

Review of substrate-integrated waveguide circuits and antennas

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Abstract: Substrate-integrated waveguide (SIW) technology represents an emerging and very promising candidate for the development of circuits and components operating in the microwave and millimetre-wave region. SIW structures are generally fabricated by using two rows of conducting cylinders or slots embedded in a dielectric substrate that connects two parallel metal plates, and permit the implementation of classical rectangular waveguide components in planar form, along with printed circuitry, active devices and antennas. This study aims to provide an overview of the recent advances in the modelling, design and technological implementation of SIW structures and components.

1 Introduction

Wireless components and systems have received increased interest in recent years, as new applications for millimetre-waves (mm-waves) are being introduced and developed. In fact, a variety of applications have been recently proposed in the frequency range over 60–94 GHz, including wireless networks [1], automotive radars [2], imaging sensors [3] and biomedical devices [4].

The deployment of mm-wave technologies is critical for the evolution of wireless systems as broadband and high-resolution techniques are naturally supported by the use of mm-waves. In most of these systems, the success mainly depends on the availability of a cost-effective technology, suitable for the mass-production of components and systems. It is expected that high-density integration techniques, combined with a low-cost fabrication process, should be able to offer widespread solutions for mm-wave commercial applications.

The core of these systems is related to the active part, which includes components such as local oscillators, mixers and possibly low-noise amplifiers among others. Nowadays, such components can be integrated in the form of chip-sets at reasonably low cost. Several semiconductor companies are currently working towards the development of chip-sets operating at 60 GHz or even at higher frequencies [5]. Nevertheless, other components are needed in mm-wave systems, which cannot be conveniently integrated in the chip-set, because either they are too large or the required performance cannot be achieved by integrated components (such as antennas, selective filters and power amplifiers). These additional components could be simply considered as the package that embeds the chip-set, but they actually

represent a significant portion of the system. With the concept of the system in package (SiP) [6], one or more chip-sets are combined in a single package with the other components, which are fabricated with different technologies. At low frequencies these components are typically fabricated in planar technology (microstrip or coplanar waveguides); at frequencies higher than 30 GHz, however, transmission losses and radiation prevent the use of microstrip or coplanar waveguides and other technological solutions have to be identified. We can therefore conclude that the successful development of mm-wave wireless systems requires the definition of a platform for implementing all these components with a high performance, low-cost and reliable technology.

A promising candidate for developing this platform is substrate-integrated waveguide (SIW) technology [7–11]. SIWs are integrated waveguide-like structures fabricated by using two rows of conducting cylinders or slots embedded in a dielectric substrate that electrically connect two parallel metal plates (Fig. 1). In this way, the non-planar rectangular waveguide can be made in planar form, compatible with existing planar processing techniques (e.g. standard printed circuit board (PCB) or low-temperature co-fired ceramic (LTCC) technology). SIW structures exhibit propagation characteristics similar to the ones of classical rectangular waveguides, including the field pattern and the dispersion characteristics. Moreover, SIW structures preserve most of the advantages of conventional metallic waveguides, namely high quality-factor and high power-handling capability with self-consistent electrical shielding. The most significant advantage of SIW technology is the possibility to integrate all the components on the same substrate, including passive components, active elements and even

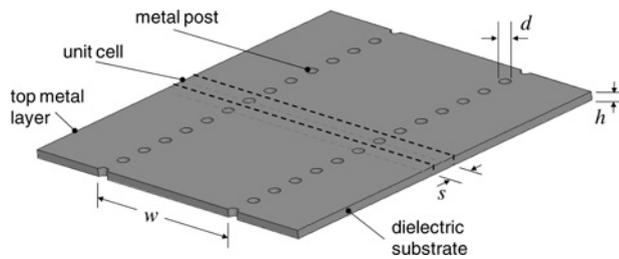


Fig. 1 Geometry of a SIW

antennas. Moreover, there is the possibility to mount one or more chip-sets on the same substrate. There is no need for transitions between elements fabricated with different technologies, thus reducing losses and parasitics. In this way, the concept of SiP can be extended to the system-on-substrate (SoS) [12]. SoS represents the ideal platform for developing cost-effective, easy-to-fabricate and high-performance mm-wave systems.

First developed with the name of post-wall waveguide [7] or laminated waveguide [8] for feeding networks in antenna arrays, the SIW technology has been applied to several microwave components, including post and cavity filters [9, 13], directional couplers [14], oscillators [15, 16], power amplifiers [17, 18], slot array and leaky antennas [19, 20], six-port circuits [21] and circulators [22]. From the technical literature, it is seen that most of the classical waveguide components have been implemented in SIW technology. Nevertheless, most of these SIW components operate in the frequency range up to 30 GHz, with a few exceptions at higher frequency [23–26]. This phenomenon is not due to physical limitations of SIW components, but to the increased technological difficulties encountered in designing and manufacturing SIW structures over the mm-wave range, namely, miniaturised dimensions, higher losses and the need to carefully select the material. Therefore current and future research activity aims to overcome these difficulties, in order to use SIW technology as an interconnection and integration platform for mm-wave systems.

2 SIW structures

2.1 Operation principles

SIW structures exhibit propagation characteristics similar to those of rectangular metallic waveguides, provided that the metallic vias are closely spaced and radiation leakage can be neglected (Fig. 2). More specifically, SIW modes practically coincide with a subset of the guided modes of the rectangular waveguide, namely with the TE_{n0} modes, with $n = 1, 2, \dots$. TM modes are not supported by SIW,

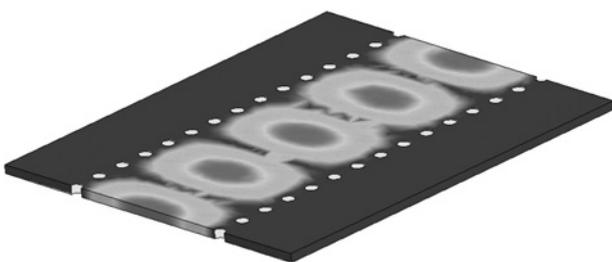


Fig. 2 Plot of the amplitude of the electric modal vector of the fundamental SIW mode

due to the gaps between metal vias: in fact, transverse magnetic fields determine longitudinal surface currents, which are subject to strong radiation due to the presence of the gaps [1]. In particular, the fundamental mode is similar to the TE_{10} mode of a rectangular waveguide (Fig. 2), with vertical electric current density on the side walls.

Owing to this similarity between SIW and rectangular waveguide, empirical relations have been obtained between the geometrical dimensions of the SIW and the effective width w_{eff} of the rectangular waveguide with the same propagation characteristics. These relations allow for a preliminary dimensioning and design of SIW components, without any need of full-wave analysis tools. One of the most popular relations was derived in [27]

$$w_{\text{eff}} = w - \frac{d^2}{0.95s} \quad (1)$$

where d is the diameter of the metal vias, w represents their transverse spacing and s is their longitudinal spacing (Fig. 1).

Relation (1) was subsequently refined in [10]

$$w_{\text{eff}} = w - 1.08 \frac{d^2}{s} + 0.1 \frac{d^2}{w} \quad (2)$$

Another relation was proposed in [28]

$$w = \frac{2w_{\text{eff}}}{\pi} \cot^{-1} \left(\frac{\pi s}{4w_{\text{eff}}} \ln \frac{s}{2d} \right) \quad (3)$$

A more rigorous determination of the propagation characteristics of SIW structures can be based on full-wave analysis tools, either commercial software (e.g. Ansoft high-frequency simulation study or computer simulated technology Microwave Studio) or in-house developed electromagnetic simulators: among them, the most common techniques are based on the finite-difference time domain method or the finite-difference frequency domain [29], the boundary integral-resonant mode expansion (BI-RME) method [27], the method of lines [30] and the transverse resonance method [11]. In all cases, the determination of the propagation and attenuation constants of SIW structures is based on the analysis of the single unit cell (Fig. 1), thus significantly reducing the computational effort.

2.2 Loss mechanisms

A key issue in the design of SIW structures is related to loss minimisation, which is particularly critical when operating at mm-wave frequencies. Three major mechanisms of loss need to be considered in the design of SIW structures [31, 32]: they are conductor losses (due to the finite conductivity of metal walls), dielectric losses (due to the lossy dielectric material) and possibly radiation losses (due to the energy leakage through the gaps). The behaviour of conductor and dielectric losses is similar to the corresponding losses in rectangular waveguides filled with a dielectric medium, and the classical equations can be effectively applied. It transpires that conductor losses can be significantly reduced by increasing the substrate thickness, being the corresponding attenuation constant almost proportional to the inverse of substrate thickness h . The other geometrical dimensions of the SIW exhibit a negligible effect on conductor losses. Conversely, dielectric losses depend only on the dielectric material and not on the geometry of the

SIW structure, and therefore they can be reduced only by using a better dielectric substrate. Finally, radiation losses can be kept reasonably small if $s/d < 2.5$, with $s/d = 2$ being the recommended value. In fact, when the spacing s is small and the diameter d of the metal vias is large, the gap between the metal vias is small, thus approaching the condition of continuous metal wall and minimising the radiation leakage. Generally speaking, the contribution of dielectric losses is predominant at mm-wave frequencies, when using typical substrate thickness and commercial dielectric material [32].

The insertion loss usually calculated for SIW structure, which accounts for conductor, dielectric and radiation losses, can be significantly increased by the effect of surface roughness in conductors. Analytical models of losses due to the surface roughness have been developed for classical waveguides, and are incorporated in commercial electromagnetic simulators. Recently, this issue has been carefully investigated through numerical and experimental studies in the case of microstrip transmission lines [33], whereas no publications have been reported yet in the case of SIW structures.

It is also particularly relevant to compare losses in SIW structures and in other traditional planar structures, for example, microstrip or coplanar lines. A systematic comparison of SIW and microstrip components is not easy, because SIW circuits are usually implemented on a thick substrate with low dielectric constant (which is not suitable for the implementation of microstrip circuits), with the aim of minimising conductor losses. In principle, microstrip component losses could also be mitigated by increasing the substrate thickness; in practice, however, this cannot be exploited due to the unacceptable increase in radiation loss and excitation of surface waves. A detailed comparison of losses in SIW structures, microstrip lines and coplanar waveguides is reported in [34]: it is seen that SIW structures can guarantee comparable or lower losses, compared to traditional planar transmission lines.

2.3 Size and bandwidth

Another important topic to be accounted for in the design of SIW structures is the performance in terms of size and operation bandwidth. In fact, similar to rectangular

waveguides, SIW structures are limited in compactness and bandwidth. The width of the SIW determines the cutoff frequency of the fundamental mode (with a reduction of factor $\epsilon_r^{-1/2}$ over hollow rectangular waveguides). The operation bandwidth is limited to one octave (from the cutoff frequency f_1 of the TE₁₀ mode to cutoff frequency $f_2 = 2f_1$ of the TE₂₀ mode), corresponding to the mono-mode bandwidth of the waveguide.

Different waveguide topologies have recently been proposed to improve the compactness of SIW structures (Fig. 3). The substrate-integrated folded waveguide (SIFW) was proposed in [35] (Fig. 3a): a metal septum permits folding of the waveguide, thus reducing the size by a factor of more than two at the cost of slightly larger losses. The half-mode substrate-integrated waveguide (HMSIW) was introduced in [36] (Fig. 3b): based on the approximation of the vertical cut of the waveguide as a virtual magnetic wall, it permits a size reduction of nearly 50%. A combination of the two techniques was also proposed [37], resulting in the folded half-mode substrate-integrated waveguide (FHMSIW), which leads to a further size reduction.

To improve the bandwidth performance, some waveguide configurations have been developed. The substrate-integrated slab waveguide (SISW) was proposed in [38] (Fig. 3c): it consists of an SIW where the dielectric medium is periodically perforated with air-filled holes, located in the lateral portion of the waveguide. This approach enabled the design of a waveguide with a mono-modal band from 7.5 to 18 GHz (with 40% bandwidth enhancement). The implementation of a ridge waveguide in SIW technology was proposed in [39], where the ridge was implemented through a row of thin, partial-height metal posts located in the centre of the longer side of the waveguide. This structure allowed coverage of the mono-modal band from 4.9 to 13.39 GHz, thus achieving a 73% bandwidth enhancement. A significant improvement in the performance of the ridge SIW was introduced in [40] (Fig. 3d): the modified ridge SIW is based on a row of partial height metal cylinders located in the broad side of an SIW and connected at their bottom with a metal strip. A modified ridge SIW covering the frequency band 6.8–25 GHz was designed and fabricated (with 168% bandwidth enhancement). A further improvement has been obtained with the ridge SISW, where air-filled holes have been added. In this case, a ridge SISW covering the entire

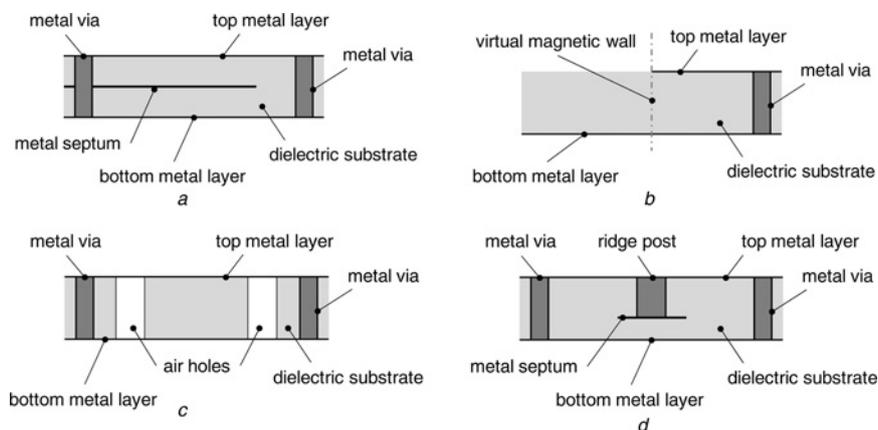


Fig. 3 Different topologies of SIWs

- a Substrate-integrated folded waveguide
- b Half-mode SIW
- c Substrate-integrated slab waveguide
- d Ridge SIW

frequency band 7.1–30.7 GHz was designed and fabricated (with 232% bandwidth enhancement). This last configuration allows obtaining compact and broadband interconnects, which are 40% smaller than a conventional SIW and exhibit a three times broader bandwidth, and are suitable to fabrication by using standard PCB or LTCC technology.

3 SIW passive circuits

Owing to the similarity between SIW structures and classical rectangular waveguides, most of the planar (H-plane) waveguide components have been implemented in SIW technology. This solution usually permits a substantial reduction in size and in weight of components if compared to classical waveguide; moreover, losses of SIW components are lower than in the corresponding microstrip devices and there are no radiation and packaging problems. SIW components are a good compromise between air-filled rectangular waveguide and microstrip line, especially in the mm-wave range, where microstrip is too lossy to design high Q components.

3.1 Filters and couplers

Among the passive components, filters have received a particular attention. A variety of filter topologies have been proposed (Fig. 4): among them, a filter with inductive-post operating at 28 GHz [9] and a filter with irises operating at 60 GHz [24] were designed and fabricated. Subsequently, cavity filters with circular [13] and rectangular cavities [41] were developed: they permit a better design flexibility and exhibit higher selectivity, thanks to the cross-coupling that introduces transmission zeros. A multi-layered structure was adopted in [42]: the use of a two-layer substrate permitted the design of an elliptic filter with four cavities operating in C band. Compact and super-wide band-pass filters were presented in [43]: owing to the use of an electromagnetic band-gap structure in the ground plane, a band-pass filter covering the frequency range 8.5–16.5 GHz was designed and tested. While most of these filters operate in the microwave range, filters operating at 60 GHz [24] and up to 180 GHz [25] were also proposed.

Besides filters, several other passive components have been developed in SIW technology. Among them, two configurations of directional couplers were proposed: the former, based on two adjacent SIW with apertures in the common wall, was used to design 3, 6 and 10-dB couplers

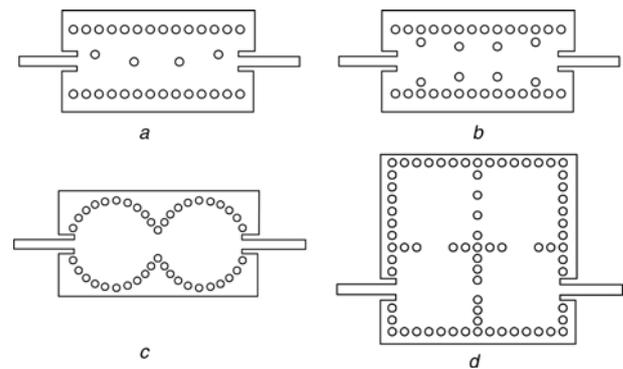


Fig. 4 Different classes of SIW filters

- a Filter with inductive-post
- b Filter with iris windows
- c Filter with circular cavities
- d Filter with rectangular cavities and cross-coupling

[44]; the second configuration presents a cruciform shape, and was adopted to design a super-compact 3-dB directional coupler [45]. Planar SIW diplexers operating at 5 and 25 GHz were proposed [46, 47]. A magic-T [48], six-port circuits [21, 26] and circulators [22] were also implemented and experimentally verified.

3.2 Transitions

The transitions between planar transmission lines and SIW structures represent another important element related to SIW components. Several broadband transitions between microstrip or coplanar waveguide and SIW have been developed [49, 50] (Fig. 5). In particular, microstrip-to-SIW transitions are typically based on a simple taper (Fig. 5a), provided that the microstrip and the SIW structure are integrated on the same substrate [49]. Recently, design equations have been proposed for the fast implementation of microstrip-to-SIW transitions [51]. Microstrip-to-SIW transitions in a multi-layer substrate environment have been proposed [52], to connect a microstrip implemented in a thin substrate with a thicker SIW structure. The use of thick substrates allows for reduced conductor losses in SIW structures. Conversely, two solutions have been proposed for coplanar-to-SIW transitions. The first solution makes use of a current probe (Fig. 5b): the current flowing through the probe generates a magnetic field, which matches with the magnetic field inside the SIW structure [50]. Another possible configuration was proposed in [53] and consists of

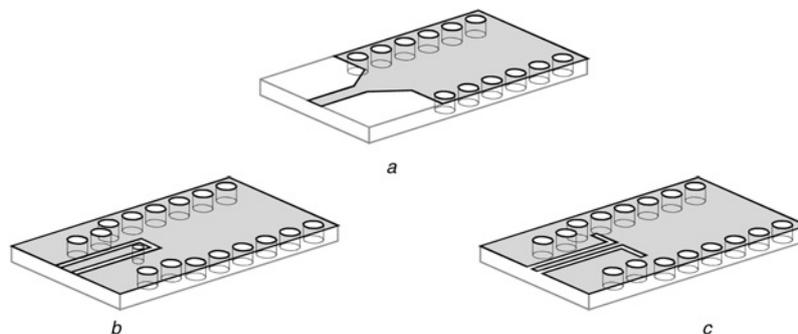


Fig. 5 Transitions between printed transmission lines and SIW

- a Microstrip-to-SIW transition, based on a taper
- b Coplanar-to-SIW transition, based on a current probe
- c Coplanar-to-SIW transition, based on a 90° bend

a coplanar waveguide with a 90° bend on each slot inside the SIW structure (Fig. 5c). It is noted that using coplanar waveguides may be convenient when thick substrates are adopted to reduce conductor losses, and consequently the use of microstrip lines is not possible. Finally, transitions between air-filled waveguide and SIW structure have also been proposed [54]: this transition is based on a radial probe inserted into a tapered metallic waveguide.

3.3 Electromagnetic modelling

The development of SIW technology has stimulated the application of several numerical techniques to the electromagnetic modelling and design of SIW components. Among them, full-wave numerical techniques have been widely adopted. Both commercial electromagnetic software and specifically developed numerical techniques have been used. Electromagnetic codes based on integral-equation, finite-element or finite-difference methods have been implemented [29, 31, 55]. In most cases, these methods deal not only with metallic posts, but also with inhomogeneous substrates (for instance, with air-filled holes as in SISW).

A particularly efficient numerical technique for the modelling of arbitrarily shaped SIW components is based on the BI-RME method [31]. The BI-RME method, originally developed for the modelling of classical waveguide components, allows for characterising SIW components through their generalised admittance matrix Y expressed in the form of a pole expansion in the frequency domain, relating modal currents and voltages of the port modes. Owing to this peculiarity, the BI-RME method is used to determine in one shot the wideband expression of the frequency response of SIW components, thus avoiding repeated frequency-by-frequency electromagnetic analyses. Consequently, the BI-RME modelling of SIW components typically requires a few seconds on a standard personal computer.

Another significant advantage of the BI-RME method is the possibility to directly determine equivalent circuit models of SIW discontinuities [56]. Owing to the particular representation of the admittance matrix, each term of the Y matrix can be represented as the parallel of an inductance, a capacitance and a number of LC-series resonators (Fig. 6). The most important application of this method is the determination of parametric multi-modal equivalent circuit models, where the values of the lumped elements depend on the geometrical dimensions of the component [56]. In

fact, once a library of equivalent circuit models is available, the direct synthesis of a component can be performed in a short time by using conventional circuit computer-aided design tools, with no need of electromagnetic full-wave analysis codes. Equivalent circuit models of SIW discontinuities have been adopted, in conjunction with a space-mapping optimisation technique, for the fast design of SIW filters [57].

4 SIW active circuits

The implementation of active circuits using SIW technology has received less attention compared to the reported developments regarding passive circuits. Nonetheless, the field of active circuits without doubt opens numerous new design possibilities for SIW technology towards a complete SoS integration. Essentially, the design and optimisation of active circuits consists of embedding active devices in passive SIW circuits and interconnects thus utilising the advantages of the technology such as, for example, low loss, high isolation and compact size to achieve good performance with low-cost fabrication techniques. Recent developments in oscillators, mixers and amplifiers are provided in the following subsections.

4.1 Oscillators

SIW technology is particularly suitable for oscillator design in terms of designing compact cavity resonators. In the microwave and millimetre frequency range the oscillator topologies are either of (a) feedback type, where a resonator is embedded in the feedback loop of an amplifier circuit, or (b) reflection type, where a resonator is coupled to an active device port presenting a negative resistance. In Fig. 7, illustrative block diagrams of various oscillator topologies using SIW resonators are shown.

The first reported that SIW oscillator used a rectangular SIW resonator appropriately placed in the feedback path between the input and output nodes of an amplifier circuit designed using the Agilent ATF36077 pHEMT transistor [58]. The obtained oscillation frequency was 12.02 GHz with an output power of 0 dBm. The phase noise of the oscillator was measured to be -105 dBc/Hz at an offset of 1 MHz from the carrier. Linear analysis was used to design the oscillator. The feedback transmission lines and various input and output lines of the oscillator circuit were fabricated in microstrip technology and were connected to

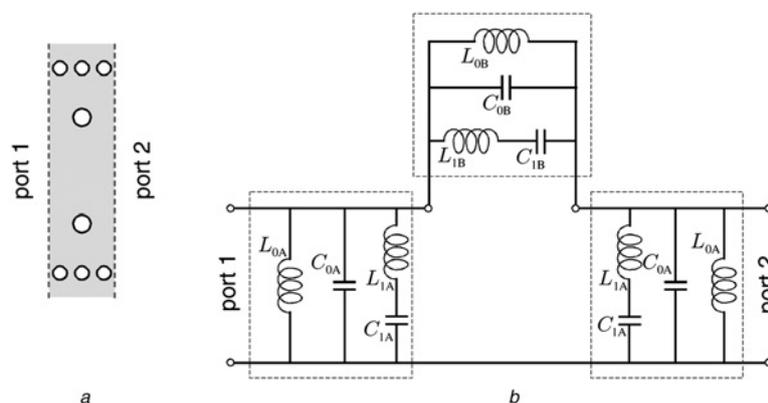


Fig. 6 Equivalent circuit model of an iris-type discontinuity in SIW technology

a Geometry of the discontinuity

b Equivalent circuit model of the discontinuity, derived from the BI-RME method

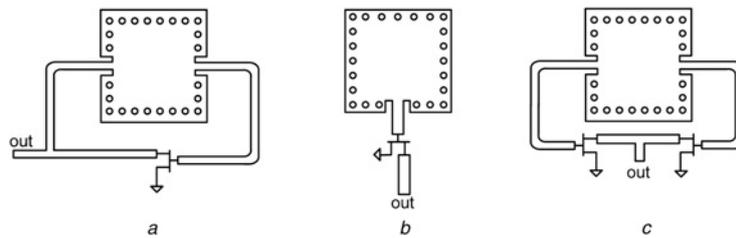


Fig. 7 Oscillator circuits in SIW technology

- a Feedback oscillator
- b Reflection oscillator
- c Push–push reflection oscillator

the SIW cavity using appropriately designed transitions. The design first demonstrated the capability of utilising SIW technology to fabricate low-cost, compact, high Q resonator structures for microwave oscillator circuits.

A reflection type oscillator at 35.259 GHz based on a Gunn diode was reported in [59]. A rectangular SIW cavity was also used in this design, and the active device is integrated into the edge of the cavity by removing a number of metallised via holes, and placing the diode vertically in the substrate inside an un-metallised via hole with a diameter size large enough to accommodate the diode. A DC-bias network, including a low-pass filter composed of high-low impedance microstrip line sections, was implemented externally to the cavity resonator in order to provide the bias. The oscillation frequency was predicted using linear simulation. The oscillator demonstrated an output power of 15.7 dBm, and a measured phase noise performance of -91.2 dBc/Hz at an offset of 100 KHz from the carrier frequency. The DC to radio frequency (RF) efficiency of the oscillator is 0.74%, as the oscillator dissipates 1000 mA from a 5 V supply. However, this work demonstrates the capability of integrating active devices into the SIW cavity as well as the suitability of this technology in order to manufacture low-cost frequency sources in the millimetre wave range. The authors also published a very similar oscillator at [60], with the difference that the Gunn diode was placed near the centre of the cavity and it was biased using a bond wire which connected the diode anode to a pad forming a DC-bias network implemented externally to the cavity resonator. In a more recent publication, the authors of [61] proposed a Gunn oscillator with frequency tuning capability by integrating a varactor diode in addition to a Gunn diode inside a rectangular SIW cavity. The varactor diode is equivalent to a tunable capacitive post within the cavity and consequently it permitted the authors to vary the resonance frequency of their cavity by 3.16% at a centre frequency of 36.271 GHz. The output power varied from 9.3 to 11.3 dBm over the tuning range. The oscillator had a low conversion efficiency of 0.55% at 11.3 dBm output power, while it dissipated 2.448 W from a 4.8 V supply. The phase noise of the oscillator was measured to be -102.1 dBc/Hz at 1 MHz offset at 36 GHz. The oscillator output was obtained by tapping the signal of the resonant cavity using properly designed SIW-to-microstrip transition.

Both previous oscillator designs relied on linear simulation techniques in order to estimate the resonance frequency of the oscillators. In [58] the oscillator design consisted of adjusting the electrical length of the feedback path in order to obtain a phase shift that is an integral multiple of 2π over the loop while maintaining a positive closed-loop gain, and therefore guaranteeing the oscillation condition [62]. The reflection oscillators of [59–61] were also designed using linear

simulation. The oscillation frequency was predicted by ensuring that the parallel combination of the resonator and diode admittance had a negative real part and zero imaginary part [63, 64]. Metallised via holes inside the SIW cavity were used to control the admittance presented to the Gunn diode and therefore set the oscillation frequency. Linear simulation techniques are highly intuitive and provide a means for estimating and investigating potential parasitic oscillations in undesired frequency bands. In addition, they do not require complicated non-linear models for the active devices and can be applied using easily obtained S-parameter measurements of the passive resonator and the active device for the selected bias conditions. However, they only provide an estimate of the oscillation frequency due to the fact that they rely on small signal operating conditions of the active device, and they are unable to estimate the output power or the harmonic content of the oscillator.

When a non-linear model of the active device is available, non-linear simulation techniques such as harmonic balance provide accurate estimates of both the oscillation frequency and amplitude. Harmonic balance simulation presents an inherent difficulty in simulating oscillator circuits due to the fact that the frequency of oscillation which represents the basis of the Fourier expansion used in the harmonic balance analysis is unknown a priori. This difficulty can be circumvented by introducing an additional (complex) equation to the harmonic balance formulation corresponding to the total admittance at some frequency looking into an arbitrary circuit node must be equal to zero for the circuit to oscillate [64]. The inclusion of this requirement in the harmonic balance simulation allows one to introduce two additional unknowns, which typically are the oscillation frequency and amplitude at the selected node. This observation was first proposed by [65] and it can be easily implemented in commercial simulators as demonstrated by [66]. The use of harmonic balance analysis permits the designer to optimise directly the oscillator amplitude and frequency and its accuracy is limited only by the accuracy of the available device models.

Two X-band oscillators using a rectangular SIW cavity resonator were also reported in [67]. In this work, the oscillators were designed by harmonic balance simulation where the S-parameters of the cavities were imported from an electromagnetic simulation. The proposed designs consisted of a feedback and a reflection oscillator implemented in 0.508 mm thick Rogers 4003 substrate and used the NEC NE3509 HJFET device. The feedback oscillator had a resonance frequency of 12.64 GHz, output power of 4.5 dBm with a DC to RF conversion efficiency of 11.2% from a 1.5 V supply. The reflection oscillator operated at 13.03 GHz with an output power of 7.1 dBm and efficiency 16.3% from a 1.5 V supply. The phase noise of

both oscillators was -118 dBc/Hz at an offset of 1 MHz. In both cases, the SIW resonator was coupled to the active device using microstrip lines. The measured loaded Q of the cavity used in the feedback oscillator was 61, whereas the measured loaded Q of the cavity used in the reflection oscillator was 95.3.

A push–push oscillator using a rectangular SIW cavity was reported in [68]. In this design, a rectangular SIW cavity with fundamental resonance at 14.75 GHz was used. Two reflection-type oscillators were designed based on the NEC NE3210S01 HJFET sharing the same cavity in a topology that is similar to the one illustrated in Fig. 7c. The oscillator outputs were combined using a Wilkinson power combiner. Harmonic balance simulation was used to predict the oscillation frequency and amplitude. The transmission lines coupling the active devices to the cavity and to the Wilkinson divider were adjusted in order to ensure that the two oscillators are synchronised in frequency and oscillate out-of-phase. As a result, the fundamental frequency components are cancelled at the output and the second harmonic components at 29.5 GHz are summed together. The measured output power of the oscillator was -14.7 dBm, compared to a simulated value of -9.6 dBm. The difference was attributed to component yield variations which led to a 6 dB variation between the second-harmonic component of the two individual fabricated oscillators. The simulated DC to RF conversion efficiency of the push–push oscillator was quite poor 0.61% and it is attributed to the low second-harmonic content of the designed oscillators. Optimised designs are able to obtain second-harmonic content comparable to the fundamental component, thus being able to demonstrate higher efficiency values. The oscillator had a phase noise of -105.7 dBc/Hz at 1 MHz offset.

Even though notable oscillator circuits have appeared in the literature, one can identify several research areas where oscillator design can take advantage of SIW technology: (i) Phase noise optimisation: Phase noise may be improved by designing resonator cavities with higher unloaded quality factors, utilising higher-order resonances and multiple substrate topologies. Non-linear analysis can be used to investigate the optimum loaded quality factor and coupling of the resonator to the oscillator circuit in order to minimise phase noise; (ii) High-frequency generation: The existing publications have demonstrated the capability of fabricating low-cost frequency sources in the millimetre wave range; however, there is still a large room for improvement utilising harmonic resonances and multiple device oscillators such as N-push topologies in order to demonstrate frequency generation in the frequency range of 100 GHz and above; (iii) Tunable oscillators: Tuning ranges of 2–3% have been demonstrated by placing a varactor diode inside the cavity. Work is still necessary in order to obtain tuning bandwidths of 10% or more, while maintaining a high quality factor, potentially using more than one varactor diodes or multiple resonator topologies; (iv) Coupled oscillator arrays: Coupled oscillators can be used in addition to high-frequency generation (such as in an N-push topology) in order to distribute a local oscillator (LO) signal with a desired phase distribution to a transmitting or receiving array. Preliminary work on coupled oscillator arrays in SIW technology has been demonstrated in [69] and is further described in Section 5.2.

4.2 Mixers

Frequency conversion circuits have been demonstrated by coupling active devices to hybrid passive circuits

implemented in SIW technology. An X-band single balanced diode mixer using a 90° hybrid was proposed in [70]. The mixer demonstrated a measured 6.8 dB insertion loss for an RF signal of 10 GHz and a LO signal of 8 dBm at 11.92 GHz. For a fixed LO frequency of 12 GHz and power of 8 dBm, the insertion loss of the mixer was better than 10 dB over the 9–12 GHz band. The various signals were coupled to the hybrid using SIW-to-microstrip transitions.

A second single-balanced mixer using the same topology was reported in [71], as part of a 24 GHz automotive radar system-on-package front-end. The mixer had a 6.7 dB conversion loss at an RF signal of 24.1 GHz, when a LO signal of 7 dBm was applied at 22.6 GHz. The mixer return loss was better than 10 dB within the 23–26 GHz band and the radio frequency–intermediate frequency (IF) isolation was better than 17 dB within the 24–24.25 GHz Industrial Scientific and Medical band of interest for the continuous wave radar application.

Additionally, SIW filters can be used to improve the performance of mixer circuits, although in this case the technology is not applied in the mixer circuit design itself. An example is a folded SIW filter with 3.3 dB insertion loss and a 3-dB bandwidth of 1.45 GHz centred at 36.225 GHz which was used as an image rejection filter of a monolithic microwave integrated circuit mixer operating at 34–40 GHz [72].

Self-oscillating mixers are compact circuits providing the functionality of both the oscillator and the mixer. They are designed by appropriately biasing and loading oscillator circuits in order to optimise conversion gain. The first reported self-oscillating mixer in SIW technology was based on a feedback oscillator [73]. The oscillator used a rectangular SIW cavity placed in the feedback path of a field-effect transistor (FET)-based amplifier circuit. The measured oscillation frequency was 14.347 GHz. The designed circuit was a sub-harmonic mixer, where the second harmonic of the oscillator was mixed with a 30.4 GHz RF signal to produce an IF at 1.7 GHz approximately, and had a measured conversion loss of 8.6 dB.

Similarly to the oscillator circuit developments, there are several areas where the application of SIW technology in mixer circuit design can lead to improved performance, such as (i) design of mixer topologies using compact hybrid implementations involving multiple substrate layers for low insertion loss and improved isolation, (ii) integration of active devices directly in the SIW structures, (iii) high-frequency implementations towards 100 GHz and above, where the losses associated with microstrip lines can be prohibitive and (iv) self-oscillating mixers demonstrating high-frequency performance and optimised conversion gain by employing non-linear simulation techniques. In addition, externally injection locked or coupled self-oscillating mixers can be used as phase-shifters in addition to frequency converters providing an immediate application in phased-arrays.

4.3 Amplifiers

The first reported amplifier in SIW technology was presented in [74]. The authors successfully demonstrated an X-band amplifier with 9 dB gain and less than 2 dB ripple over the entire X-band, while eliminating the use of SIW-to-microstrip transitions by coupling the active device directly to input and output SIW sections. Compact interconnects employing interdigitated series capacitors were used to couple the device terminals to the SIW sections. In addition

to DC blocking functionality, the printed series capacitors were used to provide input and output matching in combination with inductive irises consisting of plated through via holes inside the SIW.

SIW structures have also been proposed in designing bias networks for power amplifiers [75], in order to both suppress second and third-harmonic components from flowing in the bias line as well as support large DC currents. A typical bias network transmission line in microstrip technology requires a very thin high impedance line which may not be able to support very large DC currents. Using a SIW section in the bias network where the fundamental frequency and the second harmonic are both below the cut-off frequency of the fundamental propagating mode in the SIW ensures that they are effectively cancelled at the input and output bias networks. To additionally eliminate the third harmonic signal from flowing in the bias lines, shorted SIW sections with an appropriately selected length are used. It should be noted that using a SIW section in the bias line requires the use of separate substrate layer segments at the input and output of the amplifier in order to avoid shorting the DC source to the ground due to the metallised via holes of the SIW. The RF signal is then coupled to the input and output substrate sections using capacitors. The authors demonstrated the performance of their proposed bias networks by designing a 3.7–4.2 GHz power amplifier using Eudyna's C-band power GaAs FET FLM3742-4F. Their amplifier showed an output 1 dB compression point of 35.1 dBm and the suppression of the second and third harmonics were 65 and 58 dBc, respectively, showing an improvement of 22 and 13 dB over a test amplifier using traditional microstrip line-based bias networks.

The potential of SIW technology in relation to high-frequency signal amplification is further demonstrated by a Ka-band spatial power combiner amplifier reported in [76]. Power combining amplifier topologies consist of an input N-way power divider, followed by N-parallel amplifier circuits and an N-way power combiner. Utilising this architecture, the authors of [76] demonstrated an eight device power combining amplifier in SIW technology. At the input, a microstrip-to-SIW transition is used followed by a two-way SIW to HMSIW power divider. The signal into each of the two HMSIW branches is then split into four HMSIW-to-microstrip transitions effectively resulting in an eight-way power divider. Eight amplifiers are then connected to each of the microstrip interconnects. The output signals of the amplifiers are then combined using a second identical structure used as an eight-way combiner. The amplifier had a 19.5 dB small signal gain and a saturated output power of 30.6 dBm. The input and output return losses were better than 10 dB over the 33.5–35 GHz range. The measured power combining efficiency of the amplifier was 72%.

There exist several areas where further developments are expected in the application of SIW technology in amplifier design, including (i) minimisation of microstrip-to-SIW transitions in order to optimise insertion losses and increase efficiency, (ii) multi-device amplifiers and distributed amplifiers and (iii) reconfigurable amplifiers with tunable bandwidth and multi-band operation, to name a few.

5 SIW antennas

5.1 Passive antennas

In the last few years, there has been a growing interest in SIW-based antennas. Several configurations have been

proposed, starting from classical slotted-waveguides antennas. The first SIW antenna was based on a four-by-four slotted SIW array operating at 10 GHz [77]: this antenna is obtained by etching longitudinal slots in the top metal surface of an SIW. The feed network of this antenna is based on microstrip power dividers, integrated on the same substrate of the SIW antenna. A different topology is the leaky-wave antennas, introduced in [78]: this antenna exploits one of the fundamental characteristics of the SIW, namely, its property to generate radiation leakage when the longitudinal spacing of the metal vias is sufficiently large. Another leaky-wave SIW antenna, based on the TE₂₀ mode of the SIW structure, was proposed in [79] and provided better performance compared to the classical leaky-wave antenna based on the fundamental mode of the SIW.

Besides the classical waveguide-based antennas (with apertures either on the top wall or on the side wall), other antenna configurations have been proposed in the literature. A modified Vivaldi radiator was proposed in [80]: it consists of a dual V-type linearly tapered slot antenna, with centre frequency at 36 GHz. Also, this antenna topology appears particularly suited for integration in SIW technology.

Cavity-backed SIW antennas have been developed and tested [81–83]. The simplest structure was proposed in [81]: it consists of a slotted SIW cavity fed by a coplanar waveguide. The whole antenna (including the SIW cavity and the feed system) can be easily integrated on a single dielectric substrate. Another solution was proposed in [82]: it consists of an SIW cavity, slotted by a meander line and fed by a microstrip line. Besides the integration of the complete antenna and feed system on a single substrate, this solution offers improved compactness of the overall antenna. A Ku-band SIW cavity-backed antenna array was proposed in [83]: in this antenna, the radiating element is represented by a 2 × 2 array of metal patches, backed by SIW cavities. The cavities are realised by using metal vias, and the patches are fed using microstrip lines that are centrally fed by a coaxial probe. Cavity-backed SIW antennas result in planar antennas with relatively high efficiency (70% or more) and good front-to-back ratio (up to 20 dB).

Finally, an H-plane sectoral horn antenna in SIW technology was recently proposed in [84]: this antenna was also combined with a dielectric loading, integrated in the same substrate, which allows high gain and narrow beamwidths both in the E-plane and in the H-plane. This antenna topology has been used to form an array to obtain higher gain and to form a one-dimensional mono-pulse antenna array at 27 GHz.

5.2 Scanned and multiple-beam antennas

A significant effort has been devoted recently to the development of scanned and multi-beam SIW antennas. This requires, as a consequence, the design of beam-forming networks with increasing complexity. The aim of these beam-forming networks is to achieve a desired distribution of amplitude and phase coefficients at N output ports of an RF network, to properly excite the radiating elements and obtain the prescribed far-field pattern.

A simple way to realise a multi-beam antenna is based on the use of a parabolic reflector antenna: without moving the reflector mechanically, the beam produced by a parabolic reflector antenna can be steered over a limited angle by moving the feed off the reflector focal point. An SIW multi-beam antenna based on the parabolic reflector principle was proposed and implemented at 37.5 GHz [85]:

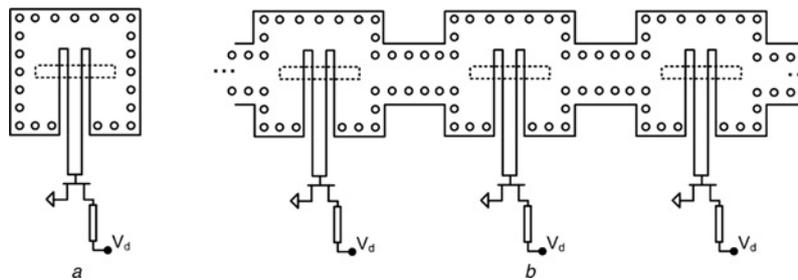


Fig. 8 Active integrated antennas in SIW technology

- a* Cavity-backed active oscillator antenna
b Cavity-backed coupled oscillator antenna array

in this antenna, seven input ports generate a corresponding number of output beams, and the parabolic reflector is implemented in planar form by an array of metallic vias.

The Rotman lens is another attractive solution to generate multi-beam antennas, because of its simple design and compact size. An SIW Rotman lens was adopted as a beam-forming network in [86]. A multi-beam antenna was implemented at 28.5 GHz with seven input ports and an antenna array constructed by nine SIW linear slot arrays, which can generate a corresponding number of beams along one dimension. Several antennas were also grouped in different ways to cover a 2-D solid angle with multiple beams.

A Blass matrix based on SIW technology was proposed in [87]. The antenna, operating at 16 GHz, consists of a double-layer structure, with a matrix of 4×16 cross-couplers. Compared with the conventional waveguide construction, this type of Blass matrix possesses the properties of lower cost, easier fabrication and low profile.

A multi-beam antenna based on the Butler matrix was implemented in [88, 89]: this solution is particularly interesting for SIW implementation, because it is based on a standard single-layer print circuit board process, which is more economical for mass production than are the advanced processes such as low-temperature co-fired ceramic and thick-film processes. In this antenna, the four components needed in the feeding network (90° hybrid coupler, cross-coupler, phase shifter and power splitter) are completely implemented in standard SIW technology.

Finally, the Nolen matrix was proposed in SIW technology. The Nolen matrix is a special case of the Blass matrix, where the termination loads are suppressed. A 4×4 Nolen matrix beam-forming network for multi-beam antenna applications, operating at 12.5 GHz, was proposed and tested in [90].

5.3 Active antennas

The term active antenna here refers to ‘circuit-antenna module’ as defined by Gupta and Hall [91]. Therefore it includes active integrated antennas where an active device is integrated in the same substrate with the radiating antenna structure [92], as well as an element in a quasi-optic array. In addition, it includes antenna elements where an active device is used to modify or reconfigure the properties of the antenna such as beam direction, polarisation or bandwidth.

Compact, single-substrate cavity-backed slot and patch oscillator antennas were proposed in [93, 94], respectively. A non-detailed schematic of the proposed circuit is shown in Fig. 8*a*. A square SIW cavity was used where the antenna was etched on the metal layer on one side of the substrate, and the antenna feed network and active device were placed on the other side, thus minimising unwanted

effects on the radiation pattern of the antenna. The cavity-backed antenna serves as the resonator structure of the oscillator. Furthermore, the ability to tune the oscillator frequency was demonstrated in [95] by removing one via hole from the cavity wall and introducing a varactor diode in its place. By tuning the varactor capacitance, it was possible to vary the resonance frequency of the oscillator by approximately 2%.

Coupled oscillator antenna arrays can be used in power combining as well as communication system applications [96]. Modulation can be easily introduced in the array through external injection, and beam-forming and beam-steering can be achieved by controlling the free-running frequencies of the array elements (the frequencies of the individual oscillator elements when they are uncoupled), thus eliminating the need for phase shifters or a complicated LO feed network. The authors of [68] proposed the use of coupled SIW cavities in the design of coupled oscillator arrays. Each array element is a cavity-backed active oscillator antenna, and coupling among the array elements is controlled by appropriately controlling the coupling between the individual SIW cavities through the use of a single or a double aperture. The proposed architecture is illustrated in Fig. 8*b*.

A cavity-backed antenna with reconfigurable circular polarisation has been proposed in [97]. A circular SIW cavity is considered and a symmetrical crossed slot is etched in the centre of the cavity wall, on the metal layer on one side of the substrate. Four shorting posts are additionally placed inside the cavity, near the slot edges and at points on the lines that extend from the centre of the cavity and along the slot arms. The posts are switched between a ground and an open state using diodes in a series topology. A circularly polarised wave is generated by altering the state of the posts corresponding to one arm of the crossed slot, effectively introducing a degree of asymmetry in the slot. By controlling the posts corresponding to a different arm of the slot, one can switch between orthogonal circular polarisation states and polarisation diversity is achieved.

Active antenna arrays, where antenna arrays implemented in SIW technology are placed in the same substrate comprised of a single or multiple layers where various transmit and receive modules are also integrated, have appeared in the literature [70, 98, 99], demonstrating the potential of the technology for complete SoS implementations.

6 Fabrication technologies for SIW components

The technological aspects are a key point for the implementation and development of SIW structures, especially for applications

in the millimetre-wave frequency range, at 60–90 GHz and at even higher frequencies. Conventional PCB techniques have been widely adopted to implement SIW structures, due to the reduced manufacturing cost and the great design flexibility. In this case, the metal holes are created either by micro-drilling or by laser cutting, and their metallisation is performed by using a conductive paste or metal plating [9]. The PCB technique exhibits an additional advantage, because it allows the integration of the complete system (including microstrip or coplanar circuitry, as well as the active elements) on the same substrate with the same fabrication technique [12].

At higher frequencies radiation issues can arise, due to some technological limitations: in fact, fabrication constraints prevent the longitudinal spacing between metal vias going below a certain value. A possible solution to this problem was proposed in [26], where the via holes are replaced by metallised slots in a circuit operating at 94 GHz.

LTCC technology has also been used in SIW implementation. The availability of several layers and the tiny dimension of the via holes permits implementation of extremely compact SIW components. SIW filters in LTCC technology were presented in [100, 101]: in these filters, the SIW resonators are vertically stacked, so that the filter size can be miniaturised.

SIW components operating above 100 GHz were fabricated using photoimageable thick-film materials, with excellent dimensional tolerances and low dielectric loss in [25]. The fabrication process is the following: first, a uniform metal layer is printed on an alumina substrate, to form the bottom wall of the waveguide. Then, a 10 μm dielectric layer is printed and photoimaged, forming the waveguide sidewalls. This step is repeated to achieve the required thickness of the SIW structure. Finally, a conductor layer is printed and photoimaged to form the upper wall of the waveguides. After printing and imaging, each layer is dried and fired prior to the processing of subsequent layers. SIW filters operating at 180 GHz, fabricated by using this technology, were designed and experimentally verified in [25].

The implementation of SIW antennas in a flexible substrate was proposed in [102]. The fabrication is based on a technique called 'ion track technology' and can be summarised as follows: first, Kapton foils are pretreated by irradiation with heavy ions and subsequent wet etching. The ions determine tracks in the material, and these tracks are selectively etched during the wet etch step, to form pores. Finally, the surface of the foil is irradiated and etched, and the pores perforate the whole thickness of the foil. An SIW-based slot array antenna in a flexible PCB was successfully demonstrated in [102].

7 Conclusion

This paper has presented the state-of-the-art of SIW technology: the recent advances in the field of passive components, active components and antennas implemented in SIW technology have been described and discussed. Issues related to the modelling and design of SIW components have been addressed, and the different technological solutions proposed for the implementation of SIW components and circuits have been presented.

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