

# A Substrate Integrated Waveguide Leaky Wave Antenna Radiating from a Slot in the Broad Wall

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**Abstract**—This paper presents the application of a substrate integrated waveguide (SIW) for the design of a leaky wave antenna radiating from a slot in the broad wall. The antenna radiates into a beam split into two main lobes and its gain is about 7 dB at 19 GHz. The characteristics and radiation aspects of the antenna are discussed here. The measured antenna characteristics are in good agreement with those predicted by the simulation. Due to the SIW technology, the antenna is suitable for integration into T/X circuits and antenna arrays.

**Index Terms**—Substrate integrated waveguide, flat slotted waveguide, leaky waves, leaky wave antenna, radiation.

## I. INTRODUCTION

Low profile planar antennas have been of great interest to researchers and designers for more than thirty years. The main advantage of these antennas is that they save space and can be located on the external surface of various bodies. They are low cost, can be easily fabricated and are therefore suitable for mass production. Open transmission lines leak power in some frequency bands and can serve as leaky wave antennas, a typical example of low profile antennas. The microstrip leaky wave antenna was reported in [1], and the first slotline leaky wave antenna in [2]. We have successively investigated planar slotted transmission lines in order to find one leaking power only into the half-space. A transmission line suitable for the design of a planar leaky wave antenna (LWA) must have a substrate with finite width and thickness, shielded on the side and bottom walls and having a longitudinal slot on the top. The flat slotted waveguide (FSW) [3] meets all these demands, and served as the transmission line for the design of a leaky wave antenna fed through the coaxial connector [4]. Its later version with an entirely flat top [5] is fed from the rear side.

In recent years, a new concept for the design of microwave and millimeter-wave waveguide structures and components has been proposed [6]. Substrate integrated waveguide (SIW) structures, also known as laminated waveguide or post-wall waveguide structures, use this concept. SIWs are based on the equivalence between well-known metallic waveguide struc-

tures [7] (usually a rectangular waveguide) and waveguide structures on a dielectric substrate using rows of metal posts (vias), see Fig. 1. Cost-effective microwave substrate integrated waveguide circuits could be designed using this approach [8].

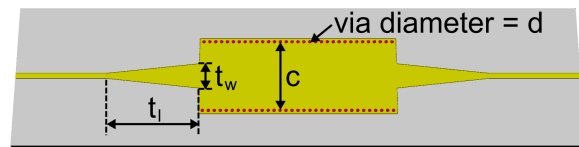


Fig. 1. Substrate integrated waveguide (SIW) with tapered line transitions from microstrip transmission lines.

A substrate integrated waveguide is a possible version of an FSW, that can be simply designed and fabricated [9]. An SIW LWA concept was proposed in [10]. This antenna radiates energy through the SIW side wall with sparsely located shortening vias. Inherent limitations of this structure were exposed. The authors of [10] demonstrated that, in order to design efficient antennas, there is a maximum value related to the substrate permittivity. A half-mode SIW LWA was later proposed, designed and fabricated, as presented in [11]. This antenna radiates from the side edge of a half-mode SIW. Wide bandwidth and a quasi-omnidirectional radiation pattern were obtained. The SIW was used as an LWA able to steer the radiation pattern main beam by changing frequency from nearly backward direction to forward direction [12]. This antenna is based on the CRLH transmission line working in the compensated mode, i.e., without the frequency gap between LH and RH bands.

The SIW LWA radiating from a longitudinal slot etched in the upper wide wall proposed in this paper was designed and fabricated. The antenna is fed through a microstrip line and the first space leaky mode with odd symmetry is excited along the line. The antenna radiates a beam that can be steered in the forward direction by changing the frequency. The SIW

structure is extremely useful for integrating the antenna into any system or antenna array. The proposed antenna is therefore more suited for applications than the LWA referred to in [4] and [5].

## II. DESIGN OF THE ANTENNA

The designed antenna can be viewed at first glance as a conductor-backed slotline with a finitely wide substrate metallized on its sided walls, or a low profile FSW [3]. The cross-section of the line, with proportions and co-ordinates, is shown in Fig. 2. Generally,  $\epsilon_{r2} > \epsilon_{r1}$ . When  $b$  is substantially greater than  $h$  we have a flat waveguide with a slot cut in the waveguide wall parallel to its longitudinal axis. Therefore, the main step in the process of SIW LWA design is based on the design of the FSW LWA [4] substituting the side walls of SIW composed of rows of vias by PEC walls. This FSW line was investigated by the method of moments applied in the spectral domain [3]. For antenna design, the complete dispersion characteristics of this line are needed. They were calculated by the method described in [3], using the APTL program [13].

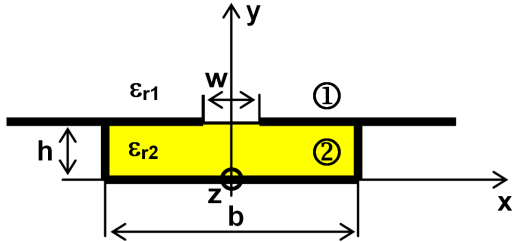
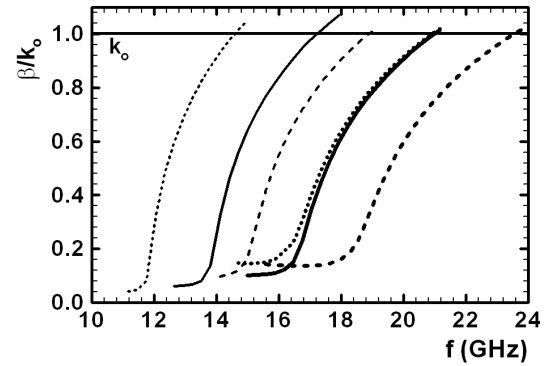


Fig. 2. Cross-section of the flat slotted waveguide.

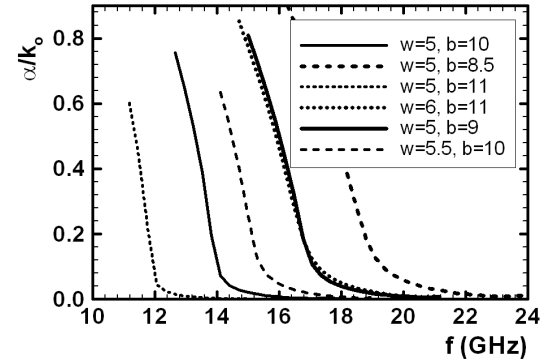
The disadvantage of using waveguides in microwave circuits is the complex transition required to couple a planar structure into a waveguide. In SIW based circuits, it is much easier to realize the transition, since the waveguide structure is integrated into the substrate. A "tapered line" method was presented to realize the transition from a microstrip line to an SIW structure [8]. The goal of such a tapered line is to approximate the field distribution of the  $TE_{10}$  mode with no reflections through the steady expansion of the microstrip line, and thus to achieve excitation of the fundamental mode in the SIW structure.

For our SIW LWA we used the Rogers 4350B substrate with  $h = 0.508$  mm in thickness with relative permittivity  $\epsilon_r = 3.48$  and loss factor  $\tan \delta_e = 0.0037$ . The system of normalized dispersion characteristics of this FSW represented by relative phase constant  $\beta/k_0$  and relative attenuation constant  $\alpha/k_0$  dependence on frequency is shown in Fig. 3,  $k_0$  is free space propagation constant. To get the maximum radiation in the frequency band around 19 GHz we chose according to the characteristics in Fig. 3 FSW width  $b = 9$  mm and slot width  $w = 5$  mm.

Next, the corresponding SIW structure was determined, i.e., the distance between the two rows of vias  $c$  and the mutual distance of the vias in the walls, using a diameter



(a) phase constant  $\beta/k_0$ .



(b) attenuation constant  $\alpha/k_0$ .

Fig. 3. Normalized dispersion characteristics of the first odd space leaky mode of the FSW with parameters defined in the text for various  $b$  and  $w$ .

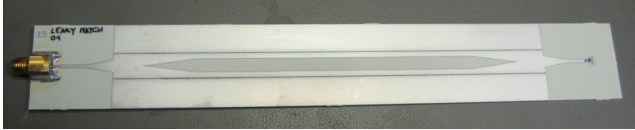
$d = 0.4$  mm of the vias, see Fig. 1. The initial dimensions of the SIW were computed on the basis of analytical formulas for equivalent rectangular waveguides and then optimized numerically using the CST Microwave Studio, resulting in a distance  $c = 9.16$  mm and a mutual distance equal 0.6 mm, both measured from a center to a center of the vias.

The transition was parametrized in its length and width –  $t_l$  and  $t_w$  in Fig. 1 – and optimized for the desired frequency range. The optimization goal was to minimize the reflection coefficient  $|s_{11}|$  and to maximize the transition coefficient  $|s_{21}|$ . The transition from the SIW to the  $50 \Omega$  microstrip line with  $t_l = 9.21$  mm and  $t_w = 4.42$  mm gives the best matching and the maximum power transfer in the specified frequency range of the leaky wave antenna.

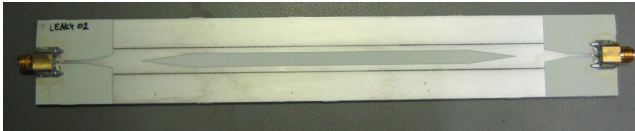
The chosen antenna length was  $l = 100$  mm, measured along the slot with the full width, see Fig. 4, due to the requirement that the amplitude of the space leaky mode at the end of the antenna should be 30% of its magnitude on the feeding tip at mean operating frequency 19 GHz. The final antenna width  $g$  was chosen three times wider than the FSW width  $b$ . The antenna gain might be increased by prolonging the antenna, since with the fabricated structure about 10% of the power coupled to the antenna is lost in the terminating impedance. The radiating slot was terminated by wedges to minimize reflections.

### III. MEASUREMENTS

We fabricated the SIW leaky wave antenna in two types. The first type is a 1-port antenna and the second type is a 2-port antenna – see Fig. 4. Both antenna types contain transitions from the microstrip transmission line to the SIW transmission line at both ends of the antenna. The 1-port antenna is connected at one end to a K-connector and is terminated at the other end with a  $50\Omega$  match. The 2-port antenna is connected at both ends to K-connectors. This enables two-port measurements of the antenna and an evaluation of the radiation losses, as discussed in Subsection III-B.



(a) 1-port SIW leaky wave antenna - with  $50\Omega$  match,



(b) 2-port SIW leaky wave antenna.

Fig. 4. Fabricated SIW leaky wave antennas with K-connectors,  $50\Omega$  match and optimized transition from microstrip transmission line to SIW.

#### A. S-Parameter Measurements

Measurements of the S-parameters of the leaky wave antenna were made using a Rohde & Schwarz R&S<sup>®</sup>ZVA50 vector network analyzer. A comparison of the measured and simulated S-parameters is shown in Fig. 5.

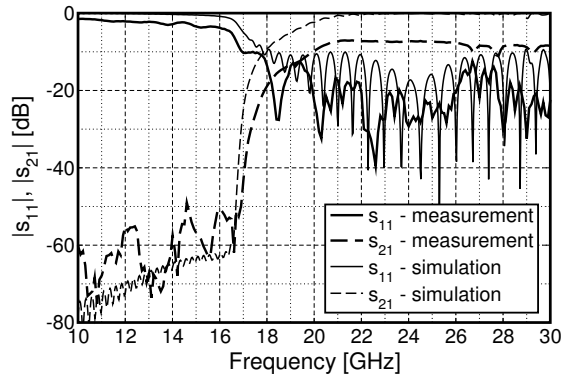


Fig. 5. Measured and simulated S-parameters of the 2-port antenna.

For the simulation, no dielectric losses and no electric losses were considered. This can be seen in the difference between the measured and simulated results, which is approximately 10 dB at  $f = 30$  GHz. The operating frequency band defined by a reflection coefficient lower than -10 dB starts at 18 GHz, and from above it is limited by the existence of the space leaky mode documented in the dispersion characteristics, which is about 20.5 GHz.

#### B. Radiation Losses of a 2-Port SIW Leaky Wave Antenna

The total power loss  $L_s$  of the 2-port antenna is computed from the measured complex S-parameters  $s_{11}$  and  $s_{21}$  using

$$L_s = 1 - |s_{11}|^2 - |s_{21}|^2, \quad (1)$$

which includes all the losses, i.e., the dielectric, conductor and radiation losses. To quantify the power losses we use the quantity  $A_s$ , defined by

$$A_s = 10 \log_{10}(1 - L_s) = 10 \log_{10}(|s_{11}|^2 + |s_{21}|^2) \quad [\text{dB}], \quad (2)$$

with  $L_s$  given by (1). The quantity  $A_s$  given in (2) is a measure of the relative available power that is not lost in the antenna, e.g.,  $A_s = \text{dB}(|s_{21}|)$  for a perfectly matched non-radiating lossless transmission line.

The  $A_s$  computed from the measured S-parameters given in Fig. 5 is shown in Fig. 6. For  $10 \text{ GHz} \leq f < 16.5 \text{ GHz}$  and  $f > 21 \text{ GHz}$  the dielectric and conductor losses dominate with no significant radiation. However, for  $16.5 \text{ GHz} \leq f \leq 21 \text{ GHz}$  the radiation losses dominate, which can be seen in the highly increased power loss  $-A_s$ .

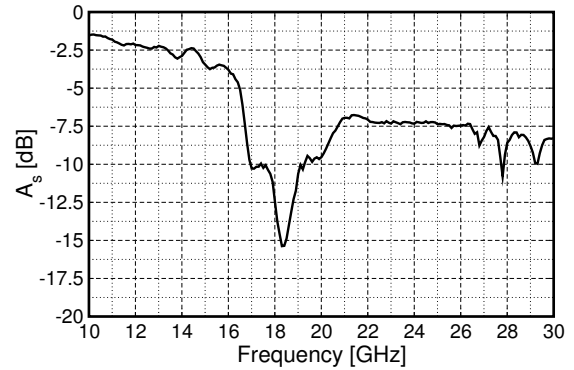


Fig. 6. Total losses of a 2-Port SIW leaky wave antenna.

#### C. Radiation Pattern Measurements

We measured and computed the electric field component  $E_\theta(\theta)$  for  $\varphi = 0^\circ$  and  $E_\theta(\varphi)$  for  $\theta = 90^\circ$ . The orientation of the antenna in the spherical co-ordinate system and the computed 3-D normalized radiation pattern at  $f = 19 \text{ GHz}$  are shown in Fig. 7. A comparison of the measured and computed normalized radiation diagrams is shown in Fig. 8 and Fig. 9.

The antenna radiates a beam steered in the vertical plane from  $48^\circ$  at 18 GHz to  $15^\circ$  at 21 GHz. The full width at half power of the main antenna beam in the vertical plane varies from around  $13^\circ$  at 18 GHz to  $21^\circ$  at 21 GHz. The side lobes are about 10 dB below the level of the maximum radiation. The radiation pattern in the horizontal plane is remarkably wider than that in the vertical plane, and at lower frequencies it is split into two sub-beams. The finite dimensions of the antenna top conductor are responsible for radiation below the antenna plane. This portion of the radiation fades away with widening of this conductor. The measured antenna gain was 7.5 dB at 18 GHz, 7.2 dB at 19 GHz, 6.5 dB at 20 GHz and 2 dB at 21 GHz.

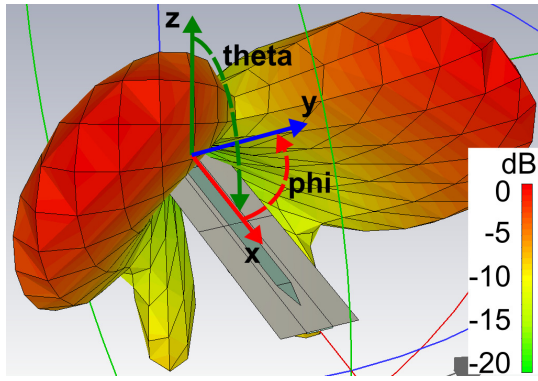


Fig. 7. Simulated radiation pattern  $-dB(|E_\theta|)$  at  $f = 19$  GHz normalized to 0 dB.

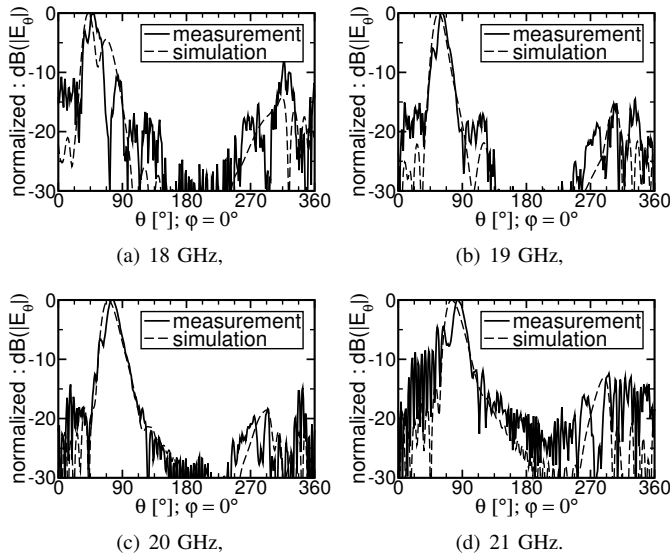


Fig. 8. Measured and simulated radiation diagrams for  $\varphi = 0^\circ$ .

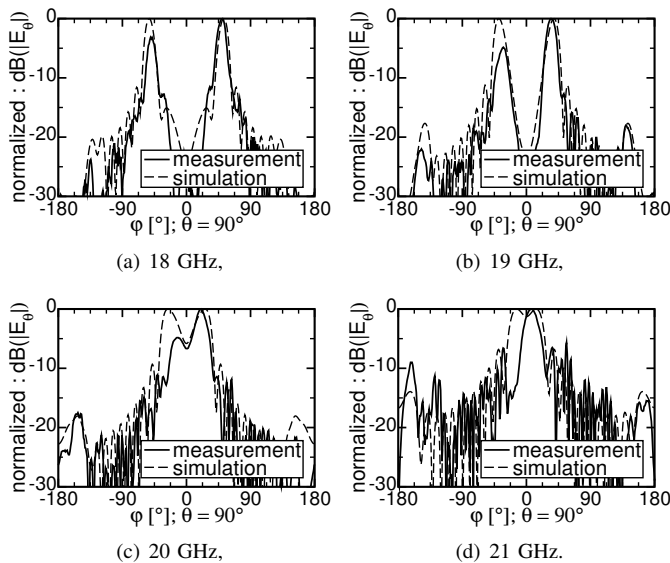


Fig. 9. Measured and simulated radiation diagrams for  $\theta = 90^\circ$ .

#### IV. CONCLUSIONS

We have designed, fabricated and measured a substrate integrated waveguide leaky wave antenna radiating from a slot in the broad wall parallel to its axis. The initial dimensions of the SIW and of the slot were obtained from analytical expressions for equivalent rectangular waveguides and from the dispersion characteristics. The SIW, the antenna layout and the feeding circuit were then optimized using the CST Microwave Studio.

The measured S-parameters and antenna characteristics fit well with the calculated results. The operating frequency band is from 18 GHz to 21 GHz and the gain is about 7 dB at 19 GHz. The antenna radiates a beam steered in the vertical plane from  $48^\circ$  at 18 GHz to  $15^\circ$  at 21 GHz. The radiation pattern in the horizontal plane is remarkably wider than that in the vertical plane, and it is split into two sub-beams.

The designed antenna is well suited for being integrated into antenna arrays and all transmitting or receiving systems. A normal PCB process is applied to fabricate this antenna.

#### ACKNOWLEDGEMENT

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