boualem Mansouri¹, Mohammed Chetioui^{1,2}, Mehdi Damou¹, Abdelhafid Lallam² and Nasreddin Benahmed²

¹Laboratory of Electronics, Advanced Signal Processing and Microwaves, Faculty of Technology, University of Saida - Dr. Moulay Tahar, Algeria

²Laboratory of Telecommunications, Faculty of Technology, University of Tlemcen - Abu Bakr Belkaid, Algeria

Corresponding author: A. Boudkhil (e-mail: boudkhil.abdelhakim@yahoo.fr) and M. Chetioui (e-mail: chetioui.mohammed@yahoo.fr).

ABSTRACT This paper presents an efficient hybrid method of filter modeling that combines a fast tuning step (FT) to a global optimization technique to design a new topology of a 4th order cross-coupled resonator bandpass microwave filter. The method can rapidly extract the coupling matrix (CM) from the electromagnetic (EM) simulated scattering parameters (S-parameters) of a rectangular waveguide resonator bandpass filter with losses that will be implemented in a computer-aided tuning (CAT) tool to develop a high performance microwave bandpass filter design simulated by successively adding a one resonator each time of a set of sequential tuning iterations within the ANSYS-HFSS (ANSYS-High Frequency Structure Simulator). This minimizes the time of designing of the microwave filter's structure as well as permits to build a higher waveguide filter's order providing a complex cross-coupling. A filter parameters' manual-coarse adjustment is finally achieved by implementing a step by step optimization algorithm (SSOA) tool based on the EM-simulator as a fast process for ideal refinement to the partial circuitual response.

INDEX TERMS Bandpass Filter, Microwaves, Coupling Matrix, Cross-Coupling, Tuning, EM-Simulation.

I. INTRODUCTION

THE synthesis of microwave filters affords an easy procedure to obtain the parameters of the equivalent circuits to meet the design requirements [1, 2]. The microwave filters with cross-coupled resonators, which gives a finite transmission zeros (TZs) [3], was broadly utilized to (1) enhance the near-to-band skirt selectivity; (2) reach in-band flat group delay; (3) separate the single-band into several pass-bands [4, 5]. Conversely, in comparison with the traditional direct-coupled resonator bandpass filter, the 4th cross-coupled resonator bandpass filter is extra complicated to be physically implemented, due to the interrelationship introduced by the cross-couplings [6].

A growing request for bandpass filter's design technique without a demand for excessive optimization [7] has been resulted. The procedure of the design timing, accuracy and complexity can be significantly enhanced by employing appropriate techniques that still less accurate specially for designing filters with cross-coupled responses [8, 9]. Filters stand on cross-coupled resonators, with real or complex transmission zeros, have been widely used to (i) enhance beside to band selectivity and (ii) attain in-band group delay linearity. The cross-coupled filter is more difficult to be physically implemented, because of the interactions

introduced by the cross-couplings than the conventional in-line resonator coupled filter.

Conventionally, the design techniques of direct-coupled filters have been adopted to excerpt the cross-coupled filter's dimensions [10, 11]. After having obtained the initial physical dimensions of each iris and cavity, an overall structure optimization is required so that to achieve the desired electromagnetic response that stills traditional and less efficient. The structure optimization becomes slow for a circuit containing a higher order and/or complex cross-coupled topology that disrupts the convergence of final results. The global optimization becomes then a more time-consuming technique and the final optimization may fail to converge to acceptable solutions because of the large amount of controlling parameters. Here, an efficient method is presented to overcome the suggested problem. Accordingly, this paper presents different steps of estimation to design microwave filters with the purpose of minimizing the encountered issues. It is important to note that precise analysis of microwave filters requires resolving Maxwell's equations in structure, thus assessing objective functions in optimization techniques may take some hours. In this research study, ANSYS-HFSS based on accurate design approach is chosen for simulation as an accurate EM

simulator to help determining the required physical initial parameters of a cross-coupled bandpass microwave filter. The design procedures enable to consider assignation of cross-coupling and provide precise desired dimensions basing on effective step by step optimization algorithm without needing for a global optimization process that allows to reduce the design timing. The proposed technique may find useful applications to design resonators based on cross-coupled filters [12].

II. Design Procedures

The design of the bandpass filter can be executed following two ways: Synthesizing the polynomial transfer functions of the filters and extracting the filter coupling matrix by optimization.

A. POLYNOMIALS DESIGN AND COUPLING MATRIX CIRCUIT

From general of any microwave filter, transmission and reflection coefficients of a two-port lossless filter network are given by characteristic polynomials $P(s)$, $F(s)$ and $E(s)$ as in [12]:

$$S_{21}(s) = \frac{P(s)/\varepsilon}{E(s)} \quad (1)$$

And

$$S_{11}(s) = \frac{F(s)}{E(s)} \quad (2)$$

Where the constant ε is used to normalize the highest polynomials' function degrees to 1. The characteristics polynomials relate on each other because the energy conversion:

$$|E(s)|^2 = |F(s)|^2 + \frac{|P(s)|^2}{\varepsilon^2} \quad (3)$$

The circuit model already investigated presents a bandpass prototype as displayed in Figure (1). It is made from a cascade of lumped-element resonator series inter-coupled by transformers. Each resonator consists of a capacitor of $1 F$ in series to a self-inductance of $1 H$ for each loop.

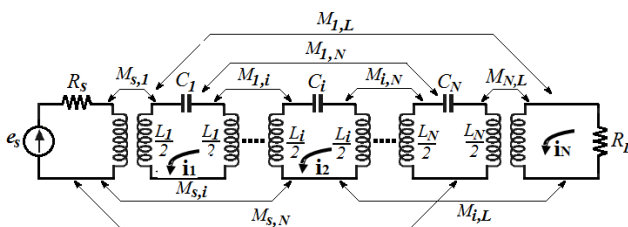


FIGURE 1. Multi-coupled series-resonator bandpass prototype.

The m matrix of a general two-port network contains one input port (S), n coupled resonators and one output port (L) and can

$$M = \begin{bmatrix} & S & 1 & 2 & 3 & \cdots & N & L \\ S & m_{S,S} & m_{S,1} & m_{S,2} & m_{S,3} & \cdots & m_{S,N} & m_{S,L} \\ 1 & m_{1,S} & m_{1,1} & m_{1,2} & m_{1,3} & \cdots & m_{1,N} & m_{1,L} \\ 2 & m_{2,S} & m_{2,1} & m_{2,2} & m_{2,3} & \cdots & m_{2,N} & m_{2,L} \\ 3 & m_{3,S} & m_{3,1} & m_{3,2} & m_{3,3} & \cdots & m_{3,N} & m_{3,L} \\ \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ N & m_{N,S} & m_{N,1} & m_{N,2} & m_{N,3} & \cdots & m_{N,N} & m_{N,L} \\ L & m_{L,S} & m_{L,1} & m_{L,2} & m_{L,3} & \cdots & m_{L,N} & m_{L,L} \end{bmatrix}$$

be expressed as followed :

It presents a symmetric matrix with respect to the diagonal that contains coupling coefficients between external and internal resonator ports (m_{S1} , m_{S2}). In addition, it is possible to accommodate direct couplings between all resonators as m_{ij} which present the coupling between i resonator and j resonator. However, m_{ii} presents the resonator offset frequency. The $(N+2)$ coupling matrix, can be calculated from F , P and E polynomials using *Cameron* method [4]: the matrix entries present the de-normalized coupling coefficients given by $M_{ij} = m_{ij} * FBW$, where FBW is the filter fractional bandwidth.

B. THE STEP BY STEP OPTIMIZATION ALGORITHM

The step by step optimization algorithm technique based on an electromagnetic simulator (ANSYS-HFSS) alters at each iteration a circuit parameter. It simulates only one resonator at first time and by finishing the first resonator tuning, another resonator is added, then the circuit is tuned or optimized again. More resonators are successively tuned at each step. For each step, a new coupling matrix is required for the tuning. The optimizing process works more efficiently and generates more reliable solutions as limited number of physical dimensions needs to optimize in each step. The filter topology tuning process is built successively by adding one resonator at a time.

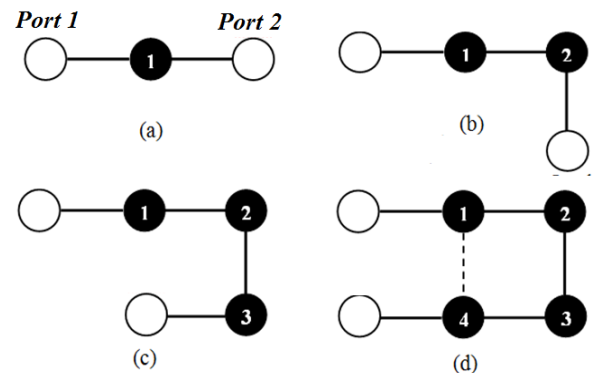


FIGURE 2. The four stages (a-d) of filter topology tuning process.

In each design step a part of the structure is selected and the electromagnetic simulation is compared with the coupling matrix ideal response correspondent to that part. Selected steps can be depicted by means of the suitable diagram of Figure (2), in which circles represent cavities and M_{ij} and describe the coupling (inverters in the circuit model) between resonators. The main point of the step by step optimization algorithm technique is to evaluate the S -parameters (S_{11}/S_{21}) in each stage and apply the responses as the objective ones for the physical optimizing. From the coupling coefficient of a K -inverter between resonators i and j is given by the following equations:

- For input and output couplings:

$$K_{S,1} = \frac{M_{S,1}}{\sqrt{\alpha}}, K_{4,L} = \frac{M_{4,L}}{\sqrt{\alpha}} \quad (4)$$

- For the inter-resonator couplings:

$$K_{i,j} = \frac{M_{i,j}}{\alpha} \quad (5)$$

Where M_{ij} are the coupling values taken from the prototype coupling matrix. K_{ij} are the RF inverter values:

$$\alpha = \left[\lambda_{g1} + \lambda_{g2} / n\pi (\lambda_{g1} - \lambda_{g2}) \right] \quad (6)$$

Where λ_{g1} and λ_{g2} are the guide wavelengths at the lower and upper band-edge frequencies, respectively. n is the number of half-wavelengths of the waveguide resonator cavity. In this case, the fractional guided wavelength bandwidth is given by:

$$FBW_{\lambda} = (\lambda_{g1} - \lambda_{g2}) / \lambda_{g0} \quad (7)$$

Where λ_{g0} is the guide wavelength at the centre frequency. In this technique of designing, the middle stage S parameter responses are evaluated from their corresponding coupling coefficients, and act as the objective responses for the tuning; to plot the desired responses at each step, the inner coupling coefficient needs to be converted into de-normalized external quality factor (Q_e). For instance, in step 1, Q_e should be calculated from M_{12} . After expressing both the external quality factor Q_{ei} and internal coupling coefficients $M_{(i,i+1)}$ using inverter value K [11], the relationship between $M_{(i,i+1)}$ and Q_{ei} can be found as:

$$M_{i,i+1}^2 \cdot Q_{ei} = \frac{1}{\frac{n\pi}{2} \left(\frac{\lambda_g}{\lambda_0} \right)^2 FBW} \quad (8)$$

Where λ_g is the guided wavelength of the resonant frequency and λ_0 is the free-space wavelength. The coupling coefficients between external ports and inner resonators can be calculated by:

$$M_{i,L}^2 = \frac{1}{Q_{ei}} \quad (9)$$

i refers to the resonator number connecting to the output ports (L). Substituting (8) into (9) provides:

$$M_{i,L} = \sqrt{\frac{n\pi}{2} FBW} \left(\frac{\lambda_0}{\lambda_g} \right) M_{i,i+1} \quad (10)$$

Where $M_{i,L}$ is the equivalent external coupling coefficient of the internal coupling iris.

III. Waveguide Filter Design Simulation

A cross-coupled filter has been developed in a $WR-90$ waveguide whose dimensions are: $a = 22.86 \text{ mm}$ and $b = 10.16 \text{ mm}$. The passband is of 300 MHz , the central frequency is of 11 GHz and the fraction bandwidth (FBW) is of 0.02727 according to the method described above. The required characteristics are set as follows: The coupling matrix of a 4th order filter has been synthesized. The two transmission zeros on both sides of the passband at normalized frequencies (considering a lowpass prototype) are $[-1.5, +1.5]$ and the return loss S_{11} is of 20 dB . In the first design step, lowpass prototype filter elements with the following coupling matrix are obtained as follows:

$$M = \begin{bmatrix} 0 & 0.0276 & 0 & 0 & 0 & 0 \\ 0.0276 & 0 & 0.0221 & 0 & -0.0096 & 0 \\ 0 & 0.0221 & 0 & 0.0228 & 0 & 0 \\ 0 & 0 & 0.0228 & 0 & 0.0221 & 0 \\ 0 & -0.0096 & 0 & 0.0021 & 0 & 0.0276 \\ 0 & 0 & 0 & 0 & 0.0276 & 0 \end{bmatrix}$$

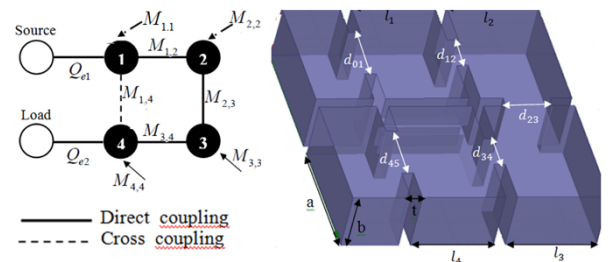


Figure (3) shows the developed 4th cross-coupled resonator bandpass microwave filter prototype where all irises have the same thickness $t = 2 \text{ mm}$.

When initial dimensions of each resonator and iris are got, ancient techniques consisted to optimize the whole topology so as to obtain the demanded results. If there is a high order of filter circuit or a complex coupled structure, the whole structure optimization becomes heavy and final results diverge. Therefore, this paper proposes a new fast technique which solves the issue standing on ANSYS-HFSS. Traditional techniques changed all filter physical dimensions for each optimization iteration. However, the proposed technique simulates only one filter cavity for the first step. When ending the first cavity adjustment, another cavity is appended to the optimization and the filter structure will be adjusted again. Further cavities are successively added to tuning step by step. The design can be split into four sub-step as presented in Figure (3). Initially, only 1st resonator dimensions (three or less parameters) are tuned. Therefore, a desired set of dimensions, whose corresponding responses match the objective ones, can be obtained within a short time. The response and obtained physical structure by the four sub-steps are displayed in Figure (4) to (10):

• Step one

The SSOA method starts with optimising resonator 1 of the crossed waveguide filter.

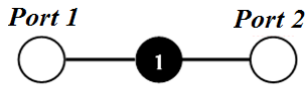


FIGURE 4. Schematic of the resonator 1 in step one.

Resonator 1 is coupled to two ports so the $(N+2)$ coupling matrix (with $N = 1$) is applied as:

$$M = \begin{bmatrix} 0 & 0.0276 & 0 \\ 0.0276 & 0 & 0.005699 \\ 0 & 0.005699 & 0 \end{bmatrix}$$

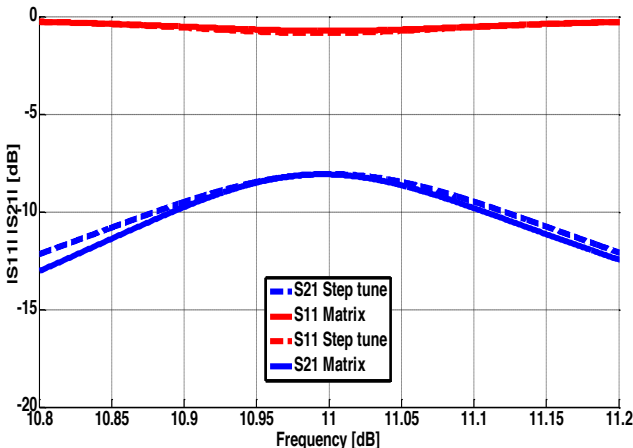


FIGURE 5. Step one: EM response of tuned results (solid lines) and target results (dashed lines).

According to Figure (3), Resonator 1 is coupled to Port 1 and Resonator 2. In Step One, Resonator 2 is replaced with a Port 2. Resonator 3 to 4 are removed. The schematic of the circuit in step one is illustrated in Figure 4. Figure 5 shows the EM response of tuned results displayed solid lines compared to EM response of the target results displayed in dashed lines. The EM response seems to be less effective and needs more optimization in the following step by adding at each step a filter cavity to be adjusted.

• Step two

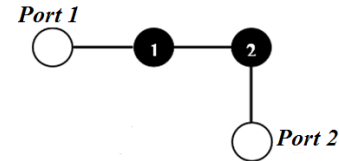


FIGURE 6. Schematic of the resonators 1 and 2 in step two.

$$M = \begin{bmatrix} 0 & 0.0276 & 0 & 0 \\ 0.0276 & 0 & 0.0221 & 0 \\ 0 & 0.0221 & 0 & 0.00587975 \\ 0 & 0 & 0.00587975 & 0 \end{bmatrix}$$

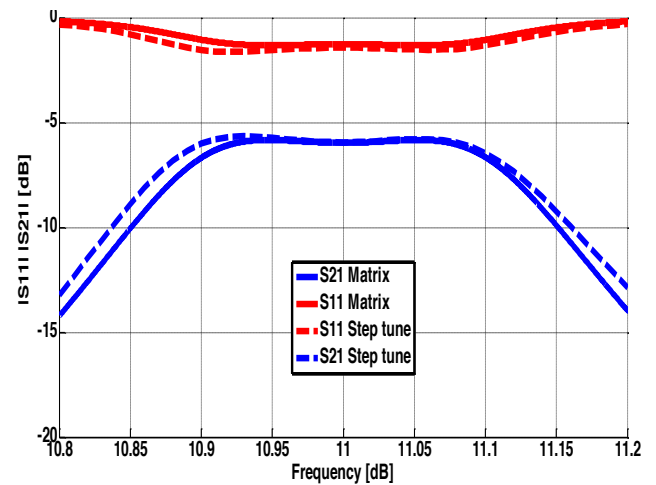


FIGURE 7. Step two: EM response of tuned results (solid lines) and target results (dashed lines).

• Step three

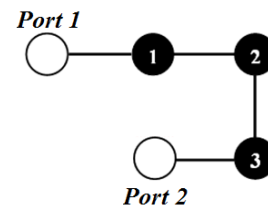


FIGURE 8. Schematic of the resonators 1, 2 and 3 in step three.

$$M = \begin{bmatrix} 0 & 0.0276 & 0 & 0 & 0 \\ 0.0276 & 0 & 0.0221 & 0 & -0.00244759 \\ 0 & 0.0221 & 0 & 0.0228 & 0 \\ 0 & 0 & 0.0228 & 0 & 0.005699 \\ 0 & -0.00244759 & 0 & 0.005699 & 0 \end{bmatrix}$$

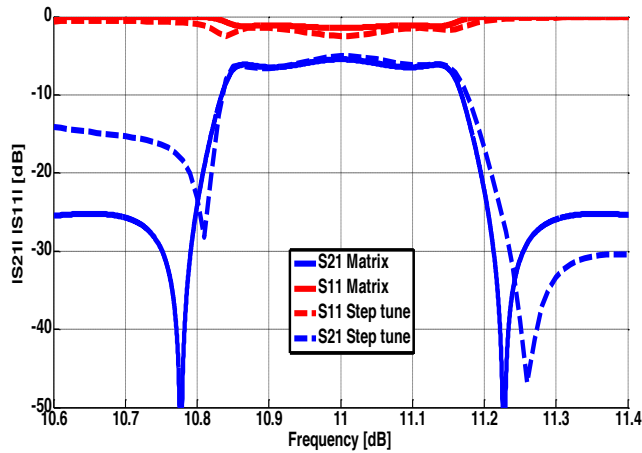


FIGURE 9. Step three: EM response of tuned results (solid lines) and target results (dashed lines).

• Step four

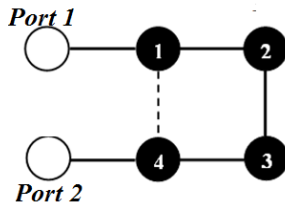


FIGURE 10. Schematic of all resonators (1, 2, 3 and 4) in final step (step four: completion of all stages).

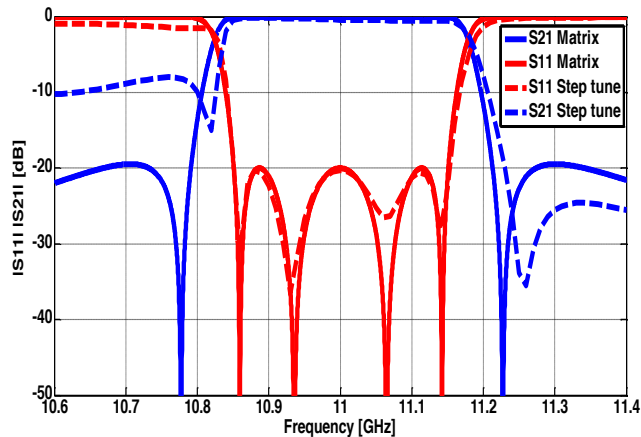


FIGURE 11. Step four: EM response of tuned results (solid lines) and target results (dashed lines).

The fourth and final step provides all dimensions of the optimized filter structure. The final EM response of the 4th cross-coupled resonator bandpass microwave is presented in Figure 11 that demonstrates good agreement of S_{11} and S_{21} in the whole filter passband due to the symmetrical transmission zeros pair.

The filter parameters' manual-coarse adjustment is finally achieved by implementing a step by step optimization algorithm as a fast process for ideal refinement to the partial circuital response. Final filter dimensions may be slightly altered to tune the response towards the desired one, as demonstrated in Tables 1 and 2. It can be observed that very small adjustments are required for the obtained dimensions in foregoing stages, to account for the influence from the subsequent resonators.

TABLE I. Bandpass filter iris dimensions for each step

Step	Dimensions of iris (mm)					
	$d_{0,1}$	$d_{1,2}$	$d_{2,3}$	$d_{3,4}$	$d_{4,1}$	$d_{4,5}$
1	10.100	6.200	-	-	-	-
2	10.205	6.399	8.000	-	-	-
3	10.205	6.378	8.000	6.510	0.700	-
4	10.205	6.378	8.000	6.380	0.720	10.208

TABLE II. Bandpass filter resonators' length for each step

Step	Length of resonators (mm)			
	l_1	l_2	l_3	l_4
1	14.430	-	-	-
2	14.285	15.795	-	-
3	14.385	15.800	15.800	-
4	14.385	15.800	15.800	14.385

IV. CONCLUSION

This paper describes a new process of filter designing based on mechanical dimension calculations using a fast tuning (FT) and step by step optimization algorithm (SSOA) techniques to develop a cross-coupled resonator bandpass microwave filter [13]. The final structure of the filter is developed step by step by adding to the simulated design a one resonator at each time. The dimensions of the resonator are set around the target mid-stage response as wished. Accordingly, a 4th order bandpass filter topology with a symmetrical transmission zeros pair has been successfully designed and attested using the proposed method that surpasses the necessity to employ a global optimization technique for dimensions and carries on to a significant decreasing in the design procedure timing. Furthermore, such a method permits to build a high waveguide filter's order with a complex cross-coupling and finds useful applications in the microwave field ranging from 10.7 GHz to 11.3 GHz.

REFERENCES

- [1] G. Macchiarella and S. Tamiazzo, "Synthesis of star-junction multiplexers," *IEEE Trans. Microw. Theory Techn.*, vol. 58, no. 12, pp.3732–3741, Dec. 2010.
- [2] G. Macchiarella, S. Tamiazzo "Novel approach to the synthesis of microwave duplexers" *IEEE Trans. Microwave Theory Tech.*, vol. 54, no.12, pp. 4281-4290, Dec. 2006.
- [3] G. Macchiarella and S. Tamiazzo, "Design of triplexer combiners for base stations of mobile communications," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Anaheim, CA, May 2010, pp. 429–432.
- [4] S. Bastioli and R. V. Snyder, "Design, modelling, and manufacturing of extremely selective waveguide filters using a multi-port optimization technique," in *Proc. IEEE MTT-S Int. Conf. Numer. Electromagn. Multiphys. Modeling Optim. (NEMO)*, Seville, Spain, May 2017, pp. 338–340.
- [5] D. Swanson and G. Macchiarella, "Microwave filter design by synthesis and optimization," *IEEE Microw. Mag.*, vol. 8, no. 2, pp. 55–69, Apr. 2007.
- [6] S. Bastioli and R. V. Snyder, "Evanescent mode filters using stronglycoupled resonator pairs," in *IEEE MTT-S Int. Microw.Symp.Dig.*, Seattle, WA, USA, Jun. 2013, pp. 1–3.
- [7] S. Bastioli, R. V. Snyder, and P. Jovic, "High power in-line pseudoelliptic evanescent mode filter using series lumped capacitors," in *Proc. 41st Eur. Microw. Conf.*, Oct. 2011, pp. 87-90.
- [8] Z. Xu, J. Guo, C. Qian, and W. Dou, "A novel quasi-elliptic waveguide transmit reject filter for ku-band vsat transceivers," *Progress in Electromagnetics Research*, vol. 117, pp. 393–407, 2011.
- [9] J.-T. Kuo, S.-C. Tang, and S.-H. Lin, "Quasi-elliptic function bandpass filter with upper stopband extension and high rejection level using crosscoupled stepped-impedance resonators," *Progress in Electromagnetics Research*, vol. 114, pp. 395–405, 2011.
- [10] T.F. Skaik, M.J. Lancaster and F. Huang, "Synthesis of multiple output coupled resonator circuits using coupling matrix optimisation," *IET Microwave Antenna Propag*, vol. 5, pp. 1081–1088, 2011.
- [11] R.J. Cameron, C.M. Kudsia, and R.R. Mansour, *Microwave filters for Communication Systems*, Wiley, Hoboken, NJ, 2016.
- [12] J.S. Hong and M.J. Lancaster, *Microwave Filters for RF/Microwave Applications*, Wiley, New York, NY, 2011.
- [13] A. Boudkhil, M. Chetoui, F. Bouasria, M.Damou, B. Bouras and N. Benahmed, "Diagnosis of coupled resonator waveguide filter using tuning step optimization technique," in *National Conference of Telecommunications and its Applications (CNTA21)*, Ain Tmouchent, Algeria, Dec 20-21, 2021.