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SIW/HMSIW Bandpass Filters for Ka Frequency Band Serving Wireless Applications

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***Abstract*—This paper describes a novel Substrate Integrated Waveguide (SIW) bandpass filter using *Chebyshev* approximation and Half Mode Substrate Integrated Waveguide (HMSIW)modeling technique. The developed 3rd order filter structure uses an inductive iris and an inductive post station in a way it resonates in Ka frequency band serving wireless applications. The paper presents in details steps of the filter design formed by specific analytical equations to extract its different synthesizable parameters including coupling matrix, quality factor and initial geometric dimensions. The ideal frequency response of the filter is determined from an equivalent circuit that uses localized elements developed by AWR Microwave Software. High Frequency Structure Simulator (HFSS) is then employed to model the proposed filter structure and optimize its initial parameters until meeting the target specifications initially fixed in order to provide a high frequency response for the proposed filter design. Finally, the obtained results display a good performance for the proposed filter design and demonstrate a high usefulness for the employed technology that allows a low design volume.**

***Keywords*—Filter, SIW, HMSIW, CM, Ka band, Wireless applications**

I. INTRODUCTION

WIRELESS communication systems become today very saturated because of the huge demand for operated frequency bands serving different applications that needs to broaden the frequency range by developing new effective techniques and technologies to design the different passive and active microwave circuits. In fact, many research studies have carried out to investigate superconducting front-end planar devices at Ka frequency band in order to provide additional operating frequencies to satisfy the increasing requirements of wireless communications in the future [1]. Ka frequency band extending from 26 to 40 GHz checks a large range to design efficient microwave components with reduced weight and consumption and inexpensive cost.

Waveguides were traditionally used to design high-performance filters with important sizes, but by the time, it will be necessary to transform such bulk designs into planar forms easy to be integrated that claimed a very complicated process as a straight solution to combine rectangular waveguides into microstrip structures knowing by substrate integrated waveguide technology [2] that provides a good quality factor (Q) at the filter input due to the dielectric filling resulting from the size reduction [3]. Accordingly, SIW and half-mode SIW technologies [4-5] become very popular manufacturing processes for microwave filter designs, very easy to be integrated to planar forms and allow a high performance in terms of low volume, low insertion loss (IL) and good quality factor comparing to conventional microstrip or coplanar (CPW) easily integrated to planar designs that low volume.

This research paper presents three topologies of a novel 3rd order bandpass filter design using SIW/HMSIW technology with shunt-inductive posts operating at Ka frequency band. The first filter topology uses a simple cavity structure however the second one employs a SIW structure to eliminate losses by avoiding the disturb nearby of resonance, whereas the third topology employs a half-mode SIW filter to suppress the higher mode harmonics and control the mutual coupling. The paper is organized as follows: Section I explains SIW processing. Section II demonstrates the coupling matrix calculation. Section III and IV describes the SIW filter designs ‘synthesis, simulation and optimization and section V presents a comparison between the proposed topologies.

II. SUBSTRATE INTEGRATED WAVE GUIDE FILTERS

Designing SIW filters be easily realized and formed by either a shunt inductive post or an shunt inductive iris aperture as reported in [6]. As shown in Figure 1, the first example of the developed SIW passband filter model employs an iris inductive shunt; different resonances are caused due to the inductive shunt coupling in addition to two microstrip-to-SIW transitions at the input and output. The smallest inductive shunts work to facilitate the input/output coupling while the biggest ones provide the inter-resonator coupling [7].

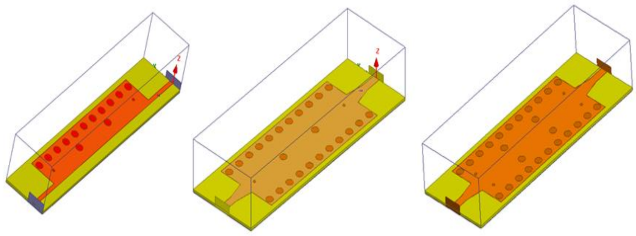

Fig. 1. The different proposed SIW/HMSIW bandpass filter designs

The filter designing procedures are similar to those of an air-filled waveguide filter basing on the coupling-matrix technique.

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Finally, the modeled SIW filter properties offer opportunity to manufacture a microwave filter with low profile and low cost.

III. Formatting Coupling Matrix From The Low-Pass Prototype Using Cross-Coupled

The coupling matrix is a theory used to represent the filter structures based on lumped element circuit [1,8]. Figure (2) shows nth order equivalent circuit with electric coupled model. The Kirchhoff's current law can be used for analyzing the equivalent circuit of the electric coupled resonators as given in Figure (2). The conductance, inductance, and capacitance of the circuit are presented by G, L and C respectively, vi present the node voltage and is is the current source [1]

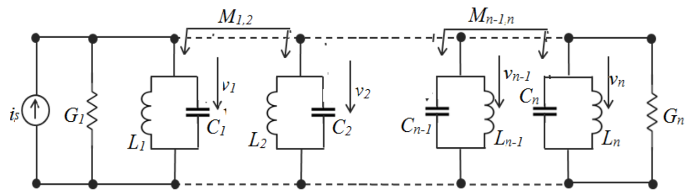


Fig. 2. Equivalent circuits of n[th] coupled resonators electric coupled resonators

By applying the Kirchhoff's node equations to the circuit of electric coupled resonators as given in [8]

$$\begin{aligned}
&\left(G_1 + j\omega C_1 + \frac{1}{j\omega L_1}\right) v_1 - j\omega C_{12} v_2 \ldots - j\omega C_{1n} v_n = i_s \\
&-j\omega C_{21} v_1 + \left(j\omega C_2 + \frac{1}{j\omega L_2}\right) v_2 \cdots - j\omega C_{2n} v_n = 0 \\
&\vdots \\
&-j\omega C_{n1} v_1 - j\omega C_{n2} v_2 \cdots + \left(G_n + j\omega C_n + \frac{1}{j\omega L_n}\right) v_n = 0
\end{aligned} \tag{1}$$

where $C_{ij} = C_{ji}$ are the mutual capacitance across resonators i[th] and j[th]. The node voltages are presented by respect to the ground. The Kirchhoff's currents equations (1) are arrange in matrix form as [8]

$$\begin{bmatrix} G_1 + j\omega C_1 + \frac{1}{j\omega L_1} & -j\omega C_{12} & \cdots & -j\omega C_{1n} \\ -j\omega C_{21} & j\omega C_2 + \frac{1}{j\omega L_2} & \cdots & -j\omega C_{2n} \\ \vdots & \vdots & \vdots & \vdots \\ -j\omega C_{n1} & -j\omega C_{n2} & \cdots & G_n + j\omega C_n + \frac{1}{j\omega L_n} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \\ \vdots \\ v_n \end{bmatrix} = \begin{bmatrix} i_s \\ 0 \\ \vdots \\ 0 \end{bmatrix} \tag{2}$$

with $[Y].[v] = [i]$

where [Y] is an (n x n) admittance matrix. If the resonator filter is supposed to be synchronously changed, this signify the microwave resonators of circuit resonate at the same frequency called center frequency $\omega_0 = 1/\sqrt{LC}$. The [Y] admittance matrix can be presented by:

$$[Y] = \omega_0 C \cdot FBW \cdot [\bar{Y}] \tag{3}$$

where $[\bar{Y}]$ represents the normalized admittance matrix and FBW is the fraction bandwidth, the $[\bar{Y}]$ of a synchronously changed circuit is given by [8]

$$[\bar{Y}] = \begin{bmatrix} \frac{G_1}{\omega_0 C \cdot FBW} + p & -j\frac{\omega}{\omega_0}\frac{C_{12}}{C}\cdot\frac{1}{FBW} & \cdots & -j\frac{\omega}{\omega_0}\frac{C_{1n}}{C}\cdot\frac{1}{FBW} \\ -j\frac{\omega}{\omega_0}\frac{C_{21}}{C}\cdot\frac{1}{FBW} & p & \cdots & -j\frac{\omega}{\omega_0}\frac{C_{2n}}{C}\cdot\frac{1}{FBW} \\ \vdots & \vdots & \vdots & \vdots \\ -j\frac{\omega}{\omega_0}\frac{C_{n1}}{C}\cdot\frac{1}{FBW} & -j\frac{\omega}{\omega_0}\frac{C_{n2}}{C}\cdot\frac{1}{FBW} & \cdots & \frac{G_n}{\omega_0 C \cdot FBW} + p \end{bmatrix} \tag{4}$$

where p is the complex lowpass frequency variable. Notice that

$$\frac{G_i}{\omega_0 C} = \frac{1}{Q_{ei}} \text{ for } i = 1, n \tag{5}$$

with Q_e is the external quality factor and the coupling coefficient is given by

$$M_{ij} = \frac{C_{ij}}{C} \tag{6}$$

The easiest expression of equation (4) is can be written by:

$$[\bar{Y}] = \begin{bmatrix} \frac{1}{q_{e1}} + p & -jm_{12} & \cdots & -jm_{1n} \\ -jm_{21} & p & \cdots & -jm_{2n} \\ \vdots & \vdots & \vdots & \vdots \\ -jm_{n1} & -jm_{n2} & \cdots & \frac{1}{q_{en}} + p \end{bmatrix} \tag{7}$$

The former parameters (p, m_{12}, q_{e1} and q_{en}) are depicted below [8]:

$$p = j\frac{1}{FBW}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \tag{8}$$

$$m_{i,j} = \frac{M_{i,j}}{FBW} \tag{9}$$

$$q_{ei} = Q_{ei} \cdot FBW \text{ for } i = 1, n \tag{10}$$

or the two-port filter as in Figure 2 the S-parameters can be calculated from [3]:

$$\begin{aligned} S_{11} &= \pm\left(1 - \frac{2}{\sqrt{q_{e1}}}[A]_{11}^{-1}\right) \\ S_{21} &= \frac{2}{\sqrt{q_{e1} \cdot q_{en}}}[A]_{n1}^{-1} \end{aligned} \tag{11}$$

The general matrix [A] formed of coupling coefficients m_{ij} and external quality factors $q_{i,ext}$ is presented in [4] as:

$$[A] = \begin{bmatrix} \frac{1}{q_{e1}} & 0 \cdots & 0 \\ 0 & 0 & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \frac{1}{q_{en}} \end{bmatrix} + p\begin{bmatrix} 1 & 0\ldots & 0 \\ 0 & 1 & 0 \\ \vdots & \vdots \cdots & \vdots \\ 0 & 0\ldots & 1 \end{bmatrix} - j\begin{bmatrix} m_{1,1} & m_{1,2}\ldots & m_{1,n} \\ m_{2,1} & m_{2,2} \cdots & m_{2,n} \\ \vdots & \vdots & \vdots \\ m_{n,1} & m_{n,2}\ldots & m_{n,n} \end{bmatrix} \tag{12}$$

Where, matrix [U] is an identity matrix, p is the complex lowpass frequency variable, ω_0 is the centre frequency of the filter, FBW is the fractional bandwidth of the filter. $q_{i,ext}$ (i=1 and n)is the scaled external quality factors of the resonator i. m_{ij} is the normalized coupling coefficients between the resonator i and j [9].

IV. SIW FILTER DESIGN USING CROSS-COUPLED RESONATORS

For the demonstration, a highly selective third-pole SIW filter with the configuration of Figure 3 has been designed. A cross-coupled is given as an example to illustrate the proposed method. The target specifications of the SIW Bandpass filter are given by:

TABLE I
SIW BANDPASS FILTER'S TARGET SPECIFICATIONS

Center frequecy	f_0= 25.96 GHz
Bandwidth	BW= 3.56 GHz (FBW=13.71 %)
Passband ripple,	L_{AR} = 0.0431dB
Passband return loss	RL=20dB
Stopband rejection level	> 40 dB

The element values of the lowpass prototype filter are found to be:

$g_1 = g_3 = 0.8516 \quad , g_2 = 1.1032$

The design parameters of the Bandpass filter, i.e the de-normalized coupling coefficients and the scaled de-normalized external quality factors can be determined by the formulas [10]:

$$Q_{eS} = \frac{g_0 g_1}{FBW} \quad , \; Q_{eL} = \frac{g_n g_{n+1}}{FBW} \tag{13}$$

$$M_{i,i+1} = \frac{FBW}{\sqrt{g_i g_{i+1}}} \quad , i = 1 \;\; to \;\; n-1 \tag{14}$$

$$m_{i,i+1} = \frac{M_{i,i+1}}{FBW} \quad , i = 1,\ldots n \tag{15}$$

From Equation (14),in the first step of the design, the lowpass prototype filter elements with the following ideal coupling matrix are evaluated:

From Equation (11), the design parameters of this bandpass filter are found:

$$Q_{ext\_e} = Q_{ext\_s} = 5.9475$$

The capacitance C_0 has been also calculated and the inductance L_0 using the following equations:

The series impedances Z_{12}, Z_{23}, Z_{34} are:

$$Z_{i,i+1} = \frac{Z}{M_{i-1,i} \times Q_{e1}} \tag{16}$$

The series impedances

Z_{01}	Z_{12}	Z_{23}	Z_{01}
50	58.1512	58.1512	50

Figure (4) shows the corresponding coupling/routing diagram comprised of SIW resonators to implement these type of filtering characteristic in a SIW technology. Here the $M_{i,i+1}$ indicate the sequence of direct coupling:

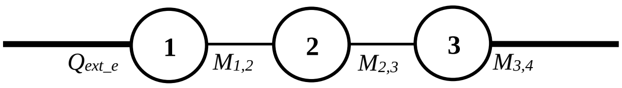
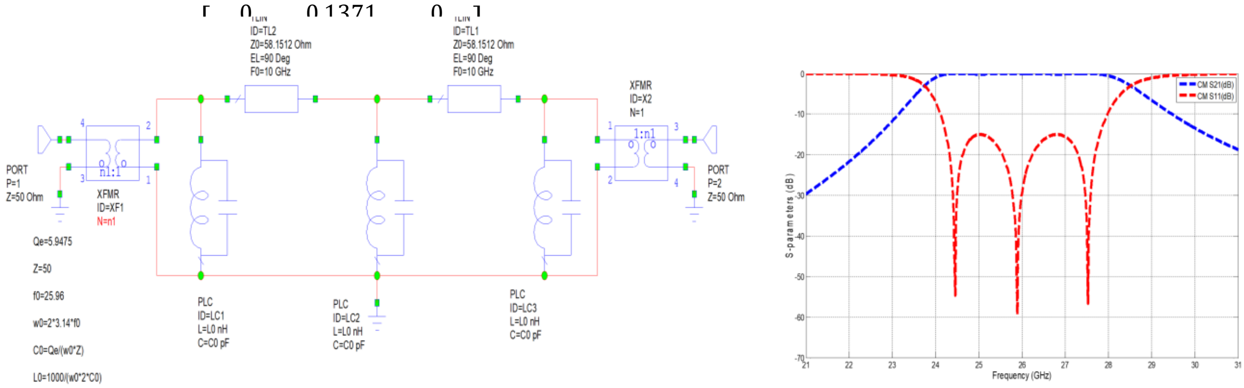


Fig. 3. A generalized model for SIW coupled resonator filter

Figure 4 shows an equivalent circuit of the BPF .This circuit consists of three circuit LCR parallel resonators which are separate through the $\lambda/4$-wavelength microstrip lines. As can be seen from figure 4, the transmission line θ=±90° at the center frequency $\omega = 1/\sqrt{LC}$ are employed to represents the coupling coefficient. This circuit model does account for the effect of losses. The RLC microstrip transmission line unit section is simulated using microwave office AWR.



a) b)

Fig. 4. a) The coarse model of the 3rd order HMSIW bandpass filter with Microwave Office, b) Ideal response of the 3rdorder HMSIW pandpass filter based on the CM technique.

The proposed filter's ideal frequency response of the equivalent circuit is displayed in Figure 4 which provides a return loss coefficient of less than 25dB for a passband of 1.05 GHz at a center frequency equal to 28.86 GHz.

A. Implementation of the SIW Filter by Shunt inductive post

In this section, a 3rd order iris SIW bandpass filter is designed and simulated using Arlon DiClad 880 substrate with a thickness of 0.25mm, a dielectric constant of 2.2 and a loss tangent of 0.009. The desired resonance frequency is of 28.86 GHz. From the fundamental frequency equation, the diameter *d* of the metal sights is given by 0.7mm with a pitch *P* of 1.04mm(*eq. 17*) to prevent radiation leakage [11].

$$\begin{cases} \lambda_c = \dfrac{c}{f_{101}\sqrt{\varepsilon_r}} = \dfrac{3\times10^8}{28.86\times10^9\sqrt{2.2}} \approx 7mm \\ \dfrac{d}{\lambda_c} \approx 0.1 \Rightarrow d = 0.7mm \\ \dfrac{d}{P} \approx 0.67 \Rightarrow P = 1.04mm \end{cases} \quad (17)$$

Leq length and *Weq* width are equal to 5mm for a single SIW square cavity as calculated in:

$$f_{101} = \frac{c}{2\pi\sqrt{\varepsilon_r}}\sqrt{\left(\frac{\pi}{Weq}\right)^2 + \left(\frac{\pi}{Leq}\right)^2} = 28.86\ GHz \quad (18)$$

$$Weq = Leq = 5mm \quad (Square\ cavity)$$

The W_{SIW} spacing between the two rows of holes is a relevent physical parameter that is necessary for designing a SIW structure. It can be determined from the empirical equations as a function of the equivalent width (W_{eq}) of the rectangular waveguide, giving the same characteristics in fundamental mode for the same height and the same dielectric constant:

$$W_{SIW} = Weq + \frac{d^2}{0.95P} \approx 5.5mm \quad (19)$$

$$W_{SIW} = L = 5.5mm \quad (Square\ cavity)$$

$$W_t = 0.4(W_{SIW} - d) \approx 7.304mm \quad (20)$$

$$\begin{cases} \dfrac{\lambda}{2} \le L_t \le \lambda \\ \lambda = \dfrac{c}{f_0\sqrt{\varepsilon_r}} \\ Lt \approx 13.12mm \end{cases} \quad (21)$$

The filter design may support the same procedure of designing an air-filled waveguide filter as it may be developed based on the coupling-matrix method as discussed in previous sections. Figure 4 shows the layout of a SIW filter designed on a dielectric substrate with a relative dielectric constant of 2.2 and a thickness of 0.25 mm. Figure. 5 plots the EM-simulated performance of the filter, obtained using available simulation tool. The simulation assumes a loss tangent tan δ = 0.001 for the substrate and a conductivity σ = 5.8 $\times 10^7$ S/m for all the metals (copper) with 17-µm thickness. The optimal structure dimensions of the 3rd order SIW bandpass filter are presented in the table below:

TABLE II
3rd ORDER BANDPASS FILTER'S OPTIMAL DIMENSIONS

Settings	***Value (mm)***	***Settings***	***Value (mm)***
W_{50}	*0.8*	l_{50}	*2.00*
W_t	*1.6*	L_t	*4.00*
W_{SIW}	*5.6*	a_1	*04.95*
d_1	*0.80*	a_1	*04.50*
d_2	*0.15*	*k*	*2.4*

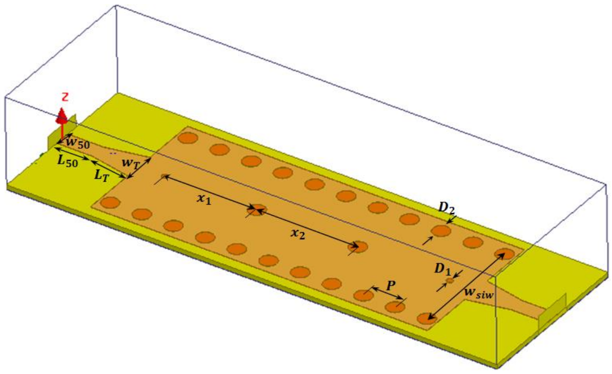


Fig. 5. Geometric design of the SIW filter using shunt-inductive posts for coupling

The SIW filter consists of four coupling posts centered in the guide and two microstrip-to-SIW transitions at the input and output, respectively. The small posts facilitate the input and output couplings, while the large posts are for the inter-resonator couplings.

Figure 6 displays the filter' frequency response when simulating the structure for a bandpass of more than 1.25GHz from 27.91GHz 29.16 GHz. Accordingly, the structure is therefore exposed for efficient optimization procedures using HFSS software to obtain optimal geometric parameters for the proposed structure and improve the filter response.

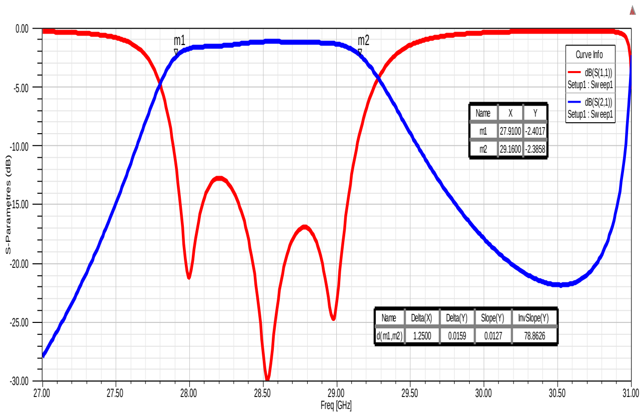


Fig. 6. S-parameters of the proposed SIW bandpass filter with initial geometric parameters

1) W_{SIW} spacing variation effect

As it appears in Figure 7, after optimizing the post inductive shunt filter by HFSS simulator, the frequency response becomes very close to the ideal specifications set before with a bandpass of 1.25 GHz, a center frequency around 28.53 GHz and reflection losses of less than 15 dB.

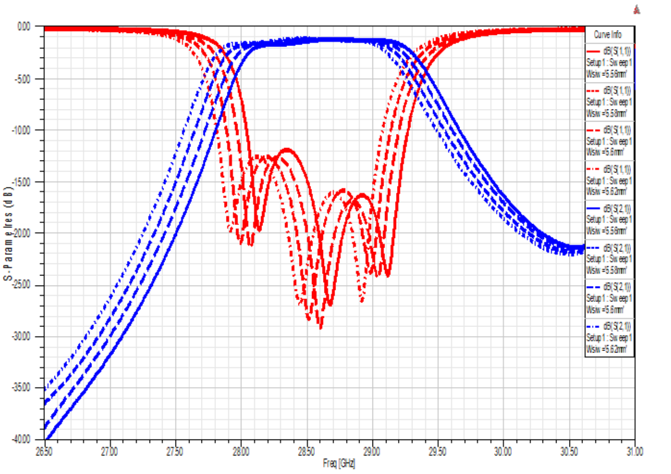


Fig. 7. W_{SIW} spacing variation effectfor the proposed 3[rd] SIW bandpass filter

Figure 8 presents the simulated EM-HFSS field layout of the proposed SIW bandpass filter for TE101 propagation mode at the center frequency. The EM-analyser demonstrates that the E-field distibution has two circular propagating zones in the single SIW filter cavity.

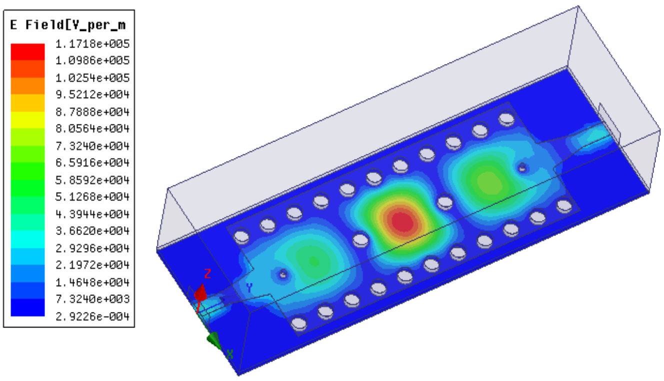


Fig. 8. Simulated E-field magnitude distributions obtained by SIW bandpass filter

B. Implementation of the SIW Filter by Shunt inductive iris

Thethree dimensional topology of the proposed 3[rd]order SIW bandpass filter is shown in Figure 9. The three resonators are coupled together using inductive iris while the two end resonators are coupled to exciteand load by means of microstrip line characteristic impedance of 50 ohm.

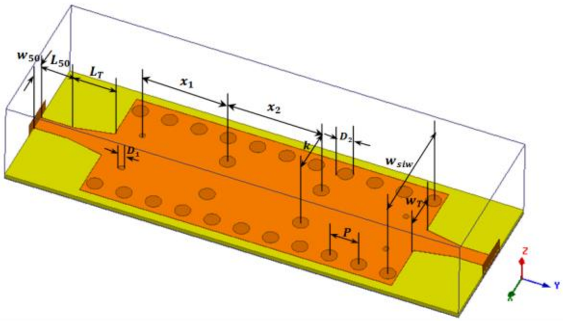


Fig. 9. Geometricdesign of the SIW filter using shunt-inductive iris for coupling

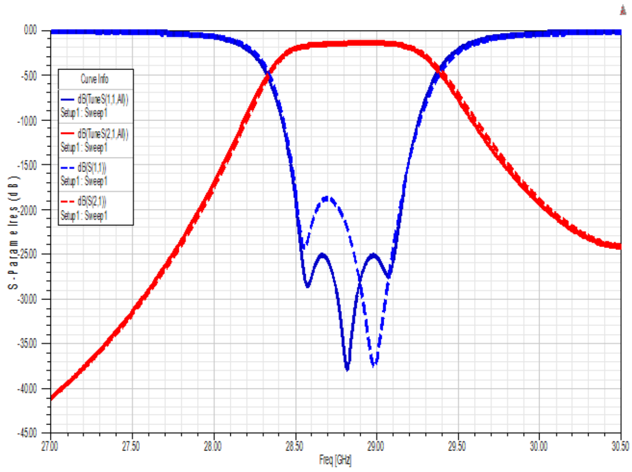


Fig. 10. S-parameters of the proposed SIW bandpass filter with optimal geometric parameters

Figure 10 shows a high frequency response for the optimized structure with optimal geometric parameters displayed in a reflection coefficient of less than 35dB and a transmission coefficient of approximately 1dB for the resonance frequency that is equal to 28.86 GHz and a bandpass of 1.05 GHz. Table 3 presents in details the different dimensional changes extracted from the optimization procedures.

TABLE III
OPTIMAL STRUCTURE DIMENSIONS OF THE 3[RD]ORDERSIW BANDPASS FILTER

Settings	*Value (mm)*	*Settings*	*Value (mm)*
W_{50}	0.8	l_{50}	2.00
W_t	1.6	L_t	4.00
W_{SIW}	5.6	a_1	04.95
d_1	0.80	a_1	04.50
d_2	0.15	k	2.4

Figure 11 presents the simulated EM-HFSS field layout of the proposed SIW bandpass filter for TE101 propagation mode at the center frequency. The EM-analyser demonstrates that the E-field distribution has two circular propagating zones in the single SIW filter cavity.

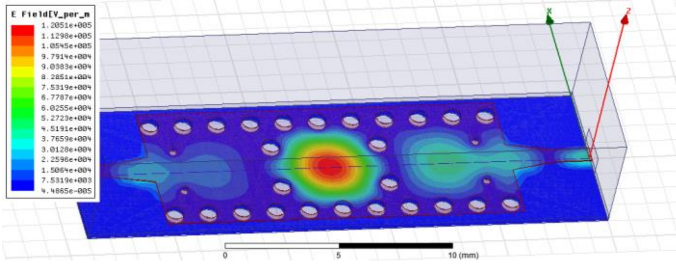


Fig. 11. Simulated E-field magnitude distributions obtained by SIW bandpass filter

SIW Variation Effects Structure

1) W_{SIW} spacing variating effect

As it appears in Figure 12, after optimizing the iris inductive shunt filter by HFSS simulator, the frequency response becomes very close to the ideal specifications set before with a bandpass of 1.045 GHz, a center frequency around 28.85 GHz and reflection losses of less than 25 dB.

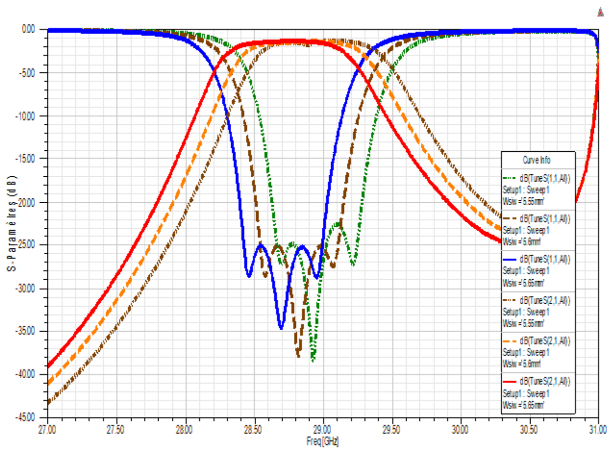


Fig. 12. WSIW spacing variating effectfor the proposed 3rd SIW bandpass filter

2) Via hole diameter variation effect

Curves obtained in Figure 13, demonstrate that variating via hole diameter influences the center frequency inversely to the W_{SIW} width variation effect. In fact, increasing the diameter provides a right shifting whereas decreasing it provokes a left shifting. As a result, the resonant frequency can be a function of both via hole diameter and W_{SIW} width.

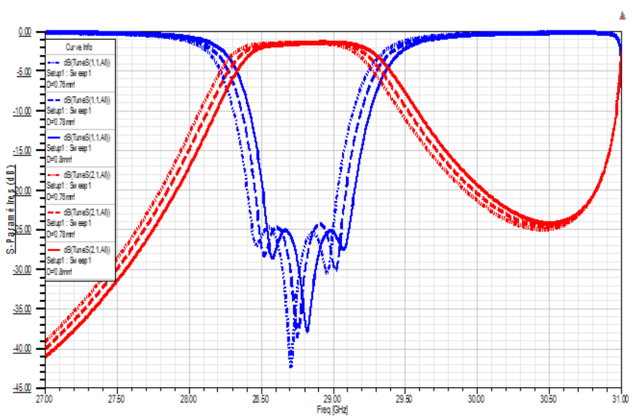


Fig. 13. Variation of the diameterD of the proposed bandpass filter SIW

C. Bandpass HMSIW Filter Using shunt-inductive posts

The structure of the proposed model of the 3rdbandpass HMSIW filter have been modified by implementing a shunt-inductive posts as shown in Figure 14.

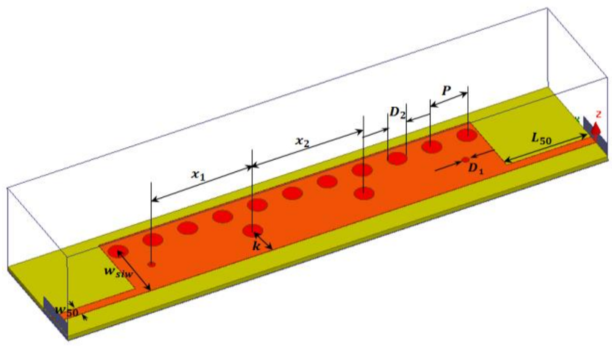


Fig. 14. Geometric design of HMSIW filter with Geometric design of SIW filter.

Simulation results presented in Figure 15 show that the new filter response becomes better in terms of low return loss of 40 dB and good transmission with 0.2dB for a central frequency of approximately 25.93 GHz and a fractional bandwidth of 13.73%.

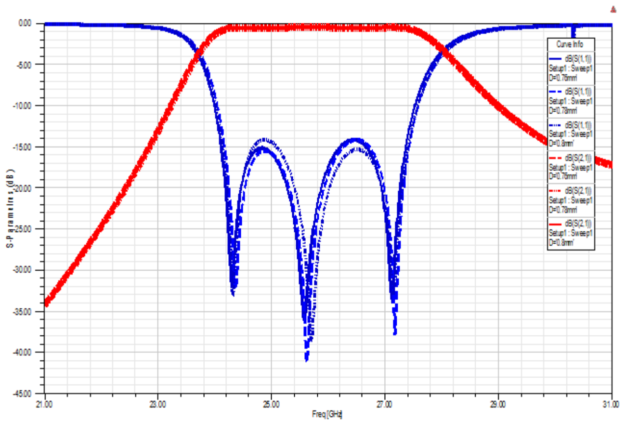


Fig. 15. S-parameters of the proposed HMSIW bandpass filter with optimal geometric parameters

1. W_{SIW} spacing variating effect

Curves obtained in Figure 16, demonstrate that variating the W_{SIW} width variation effect influences the center frequency. As a result, the resonant frequency can be a function of both W_{SIW} width.

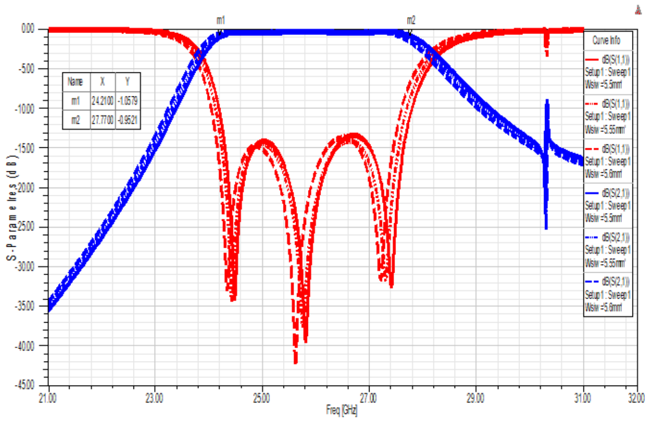


Fig. 16. S-Parameters of the new 3rd SIW bandpass iris shunt inductive filter based on DGS technique

The distribution of E-fields doesn't change when implementing HM as observed in Figure 17. This demonstrates that HMSIW technique works to enhance the filter electromagnetic response without affecting the field distribution.

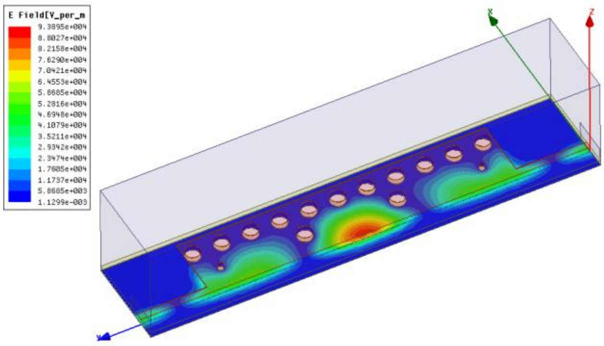


Fig. 17. Simulated E-field magnitude distributions obtained by SIW -DGS bandpass filter.

In order to validate the two proposed SIW filter models, a specific comparison to [12] using shunt inductive posts is done for the same order (3) and same bandpass ranging from 28.34 to 29.39 GHz as presented in Table 4:

TABLE IV
EM FREQUENCY RESPONSE COMPARISON

Reference	*FBW* and f_0**	*Insertion loss (dB)*	*Return loss(dB)*
[8] : Filter using inductive shunt posts	*4.38% and 28.53*	*1.2*	*15*
1st Model : SIW filter with shunt inductive iris	*3.63 % and 28.86*	*1.5*	*25*
2nd Model : HMSIW filter using shunt inductiveposts	*13.73 % and 25.93*	*0.48*	*15*

It is clearly observed that the proposed filter designs present a good electromagnetic performance for Ka band displayed by a low return loss and good transmission comparing to the other research works that validate both of the accuracy of used modeling method and the efficiency of the selected EM simulator.

CONCLUSION

This research work presents three different topologies of a new 3rd order bandpass filter using SIW technology and HMSIW technique with shunt inductive posts and iris. The filters are developed using Coupling Matrix method which presents a very useful tool for direct filter synthesis. The proposed filters designs have been simulated by HFSS for operating in Ka band applications. The first SIW filter design uses shunt-inductive posts implemented on a single-layer Arlon DiClad 880 (tm) with a permittivity of 2.2 and substrate thickness of 0.25 mm, operating on a center frequency equals to 28.53 GHz and a passband width from 27.91 GHz and 29.16 GHz. The second SIW filter design uses a shunt-inductive iris for the same center frequency and a bandpass of 1.045 GHz. The third topology employs a half-mode SIW filter from 24.21 to 27.77 GHz and a center frequency around 25.93 GHz. Analytical formulas coupling matrix (CM)are used to calculate the filter's different geometric parameters and efficient optimization techniques by EM simulators are employed to reduce the effective dielectric permittivity on certain waveguide sections in order to achieve a high performance demonstrated by a low reflection coefficient and good transmission coefficient for the whole bandpass. Such a type of filters is designed to be used in Ka frequency band that is very suitable for wireless communications by satellites.

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