A Novel HPF of Substrate Integrated Waveguide

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Summary
Utilizing filters to suppress spurious signals is inevitable in communication systems. In some applications, where the level of spurious signals is high or the frequency of LO is near to IF (in up-conversion), increasing the slope of the filter to eliminate spurious signals is unavoidable.
In this paper we are going to present an idea about usage of evanescent mode of Substrate Integrated Waveguide (SIW) as a High Pass Filter (HPF), to excellently reject spurious signals and a novel method for matching between the filter and a microstrip line for this filter is presented.
The mathematical calculations based on the presented method and the structure, and also the similarities between the calculations and the simulations done by HFSS confirm the efficiency of this new method. The validation of method is shown by measurement results.

Key words:
Evanescent Mode, SIW HPF, Matching, Microstrip Line

1. Introduction

In many communication systems, it is necessary to eliminate spurious signals of first RF stages, such as leaked signal of LO in the output RF. In this case a suitable RF filter should be used, but sometimes the desired signal and the spurious are too close together, or the level of spurious signals are high, so that the amount of rejection for this filter should be seriously considered. The implementation of a microstrip BPF after the mixer is not easy in higher frequencies because microstrip filters in higher frequencies have high in-band insertion loss due to low Q factor of their resonators, and sometimes it can make their fabrication impossible. Waveguide filters have good Q factor but in many application it is not easy to use them because of their incompatibility with planar integrated circuits.

Nowadays, the Substrate Integrated Waveguides are considerably used in Microwave and millimeter waves in various applications, including microwave and millimeter wave filters[1]. Although their loss due to dielectric materials is more than usual waveguides, in-band loss in SIW filters are definitely less than microstrip filters[2].

The pass band insertion loss for band pass filters is calculated by Eq.1 [3]

$$I.Loss = 4.43\frac{f_0\sum g_i}{\Delta f Q_u} \tag{1}$$

In the Eq.1 f_0 is the central frequency, g_i is the ith element value of Low Pass Filter prototypes, Δf is the bandwidth and Q_u is the unloaded Q factor.

As Q_u is low in microstrip filters, the insertion loss of pass band in higher frequencies is too high. For instance, to implement a direct modulator in Ku-Band at 14GHz and with 0.5GHz bandwidth, and 60dB attenuation at 12GHz (to suppress LO.), it is needed to design a 5th order filter, and if a microstrip filter is chosen, the insertion loss of pass band will be 9dB (according to Eq. 1), which is so high at this frequency.

According to Eq.1 the more filter's order and higher central frequency and less bandwidth, the more insertion loss of filter inside the pass band.

When the spurious signals are lower than the desired frequency, the band pass filter could be replaced by a high pass filter. one of the advantages of high pass filters is their easy fabrication, because there is no need to implement the middle vias as posts, particular when SIW filters are utilized, and moreover, the amount of attenuation of the out-band is selectable by increment or decrement of the length of the SIW.

If SIW is used to fabricate microwave HPF, a good solution is to choose dimensions so that the desired bandwidth and the spurious frequencies would be above and below of the cut off frequency respectively. Therefore, we will have a HPF with the same cut of frequency of the waveguide. In this case, the amount of attenuation for spurious signals should be calculated by the amount of evanescent mode attenuation along the waveguide.

As the previous methods [4], for matching between the microstrip lines and SIW near the cut off frequency could have not offered a good solution, a novel method is presented in this paper for matching between SIW and microstrip lines. Indeed, calculation of input impedance of SIW inside a small limited bandwidth and in a certain point along the width of the SIW is the key of this idea, and then performing matching between the transmission line and the SIW, by a Quarter-wave Transformer microstrip.

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2. Evanescent Mode in waveguide

In a rectangular waveguide, when the operating frequency is lower than cut off frequency of TE_{10} mode, all mode are propagated as evanescent mode, and the amount of attenuation depends of the length of waveguide and the propagation constant.

Under the cut of frequency of TE_{10} mode, the attenuation of other modes are more, and moreover as the width of waveguide is much more than the thickness of the substrate for SIW (a>>h), the propagation frequency of many other modes like TM_{11} are much higher than TE_{10} mode and the attenuation is utterly great for frequencies less than the cut of frequency of TE_{10}. Therefore, if the length of waveguide is not so short, TE_{10} is the dominant mode for calculation of the attenuation inside the waveguide.

The electrical and magnetic fields in waveguides for TE_{10} mode are obtained by Eq. 2

$$H_z(x) = A_{10}\cos\frac{\pi x}{a}e^{-j\beta z}$$
$$E_y(x) = \frac{-j\omega\mu\pi}{k_c^2 a}\sin\frac{\pi x}{a}e^{-j\beta z} \qquad (2)$$
$$H_x(x) = \frac{j\beta\pi}{k_c^2 a}\sin\frac{\pi x}{a}e^{-j\beta z}$$

Dimensions of the waveguide in a Cartesian coordinate system, illustrated by Fig. 1

Where <u>a</u> and <u>h</u> are the width and height of the rectangular waveguide respectively.

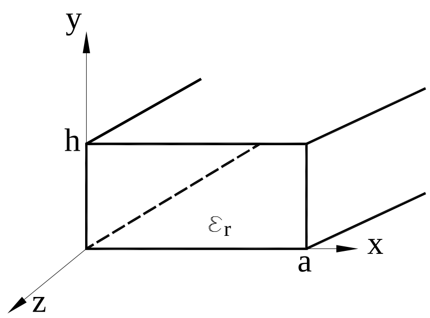

Fig.1. Waveguide dimensions in Cartesian coordinate

Eq. 3 defines the parameters of Eq. 2. Where f_c is the cut off frequency. If $k < k_c$ then β is imaginary ($\beta = j\alpha$) and consequently: $e^{-j\beta z} = e^{\alpha z}$

Consequently, the waveguide shows loss in frequencies below the cut off frequency, according to the length of the waveguide. Fig.2 illustrates the amount of loss in term of the length of waveguide in the evanescent mode for h=0.254mm, a=7.4mm, $\varepsilon_r = 2.33$, and length up to 120mm.

$$k_c = \frac{\pi}{a}$$
$$\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - (\frac{\pi}{a})^2} \qquad (3)$$
$$k = \frac{\omega}{c}\sqrt{\varepsilon_r}$$
$$f_c = \frac{k_c}{2\pi\sqrt{\mu\varepsilon}} = \frac{1}{2a\sqrt{\mu\varepsilon}}$$

According to the Fig.2 for a certain waveguide, the more increase in the length of waveguide, the more attenuation (dB) in cut off frequency, and the slope of attenuation line is increased, when the frequency is reduced.

The lower frequency the higher insertion loss, and it confirms that when waveguide is used as a HPF, the steep slope of the filter is completely acceptable in many applications.

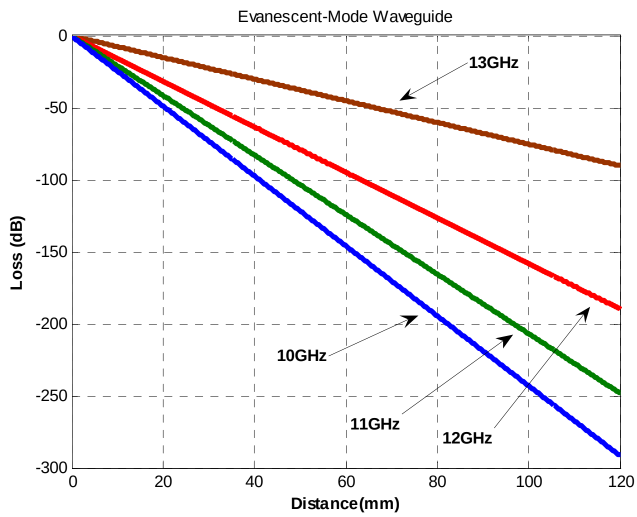

Fig.2. Insertion Loss of SIW with a=7.4mm, h=0.254mm, ε_r=2.33 and for the length of up to 120mm

3. A novel suggestion for matching

To eliminate spurious frequencies below the cut off frequency, which are close to the main signal, a HPF with steep slope is used. When a HPF is designed by SIW, the cut off frequency of the SIW is the same as the cut off frequency of the filter and also near the main signal. But near the cut off frequency, because of high variation of the propagation constant (β) in the waveguide as a function of frequency, the variation in the input impedance of SIW is very high. Therefore, matching between the SIW and the transitions lines with constant impedance characteristic (Normally 50 Ohms) is hard.

There are some researches for transition between the waveguide and the micrpstrip, which have been done ([5-8]), and also there is a good way for matching between the SIW and the microstrip in [4], where a microstrip tapered line is used for matching between quasi-TEM mode micrstrip, and waveguide in TE_{10} mode. But this method for Matching between the SIW and Microstrip lines is not suitable. where the pass band frequency is near to the cut off frequency of the lowest dominant mode of SIW, due to rapid variation of waveguide impedance as a function of frequency. The proposed method is to calculate the input impedance of SIW.

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The characteristic impedance (or the input impedance) has a certain definition for transmission lines, and is defined by the ratio of the voltage to the current at every cross section of transmission lines. The transmission lines that carry TEM Electromagnetic waves have a unique voltage and current in all frequencies and at all cross sections. Therefore, their characteristic impedance is constant and unique (certainly in all frequencies and at all cross sections) with a clear concept. But in waveguides (and also SIW), as TE and TM waves are propagated, we can not obtain a certain definition for the characteristic impedance, because it is a function of the frequency (Fig.4 and Fig.5), and also a function of physical location (Eq.5 and Eq.6). Therefore, their characteristic impedance is not unique. But if the waveguide functions in a limited frequency bandwidth and the transmission line is connected to the certain point of the waveguide, then utilizing of characteristic impedance is so useful, and can be used for impedance matching.

Therefore, the new offered method is to calculate the input impedance (or the characteristic impedance) of SIW in a narrow pass-band and in a certain point along the width of SIW, and then making matching between the transmission line and the SIW, by a Quarter-wave microstrip transformer, so that its characteristic impedance is:

$$Z_{oTR} = \sqrt{Z_o . Z_{in}} \tag{4}$$

In Eq.4, Z_{OTR} is the characteristic Impedance of the Quarter-wave microstrip transformer, Z_o is the characteristic Impedance of the transmission line and Z_{in} is the input impedance of SIW.
The direction of the voltage and the current in the SIW are illustrated in Fig.3:

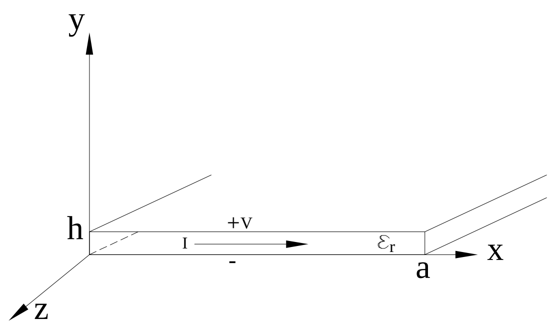

Fig.3. The directions of the voltage and the current in the SIW

The values of the voltage and the current across the input of the SIW are obtained by Eq. 5:

$$\begin{aligned} V &= \int_{y=0}^{h} E_y dy = \frac{j\omega\mu\pi}{k_c^2 a} \sin\frac{\pi x}{a} \int_0^h dy = \frac{j\omega\mu\pi}{k_c^2 a} h \sin\frac{\pi x}{a} \\ I &= \int_{x=0}^{a} H_x dx = \frac{j\beta\pi}{k_c^2 a} \int_0^a \sin\frac{\pi x}{a} dx = \frac{2 ja\beta}{k_c^2 a} \end{aligned} \tag{5}$$

As it is shown, the current is constant, and independent of x but the voltage is a function of x. On the other hand, in TE_{10} mode, the voltage and the electrical field are maximum in the middle point of the width of SIW, and they have no significant variation around the maximum point.

Also the width of microstrip line, entered into the middle of SIW, is narrow enough that we can suppose a constant voltage across the middle of SIW. So the voltage is equal to Eq. 6.

$$V = \frac{j\omega\mu\pi}{k_c^2 a} h \tag{6}$$

Therefore the input impedance of SIW is equal to Eq. 7.

$$Z_{in} = \frac{V}{I} = \frac{\omega\mu h\pi}{2\beta a} = \frac{k}{\beta}\eta(\frac{\pi h}{2a})$$
$$\eta = \frac{\eta_0}{\sqrt{\varepsilon_r}} \qquad (7)$$

The variation of Z_{in} for $\varepsilon_r = 2.33$ and a=7.4mm is illustrated in Fig.4. Obviously, the variation of input impedance near the cut off frequency (13GHz) is high, and regarding to the Fig. 4, the more thickness of the substrate, the more variation of input impedance matching. Fig. 5 illustrates the variation of Z_{in} by the frequency for different ε_r for a=7.4mm. As ε_r is increased, both the cut off frequency of SIW and the final limit of input impedance is decreased. The finial limit of input impedance is the point, where the impedance goes toward the infinity.

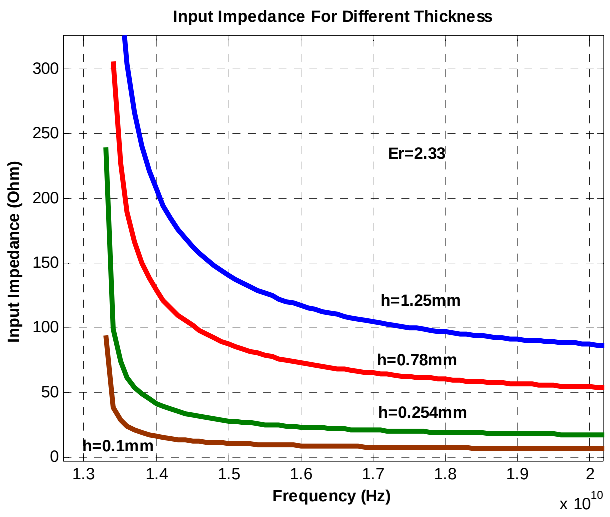

Fig.4. The variation of Z_{in} by the frequency for different substrate thicknesses and constant ε_r

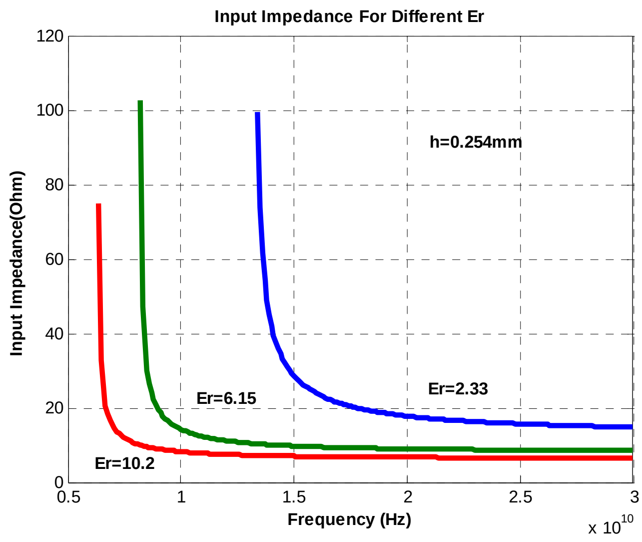

Fig.5. The variation of Z_{in} by the frequency for different ε_r and h=0.254mm

4. Simulation results

The layout of the HPF is illustrated in Fig.6, implemented on RT/Duroid 5850 (h=0.254mm, a=7.58mm, L=50mm, W=1mm, T=3.6mm, D=0.5mm and P=1.5mm). For analysis and fabrication of the SIW's walls, 0.5mm vias with 1.5mm spacing are utilized.

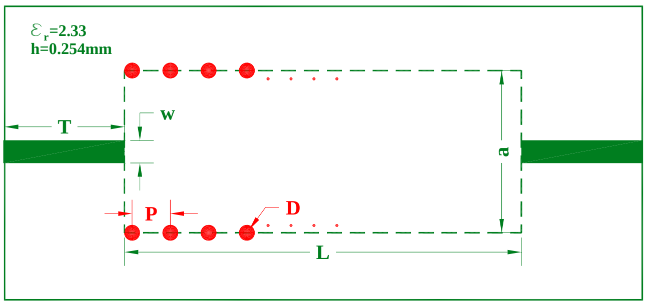

Fig.6. The Layout of the HPF

To calculate the cut off frequency of the SIW, the equivalent width of the SIW is a considerable note. In the above example, the cut off frequency is 13.4GHz, and the desired passing frequency is 14.5GHz.
For 13.4GHz cut off frequency, the width of SIW is a=7.58mm and the equivalent width of the SIW is a_{eff}=7.4mm, calculated by Eq. 8 [9].

$$a_{eff} = a - 1.08\frac{D^2}{p} + 0.1\frac{D^2}{a} \qquad (8)$$

Also for 75dB attenuation at 12GHz spurious the length of SIW (according to Fig.2) is L=50mm. According to the Eq. 7 and Fig.4 the input impedance for this SIW dimensions is 33.5ohms, and regarding to Eq.4, the characteristic impedance of the matching line is 41ohms, with length T=3.6mm. Utilizing optimization algorithms with the HFSS, the length of matching line is changed to 3.2mm for the best performance. In Fig.7, S_{11} and S_{21} of the HPF are shown:

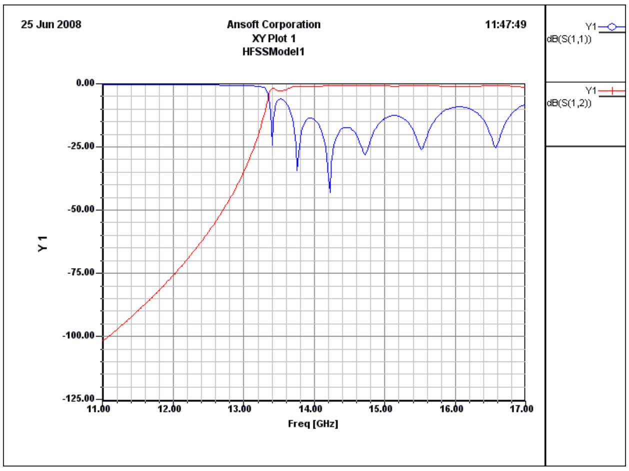

Fig.7. S_{11} and S_{21} of the HPF

As it is illustrated in Fig.7 the stop band's attenuation in 11GHz and 12GHz are 102dB and 75dB respectively, which is confirmed with the theory in Fig.2 with adequate precision. Fig.8 illustrates the Return Loss with greater zoom. The return loss is less than 14dB for frequency bands between 13.7-15GHz and less than 10dB for 13.6-15.9GHz.

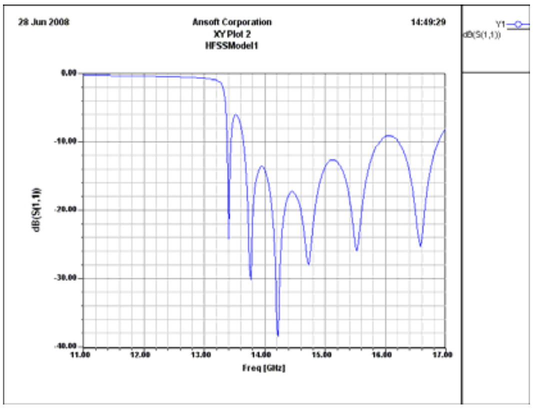

Fig.8. S_{11} of the HPF with greater zoom

5. Test Results

The test result of the designed filter, measured by HP 8722ES Network Analyzer, is shown in Fig.9.

Fig. 9. The test result of the 14GHz HPF, measured by HP 8722ES

The above design has been implemented and tested and obviously, there is a very good similarity between simulation and measurement results. As it is seen, at 12GHz the loss of filter is -75dB and the return loss is less than 14dB between 13.8-14.9GHz and less than 10dB between 13.7-15.5GHz. Certainly, we should consider that power measurement below 80dBm was impossible due to Network Analyzer's dynamic range, and therefore powers less than 80dBm in 12GHz are inside the system's noise floor.

The advantage of the method proposed in this paper to design SIW HPF, in comparison with SIW BPF is as following:

1. The required length for HPF filters is less than BPF filter for equal attenuation.
2. The pass band attenuation for SIW HPF is very less than SIW BPF.
3. Design and fabrication of a SIW HPF is much easier than SIW BPF.
4. The error that appears in the final characteristic of a SIW BPF is rapidly increased when the number of posts is increased, due to the errors in the diameter of posts, their spaces with the walls and the spaces between the posts, while these cases do not exist in the SIW HPF. Indeed, the final characteristic depends only on the length of the SIW and its matching.

To eliminate a 12GHz signal with about 75dB attenuation, the characteristics of a SIW HPF and a 9 stages SIW-BPF (with 13.7-15.5GHz Bandwidth) are compared in the Table.1.

Table1. Advantages of our methods

No.	Parameter	SIW HPF	SIW BPF
1	Filter's Length	50mm	80mm
2	Pass band Attenuation	1dB	4.5dB
3	Test & Fabrication	Easy	Hard
4	Effects of Fabrication Errors	low	high

6. Conclusion

In this paper a novel method for designing HPF, utilizing SIW is presented so that the amount of out band loss is calculated by the amount of loss in evanescent modes of waveguides. Then an innovative method for matching between SIW input and microstrip transmission line for near cut off frequency band is offered. Ultimately, a Ku band HPF has been designed, simulated and implemented, that shows a good agreement between simulation and measurement results. Also, these results confirm that good impedance matching between microstrip line and SIW at input and output ports are achieved.

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Acknowledgments

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