Design for Tapered transitions From Microstrip Lines to Substrat Integrated Waveguide at Ka Band

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Abstract

In this paper, the concept of substrate integrated waveguide (SIW) technology is used to design filter for 30 GHz communication systems. SIW is created in the substrate of RT/Duroid 5880 having relative permittivity ε_r= 2.2 and loss tangent tanφ = 0.0009. Four Via are placed on the century filter the structures of SIW are modeled using and have been optimized in software HFSS (High Frequency Structure Simulator), à transition is designed for a Ka-band transceiver module with a 28.5GHz center frequency, and then the results are verified using another simulation CST Microwave Studio (Computer Simulation Technology). The return loss are less than -18 dB, and -13 dB respectively. The insertion loss is divided equally -1.2 dB and -1.4 respectively.

Keywords: *Transition, Microstrip, Substrat Integrated Wave guide, Filter,Via.*

I. Introduction

Microstrip lines (MSL) are widely used in microwave systems because of its low cost, light weight, and easy integration with other components. Substrate integrated waveguides (SIW), which inherit the advantages from traditional rectangular waveguides without their bulky configuration, aroused recently in low loss and high power planar applications. This article proposed the design and modeling of transitions between these two common structures. Research motives will be described firstly in the next subsection, followed by a literature survey on the proposed MSL to SIW transition structures. Outlines of the following sections in this article will also be given in the end of this section.

II. THEORY OF MSL TO SIW TRANSITIONS

Rectangular waveguides are widely used in microwave systems for its high power handling ability, low radiation loss as well as low electromagnetic interference (EMI) to other circuit components. However they are also known with disadvantages such as bulky volume, heavy weight, high cost, and difficult integration with planar circuits. In addition, high precision process is required at millimeter wave frequencies. As a result, mass production is difficult for systems with rectangular waveguides. Laminated waveguides were first proposed in 1998 [1], where waveguides can be embedded in multilayer printed circuit boards with their side walls replaced by via fences. In 2001, concept of substrate integrated waveguides (SIW) was also proposed [2]. Waveguides embedded in single layer substrates are demonstrated with transitions to CPW and MSL.These kinds of waveguides can be easily integrated with other circuit components by a standard planar circuit fabrication process. Volume and weight are also significantly reduced. The structure and characteristic of SIWs will be introduced firstly in this section, followed by two kinds of excitation structures and calculation of their input resistances. Construction of the equivalent circuit models where the input resistances are associated will also be presented.The theoretical design of transition is discussed in section II. In section III modeling and optimized of transitions (paliers and conical) is described while circuit simulation and results are discussed in section IV by HFSS and then the results are verified using another simulation CST Microwave Studio (Computer Simulation Technology). The work is finally concluded in section V.

II.1. Modal Analysis of SIW

Fig. 1 shows the structure of an SIW, which is composed of the top and bottom metal planes of a substrate and two parallel via fences in the substrate. In order to replace the vertical metal walls, via pitch must be small enough.

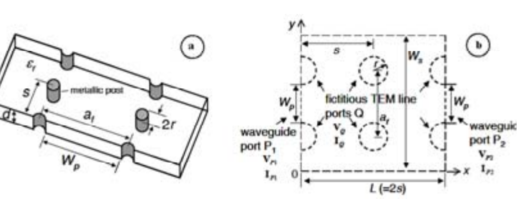

Fig. 1. Structure of SIW straight-line section (a), Planar circuit model for SIW straight-line section (b)

The vias must be shorted to both metal planes to provide vertical current paths, as shown in Fig. 2, otherwise the propagation characteristics of SIW will be significantly degraded.

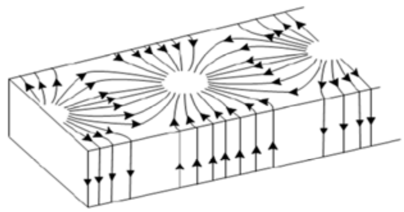

Fig. 2. Structure of RW and surface current for TE10 mode.

Since the vertical metal walls are replaced by via fences for the SIW structures, propagating modes of SIW are very close to, but not exactly the same as, those of the rectangular waveguides. This can be verified by checking the modal surface current patterns. Only patterns with solely vertical current distributed on the side wall survive in SIW. For example, Fig. 2 shows the TE_{10} mode surface current distribution of a rectangular waveguide. The current path will not be cut by the via fences, therefore TE_{10} mode can be supported in an SIW. This holds for all TE_{m0} modes since their current distributions on the side walls are similar.

On the other hand, horizontal components of the surface current exist on the sidewalls for all TM modes and TE_{mn} modes with nonzero n's. These current paths will be cut in SIW structures, which results in radiation. Therefore we can conclude that only TE_{m0} modes exist in SIW structures.

Table 1: Properties of TE_{m0} modes

Property	**TE_{m0} modes**
Generating function	$\emptyset_{m0} = \cos(m\pi x/a)$
Cutoff wave number	$k_{c,m0} = m\pi x/a$
Propagation constant	$\Gamma^2_{m0} = k^2_{c,m0} - k^2,\ k = \omega\sqrt{\mu\varepsilon}$
Magnetic field	$H = H_t + H_z.u_z$
Electric field	$E = E_t + E_z.u_z$

a and b . Because length of vertical wall of SIW becomes height of substrate, "d" and horizontal length a is much longer than height of substrate ($b >> a = d$) the cutoff frequency of the SIW is given by:

$$f_{cmn} = \frac{k_c}{2\pi} = \frac{1}{2\pi\sqrt{\mu\varepsilon}}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2} \Rightarrow f_{10} = \frac{1}{2a\sqrt{\mu\varepsilon}} \quad (1)$$

Ws is the width of the SIW, a_f is the distance between two adjacent cylinders, r is the radius of the cylinder and ae is the width of the equivalent RW. Rectangular waveguide width a can be written in term of a' [3]-[4]:

$$W_s = \frac{2a}{\pi}cot^{-1}\left(\frac{\pi W}{4a_e}ln\frac{W}{4r}\right) \quad (2)$$

There is an empirical formula for equivalent width of SIW in [5].

$$W_{eff} = Ws - \frac{4r^2}{0.95.S} \quad (3)$$

Where W_{eff} is equivalent width, Ws is the SIW width and ($D=2r$) is via diameter and S is distance between adjacent vias. A schematic view of the SIW is shown in Figure 3 [6]. The distance between two arrays (a_f) determines the propagation constant of the fundamental mode, and the via holes parameters (2.r and S) are set to minimize the radiation loss as well as the return loss. In order to insure that the synthesized waveguide section become radiation less or free of leakage loss, parametric effect of p and d have been studied [6]. These studies show that the pitch must be kept small to reduce the loss between adjacent points. The loss tends to decreases as the post gets smaller for a constant ratio d/p, which is conditioned by fabrication process. To obtain good results the ration $r/S \geq 0.25$ and $r/c < 0.2$ must be chosen Fig. 3.

II.3. MSL to SIW Transition with Shorting Via

In HMICs (hybrid microwave integrated circuits), the transition is an important bridge between non-planar and planar circuits. This transition is often very costly, bulky in size and often requires run-time tuning which is a cumbersome task.

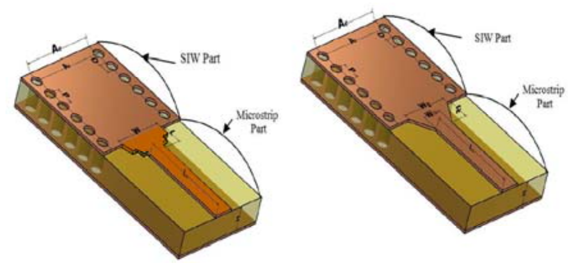

Fig. 4. SIW to Microstrip Transition Design Scheme

The synthesis of a non-planar waveguide in substrate permits the realization of efficient wideband transitions between the synthesized non-planar waveguide and planar circuits such as microstrip and coplanar waveguide (CPW) integrated circuits. With these transitions, the complexity and cost of interconnection between non-planar high-Q circuits and planar circuits are reduced to a minimum [7].

III. Optimized and Modeling

The SIW parameters are designed according to the guidelines provided by Yan et al. in [8] and the procedure to find equivalent rectangular waveguide is shown in (4)–(8). Therefore, the analytical design confirmation is performed for proposed SIW equivalent rectangular waveguide with dielectric permittivity εr = 2.22 and 30GHz frequency of operation. For the dominant mode (TE_{10}) propagation, SIW with its width a_f = 5.6 mm is needed to design that is equivalent to the dielectric filled rectangular waveguide having width Ws = 5.45 mm. This design will have only the dominant mode propagation with cutoff frequency f. Consider

$$X = x_1 + \frac{x_2}{\frac{S}{2r} + \frac{x_1 + x_2 - x_3}{x_3 - x_1}} \quad (4)$$

where the constants x_1, x_2, and x_3 are defined in (5)–(7) and their numerical values are calculated. Consider

$$x_1 = 1.0198 + \frac{0.3465}{\frac{W_{siw}}{S} - 1.0684}, \quad x_1 = 1.2597, \quad (5)$$

$$x_2 = -0.1183 - \frac{1.2729}{\frac{W_{siw}}{S} - 1.2010}, \quad x_2 = -0.646 \quad (6)$$

$$x_3 = 1.0082 - \frac{0.9163}{\frac{W_{siw}}{S} + 0.2052}, \quad x_3 = 0.7682 \quad (7)$$

$$W_s = Xa_f \quad (8)$$
$$W_s = 5.45\ mm$$

The width "Ws" of equivalent rectangular waveguide and the width of SIW "a_f" are related to each other as given in (8). The SIW equivalent rectangular waveguide is illustrated below in Fig. 3. The constant "X" is calculated using (4) and it is used in (8) to calculate SIW equivalent rectangular waveguide width. Consider

IV. Numerical results

IV.1.First Design Example

The specifications for the first design example are:

- Frequency band : 27 to 31 GHz
- Substrat : RT/Duroid 5880(ε_r =2.2,d=0.25mm)

The *S*-parameters of the SIW structure shown in Fig.3-b are computed to demonstrate the usefulness of the present analytical method. The dimensions of the guide are chosen as ε_r =2.2, d=0.25 mm, and r=0.4 mm. The spacing s and the width a_f of the posts are varied in pairs as (s, a_f) = (1.55 mm, 5.60 mm) [9]. A transition is designed for a Ka-band transceiver module w ith a 28.5GHz center frequency, a 27-31GHz transition band for 10dB return loss and 3dB insertion loss in the back-to-back transition structure. The structure is designed on an LTCC substrate with a 2.2 relative dielectric constant and a 0.0009 loss tangent at 30GHz. The feeding MSL is designed for 50Ω characteristic impedance with 0.8mm width and 0.017mm height. The SIW is chosen to be 7.5 mm wide with 0.017 mm thickness.

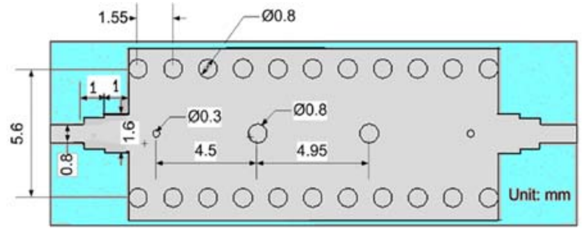

Fig. 5. Geometric configuration of the proposed filter.

Fig. 5 shows the structure of an SIW consisting of the top and bottom metal planes of a substrate and two parallel via fences in the substrate four via are placed on the century filter. The via are composed such that only patterns with vertical current distributed on the side wall can survive in SIW.

IV.1.1. Validation by CST

Response of full-wave simulation is shown in Fig. 6. The center frequency is 28.85 GHz and the 3-dB bandwidth is 2.45 GHz, which is about 5% of the center frequency. The simulated return lose is better than 8 dB and the insertion is 1.8 dB in pass band

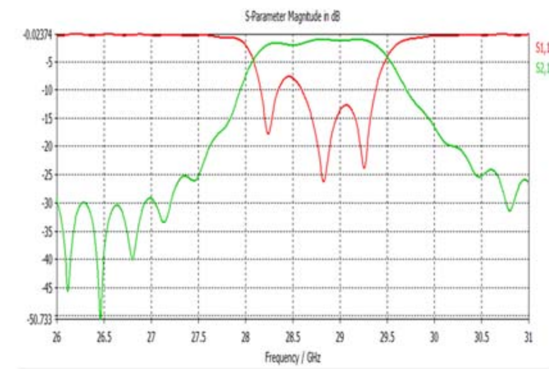

Fig 6: Return and insertion losses for SIW-microstrip tapered transition

From the simulations of the SIW filter by using CST, the Electric *E*-field is observed as shown in Fig. 7.

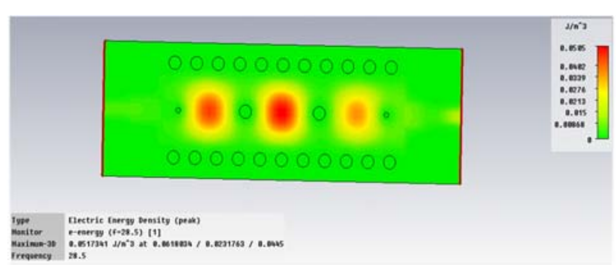

Fig. 7. Simulated Electric Energy Density obtained by CST at 28.5 GHz,

The E-field for TE_{10} mode on microstrip filter SIW at 28.85 GHz are shown in Fig. 7. There is a noted less concentration of the E-field in the filter SIW due to the fact that the micro strip taper is a transmission.

IV.1.2. Validation by HFSS

The layout design on the dual-mode (TE_{10} mode of propagation) of the SIW c filter with EM HFSS fields at 28.55 GHz is shown in Fig.8. The electromagnetic field analysis shows that the distribution of E-fields with two circular propagating is observed in the single SIW cavity filter.

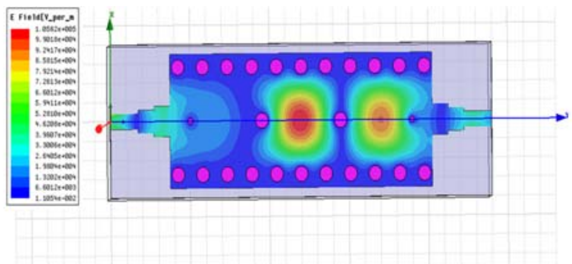

Fig. 8 . Simulated E-field magnitude distributions obtained by HFSS filter

Fig.9. shows the simulation results of the dual-mode SIW cavity filter. The return loss (S_{11}) and insertion loss (S_{21}) of about -8 dB and -1.5 dB with a bandwidth of around 2.45 GHz have been obtained.

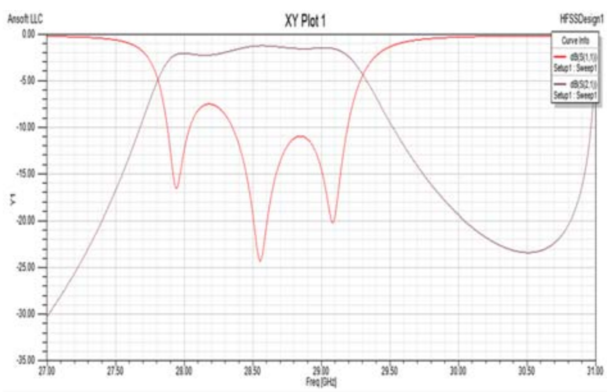

Fig 9. *S*-parameters of the proposed band-pass filter SIW

IV.2 Second Design Example

The geometry of the SIW filter is schematically illustrated in Fig. 10. The development of these types of SIW filters have been reported [10]. In this structure, the SIW filter consists of four via centered in the guide and two microstrip-to-SIW transitions coniques at the input and output, respectively.

Table 2 : Parameters of the filter with matched port configuration, Unit: mm.

Symbol	***Numerical value (mm)***
af	05.80
Wp	05.00
2.r	00.80
2.r'	00.15
S	01.55
Wt	01.60
Lt	02.00
X1	04.50
X2	04.95

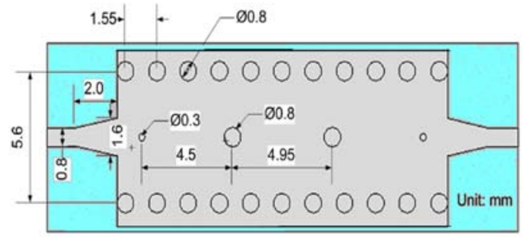

Fig 10. Layout of a three-pole SIW filter on a dielectric substrate with a relative dielectric constant of 2.2 and a

IV.2.1. Validation by CST

The frequency characteristics of the *S*-parameters are shown in Fig.11.

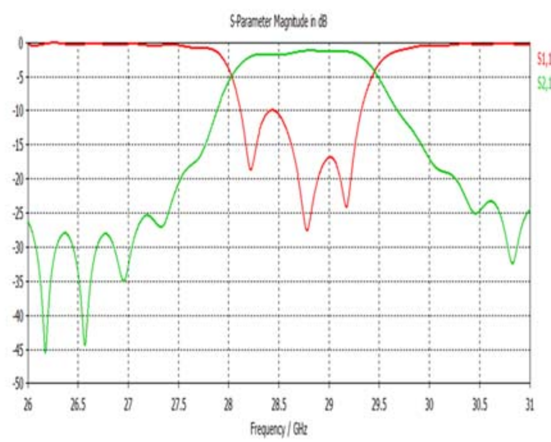

Fig 11. Frequency characteristics of the *S*-parameters

The SIW structure and the E field distribution simulated using the CST at 28.5 GHz are depicted in Fig.12

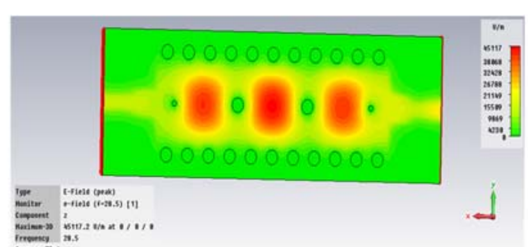

Fig 12 : Structure and Electric field distribution for the SIW straight section

IV.2.2. Validation by HFSS

The *S*-parameters obtained by the present method and the HFSS for these boundary conditions are shown in Fig.13.

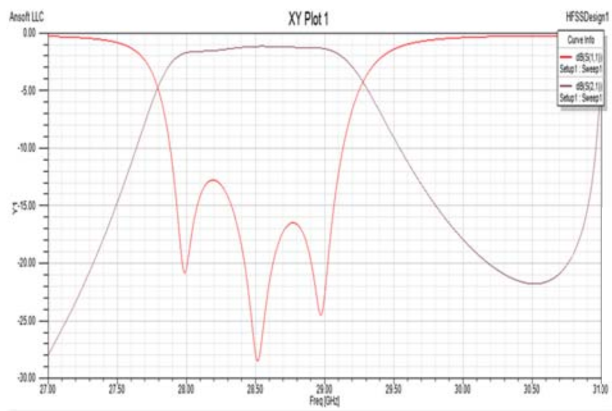

Fig 13 : Structure and field distribution simulated using HFSS (boundary: electric wall).

Fig.14 shows the field distribution calculated using the HFSS (at 28.5 GHz). The TE_{10} mode is incident from the left side of the guide. It is clear that the field is leaking. When the periphery is assigned to electric walls, a resonant mode appears.

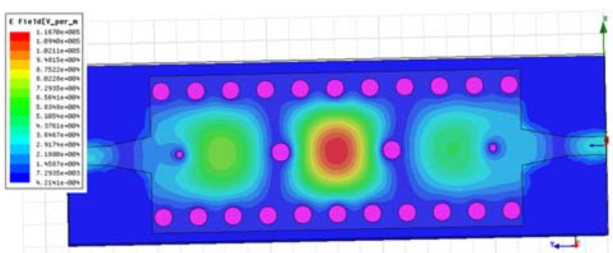

Fig.14: Electric fields distributions in SIW

Fig.14 shows the simulation result by HFSS for a single transition. 2.5 GHz bandwidth is achieved for 13dB return loss. In-band insertion loss is within 1.2 dB. Fig.12. shows the comparison between simulation result by HFSS and result by CST. As shown in the figure, 2.5 GHz the bandwidth for HFSS and CST are achieved for 18 dB and 10 dB return loss respectively.. Insertion loss is better than 1.2dB in the entire transition band.

IV.2.3. Variation Parameter D

Fig.15. shows the frequency responses for different values of *D*

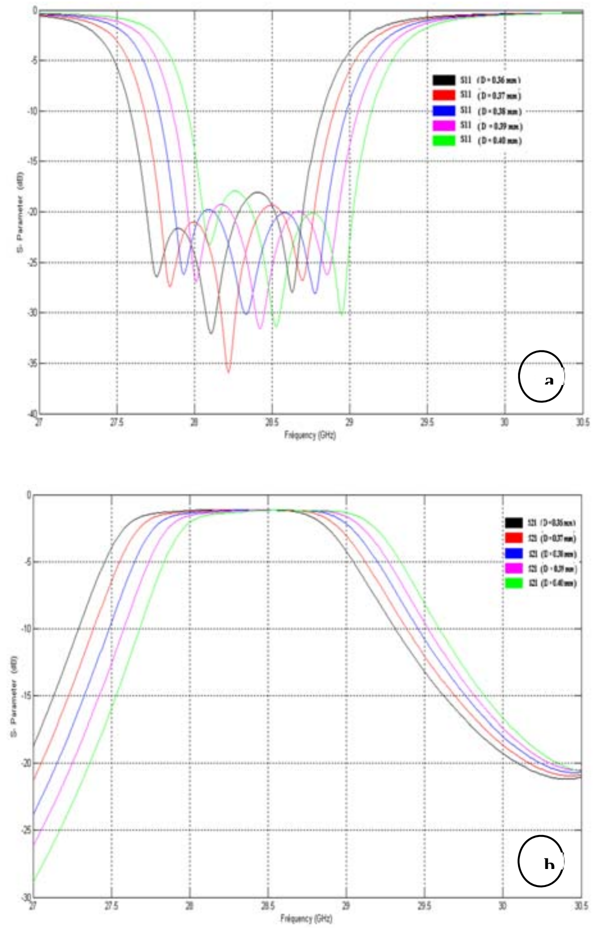

Fig. 15 Frequency response for different values of D=2r
(a) S11 and (b) S21.

V. Conclusion

In this research, this article presents systematic procedures for the design and modeling of transition structures between MSL and SIW. In this paper, the SIW filter based on coupling theory for Ka-band applications has been designed. The simulation process of the structure is done by using HFSS (FEM Method) software and CST (IFT Method). This type of filter is suitable for high density integrated microwave and millimeter wave applications. This paper describes the design of an SIW-to-Microstrip transitions at Ka-band.

The tapered microstrip transmission line is utilized for matching purposes. The key design parameters are discussed along with the governing mathematical equations. The simulated insertion loss of transition is below 1.2 dB and return loss is below 18 dB in the whole band. This
transition is useful for Ka-band substrate integrated circuits and systems and also working band can be extended to millimeter wave frequencies as well.
In the Ka-band designs, simulation by HFSS shows that for a single transition to SIW, **18 %** fractional bandwidths for 15 dB return loss can be obtained.

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